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BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

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The Nature Of Q

W. CULLEN MOORE *Engineering Manager*

A discussion of the physical concepts underlying a familiar and useful, but not always fully appreciated, quantity -- "Quality Factor."

The familiar symbol, Q, has something in common with a certain famous 19th century elephant of Indostan. You may recall that in the poem six blind men each investigated the same elephant with the agreement that they would report their findings to each other and thereby determine the true nature of an elephant. One chanced to touch the side of the elephant and reported "God bless me! But the elephant is very like a wall." Another, touching the tail, proclaimed an elephant was like a rope. The third, chancing upon a leg, avowed the elephant to be kind of a tree, and so on. The confusion of reports prompted the poet to observe in conclusion that, "Each was partly in the right, and all were in the wrong."

And so it is with Q. The concept of Q which each engineer favors is the one based on the way in which he uses Q most frequently. It might be to describe selectivity curves, or the resonant rise in voltage, or the impedance of a parallel resonant circuit, or the envelope of a damped wave train. If one were to ask for a definition of Q, the most common response probably would be "Q equals $\omega L/R_x$". But like the description of the elephant, this too is partly right and partly wrong. The reason is, that while one can obtain a numerical value for Q by dividing the quantity (ωL) by R, it tells little or nothing about the real nature of Q.

The expression $\omega L/R_x$ is a dimensionless ratio and therefore a pure number. As such it enjoys no distinction from other pure numbers. If we are to look for the *meaning* of

YOU WILL FIND. . .

Figure 1. The importance of the quantity Q in the analysis of electronic circuits and components has made the Q Meter a familiar laboratory tool. Here, H. J. Lang, BRC Sales Engineer, is checking the accuracy of a Q Meter Type 260-A with the new Q-Standard.

Q as a basis for its description, we must look for a physical concept. We may then explore the implications and applications of this concept in a variety of specific situations.

Let us go one step further in our analysis of the expression $\omega L/R_x$. It is not immediately apparent why this *particular* numerical ratio should be chosen to describe certain characteristics of components and circuits over all the other similar ratios which might be set up. The reason for this choice once again refers back to the concept involved in the establishment of a definition for Q. We shall see presently that the basic idea leads directly to a simple expression by which we can determine a numerical magnitude.

In the first place, the Q of a circuit or component has practical significance only when an alternating current, usually sinusoidal in waveform, is flowing through it. The circuit parameters associated with alternating currents, namely capacitance and inductance, have the common characteristic of being capable of storing energy. An inductor stores energy in the form of an electromagnetic field surrounding its winding. A capacitor stores energy in the form of polarization of the dielectric. Each of these systems will deliver most of the stored energy back into the circuit from which it came. These common characteristics indicate that perhaps we should look to energy relationships for an appropriate description of the behavior of circuits.

As mentioned above, most, but not all of the energy stored in an inductor or a capacitor is delivered back into the total system. If we start with this energy concept, we are in a position to derive a *figure* of *merit* for the system in terms of its ability to store energy as compared with the energy it wastes.

Continued on Page 2

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The Nature of Q (continued)

DERIVATION OF $Q = \omega L/R$

In describing the behavior of a circuit in which an alternating current is flowing (as shown in Fig. 2), it is most convenient to use as our interval one complete current cycle. During this interval the system will have experienced all of its configurations of energy distribution and will have returned as nearly as possible to the starting condition. We are interested in the ratio of the total energy stored in the system to the amount of energy dissipated per cycle by the system.

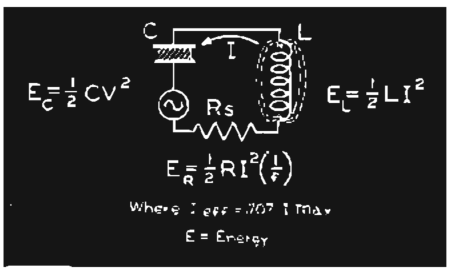

Figure 2. Energy relationships in an elementary a-c series circuit.

To calculate the total stored energy, let us select that portion of the cycle at which all the energy is stored in the field of the inductor. (This is quite arbitrary, as we could just as well assume all the energy to be stored in the capacitor.) We recall from electrical engineering that the energy stored in the field surrounding an inductor is equal to $1/2\ LI^2$. In this case I will be the peak current in amperes.

The *average power* lost in the resistor is $1/2R_sI^2$, where R_s is the total series resistance of all elements in the circuit, and I is the peak current in amperes. The factor 1/2 appears because the (*effective* current) $= .707$ (*peak* current, I), and $(.707I)^2 = 1/2I^2$.

The energy lost per cycle is equal to the average power times the time of one cycle, $T = (1/f)$, or $1/2R_sI^2T$.

The ratio of stored energy to energy dissipated per cycle becomes:

$$\frac{\frac{1}{2} L I^2}{\frac{1}{2} I^2 R_s T} = \frac{1}{T}\frac{L}{R_s} = \frac{fL}{R_s} = \frac{1}{2\pi}\frac{2\pi f L}{R_s}$$

$$= \frac{1}{2\pi}\frac{\omega L}{R_s} = \frac{1}{2\pi} Q$$

Hence: $Q = 2\pi \dfrac{\text{total energy stored}}{\text{energy dissipated per cycle}}$

Thus we see that the familiar expression giving the magnitude of the quantity Q follows directly from the basic concept of the ability of a component or circuit to store energy and the energy dissipated per cycle.

Q IN A PARALLEL CIRCUIT

The above analysis has been made on the assumption of a so-called series circuit which assumes all losses in the circuit to be represented by a single resistor in series with a lossless inductor and a lossless capacitor. We are now interested in obtaining an expression for the case in which we are looking at the circuit from the outside, or parallel connection, in which the resistor, the inductor, and the capacitor are all in parallel as shown in Fig. 3.

An equivalent expression for Q for the two circuits of Fig. 3 can be obtained most readily if we consider the current distributions when the applied alternating current has the same frequency as the resonant frequency of the R-L-C combinations. In Fig. 3-a, the current, I, flowing through the circuit from point A to point B is controlled by the parallel resonant impedance of the circuit:

$$Z_{AB} = \frac{(-j\frac{1}{\omega C})(j\omega L + R_s)}{(-j\frac{1}{\omega C}) + (j\omega L + R_s)}$$

At resonance: $\left|\frac{1}{\omega C}\right| = |\omega L| = X$,

where $|\ \ |$ indicates magnitude, so that

$$Z_{AB} = \frac{(-jX)\ (+jX + R_s)}{-jX\ + jX + R_s} = \frac{X^2 - jXR_s}{R_s}$$

$$= \frac{X^2}{R_s} + (-jX).$$

The absolute magnitude of this impedance is

$$Z_{AB} = \sqrt{\left(\frac{X^2}{R_s}\right)^2 + X^2} = X\sqrt{\frac{X^2}{R_s^2} + 1}.$$

Or, $\quad Z = \omega L\ \sqrt{Q^2 + 1}$

For most practical purposes this reduces to:

$$Z = Q\omega L,$$

which is the impedance of a parallel reasonant circuit. For the external current flowing through Figure 3-a we may then write, $I = E/Q\omega L$.

Referring to Figure 3-b, we may consider that the combination of C and L, with all losses now accumulated into the equivalent parallel resistor R_p, forms at resonance an infinite impedance circuit in shunt with a finite resistor R_p. The current flowing through such a circuit will be $I = E/R_p$.

Equating: $\dfrac{E}{Q\omega L} = \dfrac{E}{R_p}$; or, $Q\omega L = R_p$

Rewriting: $Q = R_p/\omega L$.

where R_p = total effective parallel circuit resistance in ohms.

It is convenient to remember that for the series case, R_s is in the denominator and Q becomes very large as the dissipative component R_s becomes small. In the case of the parallel resonance circuit, the larger the shunt resistance the larger the value of Q.

Summarizing:

$$Q = \frac{\omega L}{R_s} = \frac{1}{\omega C R_s} = \frac{R_p}{\omega L} = \omega C R_p$$

SELECTIVITY

We have seen how the expression $Q = \omega L/R_s$ can be derived directly from power consideration in an R-L-C circuit. By extending the analysis of power relationship in such circuits we can also derive an expression

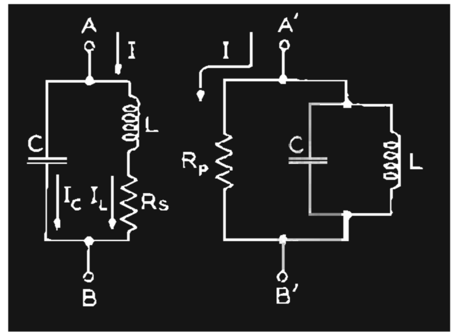

Figure 3. Current distributions in parallel resonant circuits.

which describes the selectivity, or response-versus-frequency, curve for circuits in the vicinity of their natural resonant frequency.

To begin with, we will need to establish two points on the resonance curve for reference. A convenient choice of points is one in which the net circuit inductive or capacitive reactance equals the resistance in the circuit. These two points can be shown to lie at frequencies at which the power in the circuit is one half the power at the maximum response frequency. (See Fig. 4.)

Assume that the reactance equals the resistance. Then the total circuit impedance is equal to the following:

$$Z = \sqrt{R_s^2 + X^2} = \sqrt{R_s^2 + R_s^2}$$

$$= \sqrt{2R_s^2} = 1.414R_s$$

We must remember that this new impedance consists of the original resistance plus some reactance. Only the resistance component of the impedance consumes power. If we apply the same voltage to this circuit at the selected frequency as at the resonant frequency, the current at the new selected frequency will be $I_f = 0.707\ I_0$, where I_0 is the current at resonance. The power dissipated in the circuit is then

$$W_f = I_f^2R_s = (.707 I_0)^2R_s = .5\ I_0^2R_s$$

$$= .5W_0$$

Let us now see what frequency relationships are involved. Near resonance, if we change the frequency by a small amount Δf toward a higher frequency, the net reactance of the circuit will change due to two equal contributions in the same direction: (1) there will be a small increase in the inductive reactance due to the increased frequency, and (2) there will be an equal amount of decrease in the capacitive reactance. The net change in reactance is the sum of these two equal changes. The change in reactance due to the increased inductive reactance alone is $\Delta X_1 = 2\pi \Delta f L$, and the change in the total reactance is

$$\Delta X = 2(2\pi \Delta f L) = 4\pi \Delta f L.$$

Choose Δf equal to the difference between the frequency at either of the half-power points, f_1 or f_2, and the resonance frequency, f_0. Since we have seen that at the half-power points $X = R$, we can write the two following equations:

$$R_s = 4\pi (f_0 - f_1) L$$
$$= 4\pi f_0 L - 4\pi f_1 L$$

$$R_s = 4\pi (f_2 - f_0) L$$
$$= -4\pi f_0 L + 4\pi f_2 L.$$

Adding these two equations:

$2R = 4\pi (f_2 - f_1) L.$

Re-arranging and multiplying both sides by f_0

$$\frac{f_0}{(f_2 - f_1)} = \frac{2\pi f_0 L}{R_s} = \frac{\omega L}{R_s} = Q$$

This is the application of Q which is most familiar to radio engineers; namely, an expression of the selectivity of a resonant circuit in terms of Q. As we see above, it is based on the *power* dissipated in the circuit at two selected frequencies.

RESONANT RISE IN VOLTAGE

Let us now look at another common manifestation of the Q of a resonant circuit; namely the voltage multiplication phenomena.

Consider once again the series circuit of Fig. 2 having a total equivalent series resistance, Rs, and a circulating current caused by a small sinusoidal voltage, e, injected in series with the circuit. At series resonance

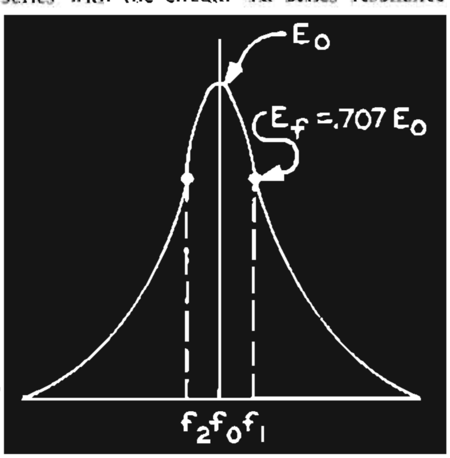

Figure 4. Resonance curve, showing half-power points.

the current circulating within the resonant circuit is limited only by the resistance and will be $I_0 = e/R_s$. This circulating current will produce a voltage across the inductor equal to $E = I_0\, \omega L = (e/R_s)\, \omega L$.

The resonant rise in voltage then is

$$\frac{E}{e} = \frac{\omega L}{R_s} = Q$$

This is often written

$E = Qe.$

For relatively high values of R_s (corresponding to low Q) we must also account for the drop across the resistor:

$$E = I_0 \sqrt{R_s^2 + \omega^2 L^2}$$

$$= \frac{e}{R_s} \sqrt{R_s^2 + \omega^2 L^2} = e \sqrt{1 + \frac{\omega^2 L^2}{R_s^2}}$$

So for this case

$$E = e\sqrt{1 + Q^2}$$

Of course we could just as well have analyzed this circuit from the standpoint of the voltage across the capacitor, but we would have arrived at exactly the same results.

POWER DISSIPATION

Proceeding directly out of the method by which we derive Q, namely from the standpoint of energy, we can see that the net Q of the complete oscillator circuit describes the manner in which the circuit causes the current to flow in alternate directions, and describes the energy lost per cycle in the process. This lost energy per cycle must be made up by the power supply of the system or oscillation will die out.

We know that a circuit consisting of an inductor, a capacitor and a resistor in series, which is charged and allowed to oscillate, will experience an exponential decay in the magnitude of the peak current. This decay follows the form $\epsilon^{\left(-\frac{R}{2L} T\right)}$. The portion of this expression R/2L is defined as the *damping coefficient*, and describes the amount by which each successive cycle is lower than its predecessor, as shown in Fig. 5. If we multiply the damping coefficient by the time for one cycle, we obtain the expression known as the *logarithmic decrement* of a circuit, which includes the effect of frequency. In each successive cycle of period T we obtain the following current ratios:

$$\frac{I_2}{I_1} = \epsilon^{-\frac{R}{2L} T} = \epsilon^{-\delta}$$

$$\text{But } T = \frac{1}{f}, \text{ so } \delta = \frac{R_s}{2fL}, \text{ or } \delta = \frac{\pi}{Q}$$

$$\text{Rewriting: } Q = \frac{\pi}{\delta}$$

We see that in this application Q is intimately linked with the rate of decay of oscillation in a dissipative circuit. Before we leave the subject of Q and power, let us mention briefly two other factors which find common usage in electrical engineering. The first of these is the *phase angle* between the

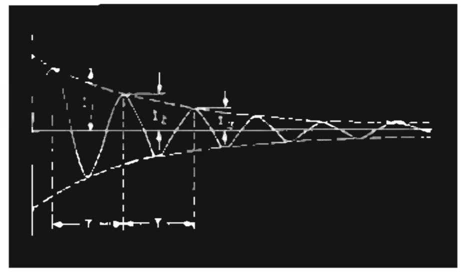

Figure 5. Q as a damping factor.

current and the driving voltage in a circuit containing reactance and resistance. If we once again arbitrarily limit ourselves to consideration of inductors, the expression for phase angle is the familiar formula:

$$\tan \phi = \omega L / R_s = Q$$

Or: Q = (tangent of the phase angle.)

Closely associated with the phase angle is the *power factor*. The power factor of an inductor is the ratio of the total *resistance* absorbing power to the total *impedance* of the device, and is designated by $\cos \phi$:

$$\cos \phi = \frac{R_s}{\sqrt{R_s^2 + \omega^2 L^2}} = \frac{R_s}{R_s \sqrt{1 + \frac{\omega^2 L^2}{R_s^2}}}$$

$$= \frac{1}{\sqrt{1 + Q^2}}$$

This is approximately $\cos \phi = \frac{1}{Q}$.

THE Q METER

Practically all of the relationships mentioned above have been used in radio and electrical engineering for a great many years. However, the expression Q and its numerical value of $Q = \omega L / R_s$ did not come into popular usage until the early 1930's. The need for the rapid measurement of Q arose with the growth of the broadcast receiver industry, and Boonton Radio Corporation demonstrated the first "*Q-METER*" at the Rochester IRE Meeting in November, 1934.

A numerical quantity for Q might be obtained by measuring each of the parameters involved in any of the various forms which have been given above. However, certain of these expressions lend themselves to direct measurement much more readily than others. Originally, the favored method was to actually measure ωL and R_s. Later, measurements of Q were based on the frequency relationship, using a heterodyne detector system. This method is feasible but demands great accuracy of the variable frequency generator in order to obtain resasonable accuracy of the final result.

An expression equivalent to the frequency relationship can be written in terms of capacitance. For the series resonant case we obtain the following:

$$Q = \frac{2C_0}{C_2 - C_1}$$

The multiplier 2 is introduced because the change in frequency is proportional to the square root of the change in capacitance. For incremental quantities this reduces to 2.

The relationship which has found almost universal acceptance in the design of instruments for the direct measurement of Q makes use of the resonant rise of voltage principle outlined above. In such instruments, a small radio frequency voltage of known magnitude is injected into the resonating circuit across a very small series resistor. At resonance this voltage causes a current to flow which is limited only by the magnitude of the total equivalent series resistance of the circuit. The current flowing through the inductor results in the resonant rise of voltage given by $E = eQ$. This magnified voltage is read by a vacuum tube voltmeter connected across the resonating capacitor. Since the series voltage injected into the circuit is known, it is possible to calibrate the scale of the voltmeter directly in values of Q.

CONCLUSION

We have seen that the conventional expression for the magnitude of Q can be derived from the basic concept of energy stored compared to energy dissipated per cycle in a resonant system. Its use as a measure of the damping effect in decaying wave trains, its relationship to phase angle and power factor, and the selectivity of a resonant circuit are seen to come out of energy and power considerations. In addition to these factors, such critical basic measurements as radio frequency resistance of a wide variety of components, the loss angle of capacitors, dielectric constants, characteristics of antennas, and transmission line parameters are all part of the continually expanding list made practical by a simple, direct-reading instrument for the measurement of Q, the *Q-Meter*.

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Vacuum Tube Oscillators, 1st Edition, 1953, William A. Edson; John Wiley and Son; Pages 20-21.

Vacuum Tube Circuits; 1st Edition, 1948, Lawrence B. Arguimbau, John Wiley and son; Pages 184-185.

THE AUTHOR

W. Cullen Moore was graduated from Reed College in 1936 with a B.A. degree in

physics. He studied advanced Electrical Engineering, specializing in microwaves and UHF, at Northwestern University between 1939 and 1942, and received an M.A. in physics from Boston University in 1949. From 1940 to 1947 he was Senior Project Engineer for Motorola, Inc., where he directed work on FM receiver design and signal generating equipment. During the war, he had charge of the development of the SCR-511 "Cavalry Set", the redesign of the SCR-536 "Handie-Talkie", and airborne communications equipment.

Between 1947 and 1951, Mr. Moore was a Project Supervisor at the Upper Air Research Laboratory at Boston University, where he supervised the design of rocket-borne electronic equipment. During the same period he taught electronics as an instructor in the B. U. Physics Department. In 1951 he joined Tracerlab, Inc., where he remained as Chief Engineer until 1953, when he accepted the position of Engineering Manager of Boonton Radio Corporation.

Basic Formulas Involving Q

A. TWO-TERMINAL IMPEDANCE

FORMULAS RELATING EQUIVALENT SERIES AND PARALLEL COMPONENTS

$$Q = \frac{X_s}{R_s} = \frac{\omega L_s}{R_s} = \frac{1}{\omega C_s R_s} = \frac{R_p}{X_p} = \frac{R_p}{\omega L_p} = R_p \omega C_p$$

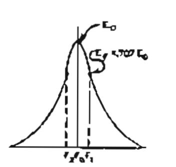
General Formula	Q greater than 10	Q less than 0.1	General Formula	Q greater than 10	Q less than 0.1
$R_s = \frac{R_p}{1+Q^2}$	$R_s = \frac{R_p}{Q^2}$	$R_s = R_p$	$R_p = R_s(1+Q^2)$	$R_p = R_s Q^2$	$R_p = R_s$
$X_s = X_p \frac{Q^2}{1+Q^2}$	$X_s = X_p$	$X_s = X_p Q^2$	$X_p = X_s \frac{1+Q^2}{Q^2}$	$X_p = X_s$	$X_p = \frac{X_s}{Q^2}$
$L_s = L_p \frac{Q^2}{1+Q^2}$	$L_s = L_p$	$L_s = L_p Q^2$	$L_p = L_s \frac{1+Q^2}{Q^2}$	$L_p = L_s$	$L_p = \frac{L_s}{Q^2}$
$C_s = C_p \frac{1+Q^2}{Q^2}$	$C_s = C_p$	$C_s = \frac{C_p}{Q^2}$	$C_p = C_s \frac{Q^2}{1+Q^2}$	$C_p = C_s$	$C_p = C_s Q^2$

B. TUNED CIRCUIT

1. Selectivity

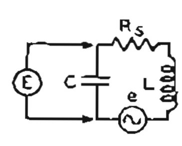
$$Q = \frac{f_0}{f_1 - f_2} = \frac{2C_0}{C_2 - C_1}$$

Where f_1 and f_2 are half-power points and C_0, C_1, and C_2 are capacitance values at f_0, f_2 and f_1 respectively.



2. Resonant Rise in Voltage $Q = \frac{E}{e}$

For relatively large R_s (low Q), $E = e\sqrt{1+Q^2}$

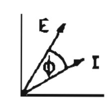


3. Power Dissipation

a. Power Factor = cos ϕ

$$= \frac{R}{\sqrt{R^2 + L^2\omega^2}} = \frac{1}{\sqrt{1+Q^2}}$$

and for inductors, $Q = \tan = \frac{\omega L}{\phi}$



b. Damped Oscillations $Q = \frac{\pi}{\delta}$, where δ is the logarithmic decrement.

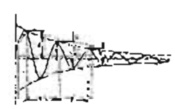


Figure 1. The Q-Standard Type 513-A

The Q-Standard

A NEW REFERENCE INDUCTOR FOR CHECKING Q METER PERFORMANCE

Dr. Chi Lung Kang
James E. Wachter

Widespread acceptance of the Q Meter as a basic tool for electronic research and development has lead, in recent years, to an increasing demand for some convenient means of checking the performance and accuracy of the instrument periodically in the field.

As a result of this demand, BRC engineers have developed the recently-announced Q-Standard Type 513-A, a highly stable reference inductor, intended specifically for use in checking the performance of Q Meters Type 160-A and 260-A. By comparing the accurately-known parameters of this inductor directly with the corresponding values read on the Q Meter, the user may now obtain a dependable indication of the accuracy with which his Q Meter is operating.

The Q-Standard is designed and constructed to maintain, as nearly as possible, constant electrical characteristics. In external appearance the unit is very similar to the inductors (Type 103-A) which are available for use as accessory coils in a variety of Q Meter measurements. This resemblance is only superficial, however, since highly specialized design and manufacturing techniques have been required to provide the high degree of electrical stability demanded of such a unit.

The inductance element consists of a high-Q coil of Litz wire wound on a low-loss steatite form. After winding, the coil is heated to remove any moisture present, coated with silicone varnish, and baked. A stable, carbon-film resistor is shunted across the coil to obtain the proper Q-versus-frequency characteristics. The coil form is mounted on a copper base which in turn is fitted to a cylindrical, copper shield can. The coil leads are brought through the base to replaceable banana plug connectors which allow the unit to be plugged directly into the Q Meter COIL posts. The low potential connector is mounted directly on the base, while the high potential connector is insulated from the base by a steatite bushing. To provide maximum protection against moisture, the unit is hermetically sealed, evacuated, and filled with dry helium under pressure.

ELECTRICAL CHARACTERISTICS

The principal electrical characteristics of each individual Q-Standard are measured at the factory and stamped on the nameplate of the unit. These include the inductance (L), the distributed capacity (C_d), and 3 values of effective Q (Q_e) and indicated Q (Q_i), determined at frequencies of 0.5, 1.0 and 1.5 mc, respectively.

The effective Q may be defined as the Q of the Q-Standard assembly mounted on the Q-Meter, exclusive of any losses occurring in the measuring circuit of the Q Meter itself. It differs from the true Q by an amount which depends largely on the distributed capacitance of the inductor. At the frequencies for which Q_e is given, the following relation is approximately correct:

$$\text{TRUE } Q = Q_e\,(1 + C_d'/C')$$

Where C' and C'_d are corrected values of resonating capacitance and distributed capacitance, respectively, as described below.

The Q of the unit as read on an average Q Meter (indicated Q) will differ from the effective Q by a small percentage which is the result of certain losses inherent in the measuring circuit of the instrument. These losses are minimized, and may usually be disregarded in all but exacting measurements. However, to provide a more accurate check on the Q Meter reading, The Q-Standard is also marked with values of indicated Q. Small variations in the calibration of both the Q Meter and the Q Standard may cause individual instruments to deviate slightly from the expected reading, but a Q Meter Type 160-A or 260-A which indicates within $\pm 7\%$ of the Q_i value marked on the Q-Standard may be considered to be operating within its specified tolerances. Although quantitative indications are not possible, it is worthwhile to note, when wider deviations are encountered, that an error which is greatest at 0.5 mc may indicate calibration inaccuracy, while one which becomes severe at 1.5 mc may be caused by excessive shunt loading effects.

In addition to checking indicated Q, the Q-Standard may be used to determine the calibration accuracy of the Q Meter resonating capacitor. This may be done readily by tuning the measuring circuit to resonance at any desired frequency within the resonant limits of the Q-Standard, and comparing the reading on the capacitor dials with the value predicted by the expression,

$$C = \frac{1}{\omega^2 L} - C_d$$

The measuring circuit of a Q Meter Type 160-A or 260-A, with a Q-Standard mounted on the COIL posts, is represented in Fig. 2-a. Here R_{sh} is the Q Meter shunt loss, Q is the

Q-indicating meter, R_i is the Q Meter injection resistor, and C' is the resonating capacitance. L, R and C_d' represent the inductance, series resistance and corrected distributed capacitance, respectively, of the Q-Standard. The equivalent circuit shown in Fig. 2-b indicates the corresponding *effective* parameters of the Q-Standard, which are related to the values in Fig. 2-a as follows:

$$L_e = \frac{L}{1-\omega\ L C_d'}$$

$$R_e = \frac{R}{(1-\omega^2 L C_d')^2}$$

$$Q_e = \frac{\omega\ L e}{R_e}$$

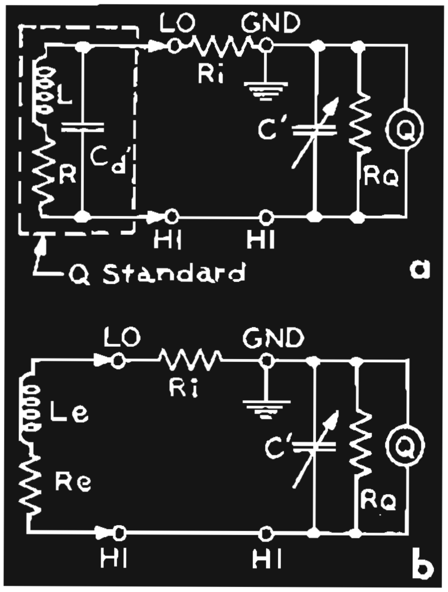


Figure 2. Schematic representation of Q Meter measuring circuit with Q-Standard attached.

It is worthwhile to consider, briefly, the corrected value of distributed capacitance (C_d') mentioned above. This value is the distributed capacitance of the Q-Standard when it is actually mounted on the Q Meter. It differs by a small, constant value from the distributed capacitance (C_d) marked on the nameplate, because of a capacitance shift caused by the proximity of the Q-Standard shield can to the Q Meter HI post. This proximity causes the transfer of a small value of capacitance from between the Q Meter HI post and ground to between the HI post and the Q-Standard shield can. This results in a change in the calibration of the resonating capacitor, and a corresponding change in the Q-Standard distributed capacity.

Thus, if the tuning dial of the resonating capacitor is adjusted to a value, C, with nothing attached to the coil posts, the actual value of tuning capacitance will be reduced by a small constant to a new value, C', when the Q-Standard is connected. At the same time, the distributed capacitance of the Q-Standard is increased to become Cd'. The magnitude of this effect is 0.4 $\mu\mu f$, and we may write,

$$C' = C - 0.4\ \mu\mu f$$

$$C_d' = C_d + 0.4\ \mu\mu f$$

When the Q-Standard is used to check the calibration of the resonating capacitor, in the manner described above the value, Cd, indicated on the nameplate is used. In other applications, however, where accurate results are desired, the corrected values, C' and Cd', must be used. In determining Q_e for example,

$$Q_e = \frac{\omega L e}{R_e} = \frac{1}{R_e\ \omega C'},$$

it can be seen that the correction may assume some importance, particularly at 1.5 mc, where C' is relatively small.

It should be noted that, in order to hold this proximity effect constant, particular care has been taken to provide for accurately-reproducible positioning of the Q-Standard with respect to the Q Meter HI post. For this purpose, the base of the high-potential connector serves as a mounting stop. When this connector is fully inserted in the HI post, the low potential connector (which is the shorter of the two) will not be fully seated in the LO post, and the insulated support attached to the Q-Standard base will not touch the top of the Q Meter cabinet.

If desired, a secondary standard inductor may be derived from the Q-Standard by means of a comparison method which is both simple and accurate. The accuracy of the Q Meter, which is the only equipment needed, has only higher order effects on the results.

The inductor selected should have electrical parameters and outside shield dimensions which are fairly close to those of the Q-Standard. The standardization (i.e. accurate determination of the effective Q of the secondary standard) is done as follows: First, plug the Q-Standard into the Q Meter and resonate the measuring circuit at one of the three frequencies (0.5, 1.0 or 1.5 mc) for which Q_e is given on the Q-Standard nameplate. Then replace the Q Standard with the secondary standard and obtain readings of ΔQ (from the ΔQ scale) and ΔC ($C_1 - C_2$). With the data given on the Q-Standard nameplate, determine C' from,

$$C' = \frac{1}{2L} - (C_d + 0.4\ \mu\mu f)$$

The effective Q of the secondary standard may then be determined from the relation,

$$Q_e(\text{unkown}) = \frac{\omega(C'+\Delta C)}{\frac{\omega C'}{Q_e} + \frac{\omega C'}{(Q+\Delta Q)}\left[(1+\frac{\Delta C}{C'})(1-\frac{\Delta Q}{Q+\Delta Q})-1\right]}$$

where C_d, L and Q_e are given on the Q-Standard nameplate.

Service Note

REPLACING THE THERMOCOUPLE ASSEMBLY TYPE 565-A IN THE Q METER TYPE 260-A

It is the function of the Q Meter thermocouple to monitor accurately the voltage injected by the oscillator into the measuring circuit. Although the unit in the Q Meter Type 260-A has been made considerably more rugged than that of the older Q Meter Type 160-A, it is necessarily a sensitive device which may be subject to damage or burnout under prolonged overload. For this reason, care is necessary in operating the instrument to avoid increasing the oscillator output (indicated on the XQ Meter) into the "red-lined" region beyond the indicated XI value.

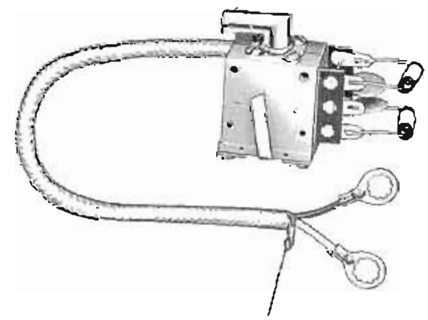


Figure 1. Thermocouple Assembly Type 565-A

If thermocouple failure should occur, the assembly may be replaced, by the user, with a new assembly obtained from the factory, if the proper care is taken. In ordering, it is necessary to include the serial number of the Q Meter in which the thermocouple is to be used since they must be individually matched. The procedure outlined below is presented as reference material for the convenience of Q Meter Type 260-A owners.

CHECKING FOR THERMOCOUPLE FAILURE

If no reading can be obtained on the XQ meter, thermocouple burnout may be sus-

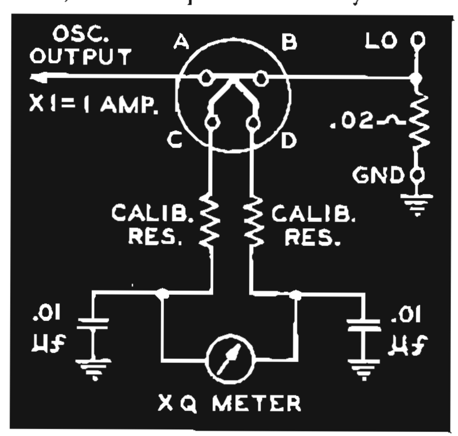


Figure 2. Thermocouple circuit of the Q Meter Type 260-A.

pected. Since this symptom may also be produced by failure of the local oscillator, however, the output of the latter should be checked first. This may be done by measuring from point A (See Fig. 2) to ground with a vacuum tube voltmeter. If the oscillator produces a voltage across these points, disconnect one lead from the XQ meter and check for continuity between points A-C, A-B, B-D and C-D. An open circuit between any of these indicates thermocouple failure. The maximum resistance of the XQ meter is 65 ohms; the total resistance of the junction circuit loop, including the XQ meter, calibration resistors and thermocouple element, can vary from 85 to 115 ohms. CAUTION: Do not disassemble the thermocouple unit.

REPLACEMENT PROCEDURE

The 565-A thermocouple replacement assembly for the Q Meter Type 260-A includes the thermocouple unit itself, a 0.02 ohm insertion resistor, two calibration resistors and two filter capacitors. Replacement of the assembly should be made as follows:

1. Remove the front panel and chassis assembly from the Q Meter cabinet and place it, face down, on a flat work surface.
2. Remove the UG-88/U plug from the receptacle at the rear of the thermocouple assembly.
3. Unscrew and remove the LO binding post terminal nut. Then, using a right-angle soldering iron (see Fig. 3), carefully unsolder the thin metal strap which connects the thermocouple unit to the bottom of the LO post.
4. Remove the terminal lugs from the XQ meter and unclamp the cable from the front panel and resonating capacitor frame.
5. Remove the four mounting screws from the thermocouple assembly, and carefully remove the assembly from the Q Meter.
6. Install the new unit and connect the attached cable to the XQ meter terminals, observing the indicated polarity. Clamp the cable to the front panel and resonating capacitor frame.

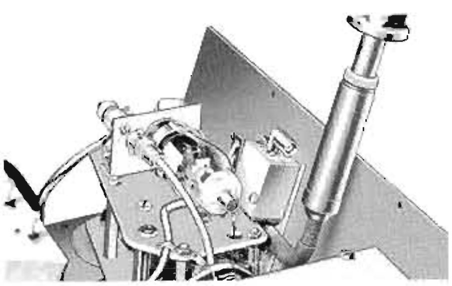
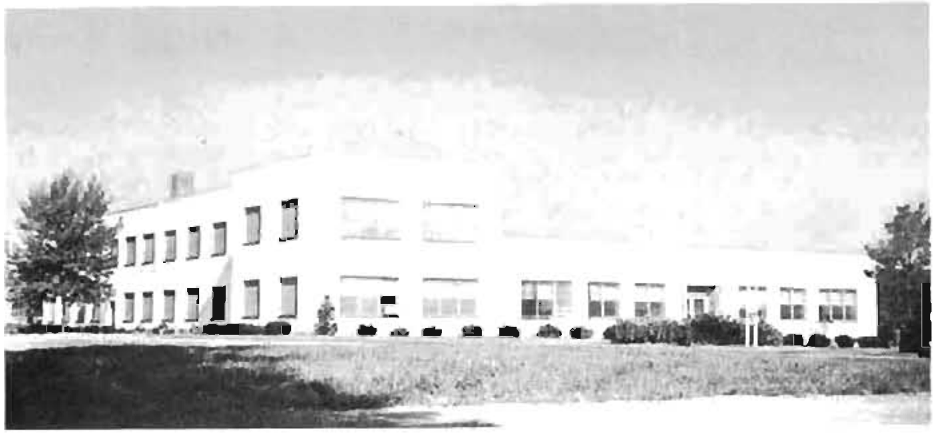


Figure 3. Using a right-angle soldering iron to solder the thermocouple connecting strap in place

7. Trim the connecting strap to a length which will permit it to reach the bottom of the LO post with a small amount of slack to allow for binding post movement. Solder this strap to the LO post, being careful not to leave the iron in contact with the strap any longer than necessary.
8. Replace the binding post nut and return the instrument to its cabinet.— **E. GRIMM**



an introduction to
BOONTON RADIO CORPORATION

Frank G. Marble, *Sales Manager*

The Boonton Radio Corporation was formed in 1934. Since that time it has been developing, designing, and manufacturing precision electronic instruments. To understand some of the details of the Company's growth, we must take a look at the field of electronics for a few years preceding 1934.

Many of the concepts that made wireless communication possible were discovered before the First World War. During this war many new ideas evolved and considerable practical experience was gained in the use of the new ideas. A keen public appreciation of the usefulness of the transmission of intelligence over a distance without wire connection appeared at this period. In the years following the war, manufacturers began devoting time and money to the use of radio devices for many purposes. They found it necessary to obtain component parts which were new to most of them, and they needed methods for testing both the component parts and their final products.

Under these conditions the Radio Frequency Laboratories was organized in Boonton, New Jersey. The staff consisted, at first, of one radio engineer, and their work concerned the manufacture of coil forms and other radio parts using insulating material. As time passed, additional technical personnel was added and the work of general engineering consultation was undertaken. This type of work naturally led to a good understanding of test equipment requirements.

In 1934, Mr. William D. Loughlin, who had been President of Radio Frequency Laboratories, together with several of his associates, formed the Boonton Radio Corporation. The first product of the new company was a Q Meter which read Q directly on a meter scale. Up until that time the measurement of Q had been made indirectly by use of bridges for measuring the effective reactance and resistance concerned. These measurements had been subject to error because of the techniques required, and useful measurements took a great deal of time. With the new Q Meter, measurements were simple and rapid, and the instrument proved capable of many additional valuable laboratory measurements on basic components and circuits. The flexible, accurate, easily used instrument was accepted almost immediately by the growing radio industry.

By 1941 a new model, replacing the earlier Q Meter, was introduced and the Company undertook development work on a frequency-modulated signal generator to meet the requirements for test equipment which the new frequency-modulated communication equipment demanded. Commercial instruments were made available and Boonton Radio Corporation continues to this date to make several forms of frequency-modulated test signal generators.

The early years of the Second World War brought the use of higher and higher frequencies, and a Q Meter similar to the earlier models, but applicable to higher frequencies, was designed. At the same time the activities of the Company were directed more and more to military applications. Its Q Meter and Frequency Modulated Signal Generators were widely used in military work and the Company produced a pulse modulated RF signal generator for use in testing radar systems. This instrument was produced in large quantities and is still used by all military services.

At the end of the War the FM Signal Generator was redesigned to permit coverage of a wider frequency range, to include AM as well as FM, and to obtain deviations in frequency which did not vary with carrier frequency. This instrument had very low leakage and a wide selection of accurately calibrated output voltages. It soon became the standard in its field and still maintains that position.

The aircraft transportation field in the 1940's was developing more accurate methods of navigation and better methods of landing in bad weather. A system for solving these problems was approved by the Civil

Aeronautic Administration and put in use both commercially and by the military services. Unusually accurate and specialized test equipment was required by this system and Boonton Radio Corporation was asked to undertake a design. A Signal Generator for Navigation equipment was produced in 1947 and an additional piece of equipment for testing receivers used in landing airplanes came very shortly after this. In 1952 the Company produced a more advanced model of the "Glide Path" testing equipment for the landing of aircraft.

In the last few years, the Company has turned its efforts to the development of self-contained, broad-band, flexible instruments containing RF bridges for measurement of components and cables. A new instrument, the RX Meter, was introduced which measures parallel resistance and parallel reactance of two-terminal networks over the LF and VHF ranges. The low frequency and high frequency Q Meters have been redesigned to include new features which increase the usefulness and accuracy of the equipment.

Companies, like people, have characteristics which identify them. From its formation to the present time, the Boonton Radio Corporation has built products of high quality. No attempt has been made to produce cheap instruments, and the quality and usefulness per invested dollar has been kept high. Close tolerances, high stability, mechanical soundness, and broad applicability have all been built into the Company's equipment. The Company regards its products as fine general-purpose tools for electronics craftsmen.

A Note From The Editor...

Since this is the first issue of THE NOTEBOOK, it seems appropriate to take a few lines to define the policies and purpose of our new publication. Briefly, THE NOTEBOOK has been planned and produced in order to distribute, to you and to as many interested persons as possible, information which we feel to be of value on the theory and practice of radio frequency testing and measurement.

In the past we have limited ourselves substantially to advertising, catalogs and instruction manuals for the broad distribution of such information. Inevitably, much important data was found to be too detailed for ads and catalogs; many new applications and techniques were learned or developed after publication of the instruction manuals. To provide a means, therefore, of informing you periodically of new methods and developments, and to furnish you with reference and background material of value in the application of our test equipment, THE NOTEBOOK has been established.

We feel that the name which we have selected is particularly appropriate, since much of the information which it will contain will be taken from our field and laboratory engineering notebooks, and since this and subsequent issues will, we believe, find a place in your own reference notebook. For the latter purpose, we have adopted standard notebook dimensions and punching in selecting our format.

Because the Q Meter is so well known and widely used, we have devoted most of the first issue to this instrument and the quantity which it measures. Our lead article discusses the nature of Q itself, using an approach somewhat different from the usual textbook handling of the subject. Then we have included some information on the recently-developed Q-Standard, a reference inductor designed to provide a check on Q Meter performance. A service note provides detailed information on the replacement of the thermocouple in the Q Meter Type 260-A. Finally, to introduce ourselves to you, we have included a brief outline of the history of our company.

THE NOTEBOOK will be published four times a year; in March June, September and December. A written request, giving your company, title and mailing address is enough to start you as a subscriber. If you have any suggestions, comments or questions concerning the contents or policies of THE NOTEBOOK, we would be happy to have you direct them to Editor, THE NOTEBOOK, Boonton Radio Corp., Boonton, N. J.

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The NOTEBOOK

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LO BS.
JXS

A Wide-Range VHF Impedance Meter

JOHN H. MENNIE, *Senior Engineer*

The design of a completely self-contained instrument for measuring impedance at Very High Frequencies entails, as might be expected, a number of interesting engineering problems. Mr. Mennie describes some of those encountered and solved in the development of the BRC RX Meter Type 250-A.

Ever since the middle of the last century when Sir Charles Wheatstone first put to practical use a curious electrical balancing network which had been devised a decade before by a fellow Englishman named Christie, the bridge circuit has been accepted as a valuable and powerful tool in the field of electrical measurement.

Even though today bridges, like almost everything else, have become increasingly complex in order to meet the demands of highly specialized applications, they retain, for the most part, the fundamental advantages of convenience, sensitivity and accuracy which characterized their more straightforward antecedents.

One of the applications in which bridge circuits have been particularly useful in recent years is the measurement of impedances at radio frequencies. A number of specialized circuits have been evolved for this purpose. One, the Schering Bridge, possesses certain features which make it outstanding. These features may be summarized as follows:

1. A constant relationship between the bridge elements is maintained regardless of the frequency impressed on the network.
2. Both of the basic variable bridge elements can be air capacitors, which are infinitely superior to other types of variable impedances for high frequency measurement work.
3. The circuit residual impedance can be kept small enough to permit compensation over a wide frequency band.
4. When arranged to measure parallel components of impedance, shielding problems are drastically reduced.

The fundamental circuit was first worked out by Schering in 1920 and proposed as a means for measuring dielectric losses at high voltages. Many other applications of this circuit have been suggested and used since then. For instance, in 1933 Dr. E. L. Chaffee suggested a form of Schering Bridge, as a method for measuring the dynamic input capacitance and resistance of vacuum tubes.

BALANCE EQUATIONS

The simplicity and wide frequency range of this bridge network can be appreciated by an analysis of the impedance relationships of Fig. 2 for the balance condition (i.e., zero voltage across the null detector).

$$Z_{AB}\,Z_{CD} = Z_{AD}\,Z_{BC} \text{ at balance,}$$

or

$$\left(R_2 + \frac{1}{j\omega C_2}\right)\left(\frac{1}{\frac{1}{R_4} + j\omega C_4}\right) = \frac{R_3}{j\omega C_1}$$

$$R_2 + \frac{1}{j\omega C_2} = \frac{R_3}{j\omega C_1}\left(\frac{1}{R_4} + j\omega C_4\right)$$

$$= \frac{R_3}{j\omega C_1 R_4} + \frac{C_4 R_3}{C_1}$$

Equating reals...

$$R_2 = \frac{C_4 R_3}{C_1}, \text{ and } \frac{R_2}{C_4} = \frac{R_3}{C_1}$$

Continued on Page 2

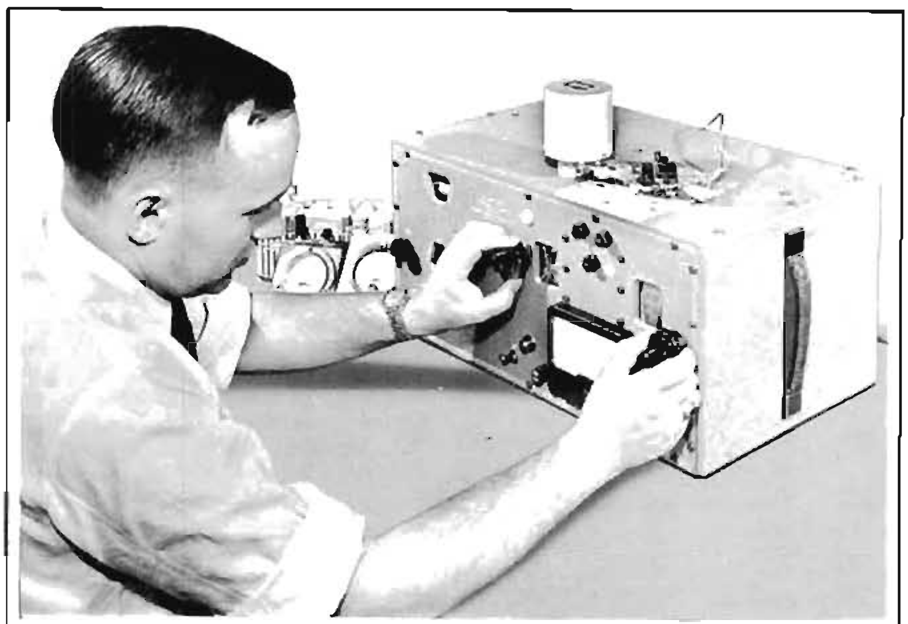

Figure 1. The RX Meter provides a simple, accurate means of measuring, independently, the RF resistance and reactance of a wide variety of materials, components and circuits. Dynamic measurements are possible, such as this one being made by C.G. Gorss, BRC Development Engineer, on a junction transistor.

YOU WILL FIND...

THE BRC NOTEBOOK is published four times a year by the Boonton Radio Corporation. It is mailed free of charge to scientists, engineers and other interested persons in the communications and electronics fields. The contents may be reprinted only with written permission from the editor. Your comments and suggestions are welcome, and should be addressed to: Editor, THE BRC NOTEBOOK, Boonton Radio Corporation, Boonton, N. J.

Impedance Meter (continued)

Equating imaginaries...

$$\frac{1}{j\omega C_2} = \frac{R_3}{j\omega C_1 R_4}$$

$$\frac{R_3}{C_1} = \frac{R_4}{C_2}$$

$$\therefore \frac{R_2}{C_4} = \frac{R_3}{C_1} = \frac{R_4}{C_2} \qquad (1)$$

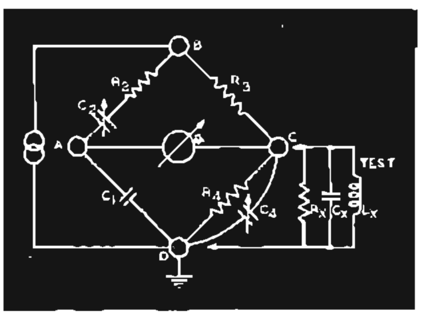

Figure 2. The Schering Bridge, arranged to provide measurement of parallel impedance components.

The test sample is connected across corners C and D of the bridge, and its parallel components of resistance and reactance effectively change the values of C4 and R4 in the circuit. In order to restore phase and amplitude balance conditions, the variable bridge capacitor C4 must be decreased by an amount equal to the equivalent parallel capacitance of the test sample. If the test sample is inductive, the capacitance of C4 is increased by an amount equal to the resonating capacitance of the parallel inductance.

The parallel resistance of the test is shunted across R4, reducing its value by a certain percentage which changes the R4/C2 ratio and unbalances the bridge. To restore phase and amplitude balance, variable capacitor C2 is reduced in value by the same percentage that R4 was reduced when shunted by the test resistance. The variable capacitor C_2 can thus be calibrated directly in terms of the parallel resistance (in ohms) of the component being measured.

THE RX METER

A refined version of the basic circuit described above forms the heart of the BRC RX Meter Type 250-A, which is designed to measure parallel resistive and reactive impedance components at frequencies from 0.5 mc to 250 mc. Unlike most RF impedance measuring devices using the null technique the RX Meter is a self-contained instrument including with the bridge circuit its associated oscillators, detector, amplifier and null indicator.

Figure 3 shows a block diagram of the instrument, the operation of which may be briefly described as follows:

The output of the test oscillator, (F1), which covers a frequency range of 0.5 to 250 megacycles, is fed into the bridge circuit. When the impedance to be measured is connected across one arm of the bridge, its parallel resistance and reactance components cause bridge unbalance, and the resulting voltage is applied to the mixer stage. The output of the local oscillator, (F2), which tracks at a frequency 100 kilocycles above F1, is also applied to the mixer, where it heterodynes with the bridge unbalance output frequency, F1, and produces a 100 kc difference frequency, having a magnitude proportional to the bridge unbalance voltage. This voltage is then amplified by a selective 100 kc amplifier to provide the desired bridge balance sensitivity. When the bridge controls are adjusted for balance (minimum indication), their respective dials serve as an accurate indication of the parallel impedance components of the test sample.

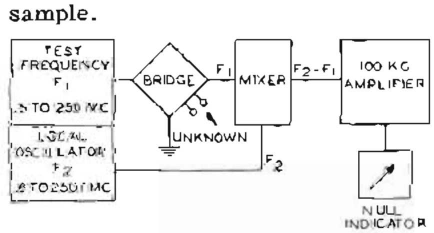

Figure 3. Block diagram of the RX Meter Type 250-A.

Particular care has been taken, in the mechanical design of the instrument, to provide adequate shielding of the oscillator and mixer in order to prevent spurious coupling between these stages. The detector is provided with automatic gain control which prevents the meter from reading off scale, thus greatly facilitating the balancing operation when approximate values of the impedance being measured are not known.

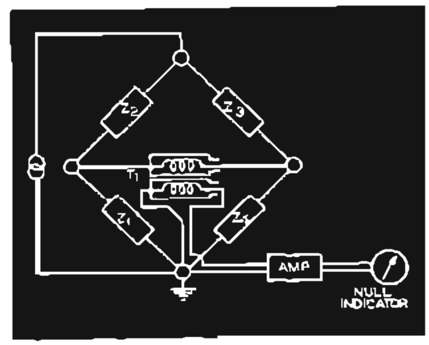

Figure 4. Conventional transformer coupling from bridge corners to amplifier.

Use of rugged castings, precision bearings and anti-backlash gears has permitted expansion of the calibrated parallel resistance scale to a useful length of twenty-eight inches, covering from 15 ohms to 100,000 ohms. The "R_p" scale of each instrument is individually calibrated and engraved in spiral form on the surface of a drum dial, the readability being one percent up to 5,000 ohms with an accuracy of approximately two percent over this range.

BRIDGE COUPLING CONSIDERATIONS

The absence of frequency terms in the balance equation (1) above suggests the applicability of this bridge circuit to impedance measurements over an almost unlimited frequency range. Several practical limitations do exist, however, and their effects must be considered in establishing the upper frequency limit for such a device. One of these involves the means used for connecting the signal source and null detector amplifier to the bridge. If the connection introduces excessive capacitance across the bridge arms or excessive loss in the oscillator or detecting system it can seriously affect the operation of the instrument at higher frequencies. This problem can be more readily appreciated by a study of Figure 4, which shows a conventional connection between the high bridge corners and the null detector, employing a special type of double-shielded transformer, T1.

A transformer of this type not only is difficult to design and expensive to manufacture but also has an upper frequency limitation imposed by leakage inductance as well as the usual low frequency limitation resulting from low impedance of the windings. The inter-shield capacitance is another handicap, as it must be incorporated in the bridge network, thus placing a severe limitation on both frequency and impedance ranges of the bridge. These shortcomings have been overcome by a new approach to the problem, as shown in Figure 5.

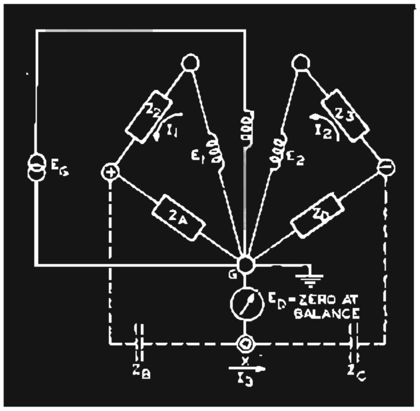

Figure 5. Specially devised system for coupling oscillator to detector.

The essential feature of this bridge network is that it is divided in two halves, one half being driven by voltage E_1 and the other half by E_2. Let us assume that voltages E_1 and E_2 are exactly equal in magnitude but opposite in phase, thus producing instantaneous currents I_1 and I_2 in the direction indicated. This arrangement enables us to detect the bridge balance conditions by coupling to the null detector through two very small but exactly equal capacitors, Z_B and Z_C. In the balanced condition, voltage E_D becomes zero, as indicated by our null detector, and the voltage across Z_B is exactly equal to that across Z_C. This is evident, since the same current, I_3, flows through both Z_B and Z_C when zero current is drawn by the detector branch. As X and G are at the same potential, it follows that Z_B may be considered effectively in parallel with Z_A, and Z_C is likewise in parallel with Z_D. Thus the voltage across Z_A is equal to that across Z_D. Now let Z_A and Z_B in parallel = Z_1 and Z_C and Z_D in parallel = Z_4. Figure 6 represents a simplified form of the bridge network for analysis purposes.

In Figure 6

$$E_1 = I_1Z_2 + I_1Z_1$$

$$E_2 = I_2Z_3 + I_2Z_4$$

$$E_1 = E_2 \text{ (by design)}$$

and $\quad I_1Z_1 = I_2Z_4 \quad$ at balance

then $\quad I_1Z_2 = I_2Z_3$

$$\therefore \frac{I_1Z_1}{I_1Z_2} = \frac{I_2Z_4}{I_2Z_3} \text{ and } Z_1Z_3 = Z_2Z_4 \qquad (2)$$

This impedance arm relationship is the same as that of a conventional bridge network.

COUPLING TRANSFORMER DESIGN

The above theory is based on the assumption that the transformer secondary voltages be extremely well balanced, and it is essential that this be the case over the entire frequency range of 0.5 mc to 250 mc. A very simple transformer was developed (Figure 7) consisting of three Formex-insulated wires twisted tightly together and wrapped once around a ferrite core ring. By using one turn the shunt capacitance and leakage inductance are held to a minimum for best high frequency performance. Leakage inductance is only 0.014 microhenry. The high permeability ferrite core serves to hold up the impedance of the transformer at the low frequency end. Interwinding capacitance C_{12} is ineffective because 1 and 2 are equipotential points. Although the transformer impedance is quite low, it effectively matches the very low output coupling loop impedance of the oscillator and drives the bridge with approximately one volt over the entire frequency range from 0.5 to 250 mc.

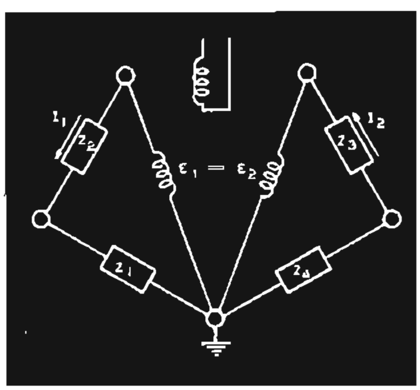

Figure 6. Simplified bridge network.

COUPLING CAPACITOR DESIGN

The capacitive network used to couple the unbalance output of the bridge to the detector must necessarily be very accurately balanced. To achieve such balance the following requirements were found to be essential:

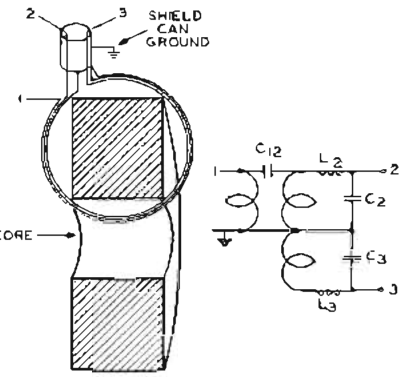

Figure 7. Balanced coupling transformer design.

1. Extremely fine adjustment to ±0.001 $\mu\mu$f.
2. Loss factor uniformly small for both capacitors.
3. Temperature coefficient uniformly small for both capacitors.
4. Mechanical stability to stand vibration and aging without change of balance.
5. Complete shielding to avoid pick-up to grid of detector from any other source than the two bridge corners.
6. Equal lead inductance to maintain effective capacitance balance to 250 mc.

The dual capacitor unit which was specially designed to fulfill these requirements is illustrated in Fig. 8.

RESIDUAL BRIDGE IMPEDANCES

As previously discussed, the bridge will balance at all frequencies, providing proper relationship between the effective capacitance and resistance is maintained up to the highest frequency to be used. Serious deviations will occur, however, if residual impedances are not either compensated for or made negligible by design.

Figure 9 shows two of the most troublesome residual inductances, L2 and L4. L2 represents the total inductance of series condenser C2

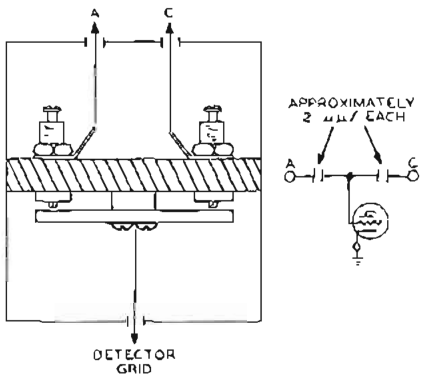

Figure 8. Balanced coupling capacitor design.

and resistor R2 and has a value of approximately 0.03 microhenry. This is sufficient reactance to completely neutralize the effective capacitance of the 20 μμf condenser C2 at 200 mc. Fortunately this residual inductance, L2, can be effectively removed from the circuit by means of a small shunt capacitance, Cp, across resistor R2. This shunt capacitance, as shown below, is equivalent to a series capacitance (C_s) whose reactance is very nearly equal and opposite to that of L2 up to the maximum frequency of 250 mc.

$$Z \text{ of parallel } R_2 C_p = \frac{R_2 - j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

The series capacitance reactance

$$\frac{1}{j\omega C_s} = \frac{-j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

For neutralization, let reactance

$$j\omega L_2 = \frac{-1}{j\omega C_s} = \frac{j\omega C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2}$$

$$\text{then } L_2 = \frac{C_p R_2^2}{1 + \omega^2 C_p^2 R_2^2} \cong C_p R_2^2$$

(within 2.4% at 250 mc.)

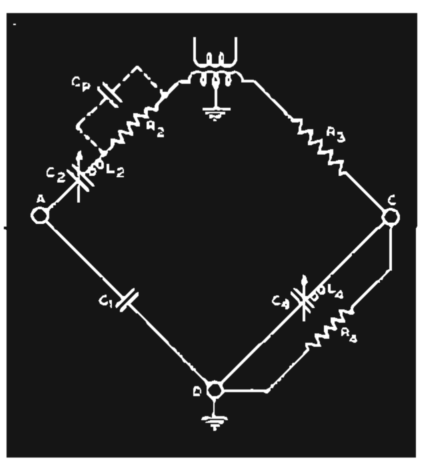

Figure 9. Residual impedance in bridge arms.

The second troublesome residual is the inductance, L4, associated with the variable condenser standard, C4. This inductance not only causes an effective capacitance increase in C4 at high frequencies but also produces an error in the resistance readings of high Q reactive components. The latter effect is the result of coupling between L4 and the inductive component of resistor R4. The best solution to this problem was found to be a series of fourteen specially designed edge-wiping rotor contact springs which provide fourteen individual parallel inductance paths through the condenser, each path having extremely low inductance. By this means, the inductance L4 is reduced to a value of only 0.0005 microhenry.

APPLICATION

The RX Meter has been found valuable in a number of diversified industrial and experimental applications. For example, it is being used to measure the RF characteristics of balanced and unbalanced transmission lines, of electrical components such as resistors, and of high and medium loss insulating materials such as phenolic vacuum tube bases.

The direct measurement of equivalent parallel resistance (R_p) of a circuit is particularly useful, since it represents the impedance seen by a vacuum tube or transistor when working into the circuit. Measurement of R_p also facilitates the determination of power dissipation in a tuned circuit since, when the voltage E across the tank is measured, power dissipation = E^2/R_p.

The RX Meter is particularly applicable to the measurement of transistor impedance since DC bias currents up to 50 milliamperes can be passed directly through the sample binding posts without harming the bridge elements. In such applications it is possible to lower the RF test voltage to as low as 20 mv with useable sensitivity.

CONCLUSION

The bridge methods conventionally used for RF impedance measurements generally require a composite test circuit consisting of a number of separate interconnected instruments. Setting up such a circuit for specific measurements is usually tedious and time consuming; leakage and undesirable coupling are frequently a problem at higher frequencies, and the resulting circuit is often limited in range and application. By combining all the necessary test components in a single, integrated instrument, the RX Meter provides ease and rapidity of measurement as well as unusual flexibility of application.

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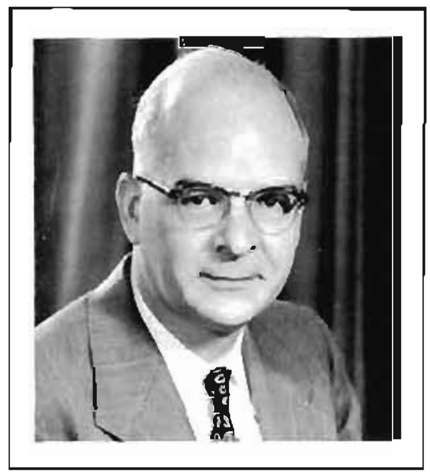
THE AUTHOR



John Mennie joined BRC as a Senior Engineer in 1951. His activities include the development of new instruments and equipment from idea to working model stage, and he is personally responsible for the basic design of the RX Meter Type 250-A.

Prior to 1951 he was employed as a design and development engineer at Western Electric Co., where he worked in close conjunction with Bell Laboratories on the development of specialized manufacturing techniques and a variety of measuring and test equipment.

Mr. Mennie was graduated from Stevens Institute of Technology in 1929 with a degree in Mechanical Engineering. He is a member of the Industrial Electronics Committee of I.R.E.

NOTE FOR OWNERS OF Q METERS TYPE 260-A

Final Instruction Manuals for this instrument are now off the press. If you have not received your copy please let us know. It will be forwarded immediately.

Q Meter Comparison

The recent redesign of the Q Meter Type 160-A resulted in a more accurate and flexible instrument - the Type 260-A. Here is a detailed discussion of the changes which were made and how they affect comparative Q readings on the two instruments.

When the Q Meter Type 260-A was introduced in 1953, its predecessors, the Types 100-A and 160-A, had been giving satisfactory laboratory service for a period of over 18 years, and had long since become a standard for Q measurements.

In spite of the fact that, during recent years, the Type 160-A had proved itself a useful and dependable tool, it had become apparent, in the light of new developments and improvements in the art, that its circuit had certain limitations. For this reason, the Q Meter Type 260-A was developed, retaining all the valuable characteristics of the 160A, but incorporating features which eliminated or minimized the limitations of the older instrument.

Because these design improvements will often cause small variations in Q readings between the two instruments, it seems desirable to investigate in detail the reasons for such apparent discrepancies and to formulate an expression for predicting them.

DESIGN DIFFERENCES

Both the 160-A and 260-A use the same basic measuring circuit, and the block diagram shown in Figure 1 applies to either instrument. This design, which makes use of the low impedance system of injecting voltage into the measuring circuit, is based on the resonant voltage rise principle described in a previous article.[1]

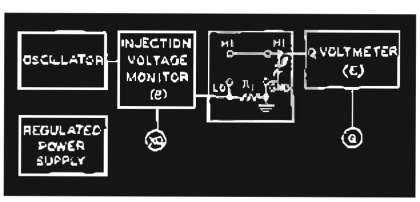


Figure 1. Q Meter block diagram.

Thus (for $Q \geq 10$) $Q = E/e$, where Q is the quality factor of the measuring circuit, e is the injected voltage and E is the voltage across the resonating capacitor.

The principle requirements for a successful Q measuring system are,

1. A low-distortion oscillator.
2. An accurate system for monitoring the injection voltage.
3. A low injection impedance across which e is developed.
4. A low-loss internal resonating capacitor.
5. An accurate high impedance Q voltmeter for measuring E.

Each of these must be carefully designed to minimize the residual parameters which tend to make the indicated Q (of the Q measuring circuit) differ from the effective Q of the unknown. Primarily, the discrepancies in Q readings which may be found between the two Q Meters at certain frequencies are caused by those improvements in the design of the 260A which result in lowered residual parameters.

SOURCES OF ERROR

1. Oscillator Harmonics

The calibration of the Q voltmeter is dependent on accurate indication of the ratio E/e. Errors in the Q readings can be caused by harmonics present in the oscillator output, due to the fact that the thermocouple used to monitor the current in the injection resistor produces a meter deflection proportional to the total heating effect of the distorted current, while, because of the selectivity of the measuring circuit, the indication of the Q voltmeter is proportional only to the fundamental component of the oscillator output. Although this effect is considerably reduced because of the square law response of the thermocouple,[2] it may still be noticeable when distortion is present. Early models of the 160-A (serial numbers below 4581) had some harmonic distortion in the oscillator output on the low frequency band (around 150 kc), which did introduce some error due to this effect. The effect is negligible in the 260-A because of the extremely low distortion present.

2. Injection Resistor

The injection resistor (R_i in figure 1) is in series with the resonant circuit, and when components are measured having low equivalent series resistance (high Q at high frequencies), the value R_i, which adds to the resistance of the unknown, becomes a large part of the total series resistance. This causes the Q Meter to indicate a Q value lower than that of the unknown component. Evidently it is desirable to keep the injection resistance as low as possible. For this reason, the 0.04 ohm injection resistor of the 160-A was reduced to 0.02 ohm in the 260-A.

It is possible for the reactive component of the injection resistance to cause a rise in injection voltage at high frequencies. Since the injection current is monitored, a constant e depends on a constant injection impedance, and it is important that the reactive component be reduced to a minimum. The dashed curve in figure 2 indicates the percent of rise in injection voltage caused by the residual 83 $\mu\mu$hy inductance present in the injection resistor of the 160-A. A new annular type of resistor has been used in the new Q-Meter, resulting in a residual inductance of the order of 0.035 $\mu\mu$hy, and reducing the error due to this inductance to a negligible amount, as indicated by the solid line curve.

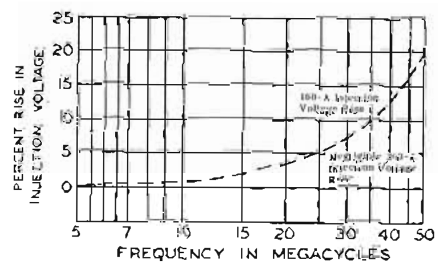


Figure 2. Calculated effect of injection resistor inductance on 160-A injection voltage at higher frequencies. 260-A resistor has negligible reactance, eliminating this source of error.

3. Resonating Capacitor

The construction of the resonating capacitor used in the Q Meter Type 160-A represents a satisfactory compromise between minimized residuals and the largest practicable capacitance range. As a result residual parameters are small enough to be ignored in the course of most measurements. This optimum mechanical design was retained in the new Q Meter and any residuals which exist may be regarded as the same for both instruments.

4. Voltmeter Circuit

The problem of providing an accurate voltmeter which will measure the RF voltage across the resonating capacitor to an accuracy of ±1% is a major one in the design of a Q Meter. In addition to maintaining the linearity (or scale calibration) characteristics, the voltmeter tube must have a very low grid current (with a grid leak of 100 megohms), the input capacity must remain nearly the same for different tubes, and the input conductance of the voltmeter circuit must be as high as possible.

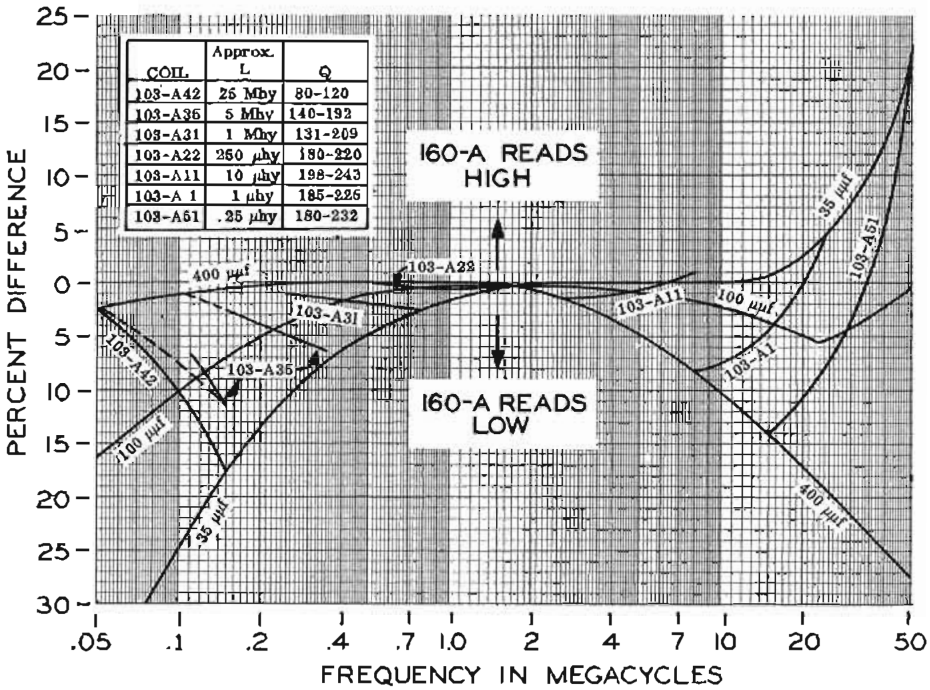


Figure 3. Difference in Q readings obtained by measuring a set of 103-A inductors on both 160-A and 260-A, with the readings of the latter used as a base. Each line labelled with a coil number indicates the percent difference in readings obtained by measuring the same coil, throughout its resonant frequency range, on both instruments. The lines labelled 35 uuf, 100 uuf, and 400 uuf respectively, indicate the percent difference in Q readings resulting at these internal resonating capacitances.

In order to achieve and maintain these characteristics, the use of a special tube is found to be imperative. The BRC 105-A is manufactured and tested specifically for application in the Q voltmeter circuit. A review of available tube types during the development of the Type 260-A confirmed the fact that this was the only tube which could meet such requirements, and it was therefore included in the design of the new instrument.

Several other changes in the voltmeter circuit were found desirable. The addition of "Lo Q" and "ΔQ" scales in the new instrument necessitated a slightly different operating point for the voltmeter tube. In addition, the physical arrangement of the grid circuit was changed to provide a higher natural resonant frequency, and additional bypassing was added in the plate circuit of the voltmeter, causing the Type 260-A to indicate Q more accurately at low frequencies and at low capacitance settings of the resonating capacitor.

TYPICAL EXPERIMENTAL DATA

To illustrate the differences which might be expected in Q readings between the Q Meters Types 160-A and 260-A, a set of standard inductors Type 103-A was measured on two representative instruments which were selected to be as nearly average as possible after a careful survey of Q readings on several hundred production units. Figure 6 is a plot of the readings obtained, and figure 3 indicates the percent difference in these readings vs frequency. Note that these graphs apply only to inductors with Q values in the ranges indicated. In general, the percent difference at any frequency is proportional to Q.

FORMULA FOR PREDICTING Q READING DIFFERENCES

The empirical formula at the bottom of this page has been suggested by R. E. Lafferty to describe the differences in Q readings obtained on the two instruments.

This formula is the result of combining actual measurements taken with both average instruments in a form indicated by the known theoretical differences in measuring circuits described previously. It will be observed that the terms of the formula are grouped according to frequency domain. Term A represents the difference in input conductance between the two Q voltmeters, term B corrects for the difference of insertion resistance value, and term C compensates for the inductance of the insertion resistor in the Type 160-A Q Meter. The latter term was derived using the average value of 83 μμhy for this inductance. It will be noted that each term is equal to 1 outside the frequency range specified. Thus, from 50 to 500 kc, only term A is significant, while from 500

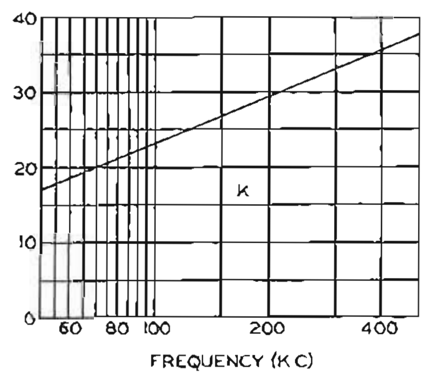


Figure 4. Values of the factor K, for use in term A of correlation equation.

kc to 5 mc, all terms are approximately 1 and the two instruments will have good agreement. From 5 to 10 mc, term B becomes significant, while terms A and C may be disregarded. Above 10 mc, B and C must be used. In general, the need for correction is proportional to the magnitude.

It should be noted that this correlation equation does not account for the effects of the harmonic content of the oscillator output of early Q Meters Type 160-A at the upper ends of the low frequency bands. Q

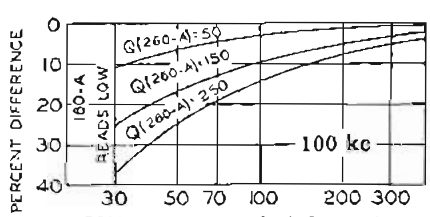


Figure 5. Calculated percent differences in Q readings between 260-A (used as base) and 160-A, at three values of Q as read on the 260-A.

$$\frac{Q160A}{Q260A} \approx \underbrace{\overbrace{\left[\frac{1}{1 + \frac{10^{-6}Q260A}{\omega CK}}\right]}^{\text{Term A}}}_{\text{50-500 kc}} \underbrace{\underbrace{\overbrace{\left[\frac{1}{1 + \frac{\omega CQ260}{50}}\right]}^{\text{Term B}}}_{\text{5-10 mc}} \overbrace{\left[\sqrt{1 + 4.32 \times 10^{-18}\omega^2}\right]}^{\text{Term C}}}_{\text{10-50 mc}}$$

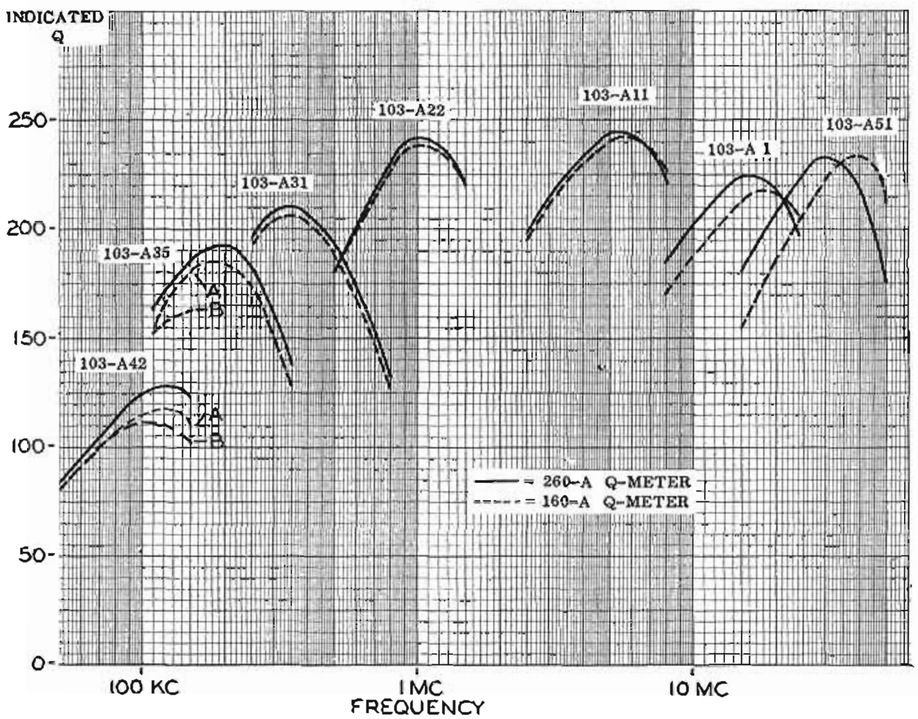


Figure 6. A comparison of Q readings obtained from duplicate measurements on both Q meters of a set of 103-A inductors. Dotted curves marked "B" indicate values obtained on older 160-A's having appreciable oscillator distortion in this range; curves marked "A" indicate 160-A's with reduced distortion.

Meters Type 160-A having serial numbers above 4581 have oscillators with reduced harmonic content.

Figure 5 illustrates the percent by which the older Q Meter will read low at 100 kc, for three values of Q. These curves were calculated by using term A of the correlation equation.

SUMMARY

The difference observed between the readings obtained on a Q Meter Type 160-A and a Q Meter Type 260-A for a given component is caused by the reduction of residual circuit parameters in the latter instrument, permitting more accurate measurement. The chart below is a summary of the causes and effects of these Q reading differences as they apply for each frequency range.

The correction equation and other data presented in this article are the result of information obtained from average Q Meter readings, and do not allow for variations within specified tolerances. Application in individual cases may require study of the parameters of the instruments concerned.

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3. "RF Microvoltages", Myron C. Selby, National Bureau of Standards.

Range	160-A reads	Cause	Term A	Term B	Term C
50-500 kc	Low	160-A voltmeter loads measuring circuit at low resonating capacitance and frequency. Early models read 3% lower at 100 kc and 8% lower at 150 kc because of oscillator harmonics. This effect is not included in Term A.	$\frac{1}{1+\frac{10^{-6}Q}{\omega CK}}$	1	1
500kc - 5 mc	Substantially in agreement with 260-A		1	1	1
5 mc - 10 mc	Low	Effect of higher 160-A injection resistance becomes appreciable with increased frequency, Q, and resonating capacitance	1	$\frac{1}{1+\frac{\omega CQ}{50}}$	
10 mc - 50 mc	Low at high C	Injection resistance effect continues, causing low readings with high resonating capacitance.	1	$\frac{1}{1+\frac{\omega CQ}{50}}$	$\sqrt{1+4.32\times10^{-18}\omega^2}$
	High at low C	Reactance of injection resistor increases injected voltage with frequency, causing high readings at low resonating capacitance and high frequency.			

Service Note

RX METER BRIDGE TRIMMER ADJUSTMENT

At frequencies above 100 MC the zero balance of the RX Meter bridge circuit is necessarily sensitive to extremely small variations in internal circuit capacitance. It is possible that minute shifts in the relative position of circuit components, caused by excessively rough handling in shipping, etc., may alter the effective capacity enough to make it impossible to obtain a null indication on the highest frequency range by adjusting ZERO BALANCE controls, "R" and "C".

In most cases, this situation can be corrected by the following screwdriver adjustments:

1. Allow the instrument to warm up, set the oscillator frequency at 200 mc, and adjust the detector tuning control as described in the Instruction Manual, with the C_p dial at 0 and R_p at ∞.

2. With a screwdriver, pry up the small metal cap located near the rear of the ground plate on top of the instrument. This provides access to a small trimmer capacitor having a vertical, slotted adjusting shaft.

3. Rotate the "R" knob and note whether the null indicator reading decreases with (a) clockwise, or (b) counter-clockwise, rotation.

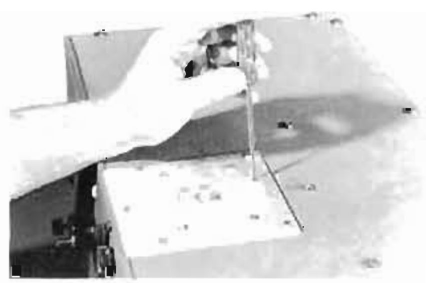


Figure 1. Adjusting RX Meter trimmer.

4. Using the screwdriver, rotate the trimmer shaft about 1/8 turn—clockwise in case (a) above, counter-clockwise in case (b) above. Then remove the screwdriver * and try to obtain balance with the "R" and "C" knobs. If a null indication still cannot be obtained, rotate the trimmer another 1/8 turn in the same direction. Continue this procedure until balance can be obtained.

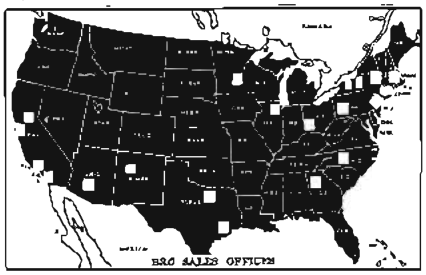
5. Check the balance at a frequency of 250 mc and repeat the above adjustment if necessary.

*Correct null indication can not be obtained while the screwdriver (or aligning tool) is near or in contact with the trimmer shaft.

BRC SALES ORGANIZATION

Frank G. Marble, *Sales Manager*

As is the case with most industrial firms, Boonton Radio Corporation makes most of its contacts with the outside world through its Sales Department. This, of course, is altogether reasonable, since the basic function of a Sales Department is actually that of communication-- the origination and transmission of data to customers, potential customers, and those seeking specific technical information.



Since many readers of THE NOTEBOOK will, at some time, fall in at least one of these catagories, we feel that it may be of interest to describe, very briefly, the set-up of our technical sales organization as it will affect them.

Our sales organization may be considered as having two major subdivisions, the internal Sales Department, and a nation-wide network of Sales Representatives. The former is an integral part of the company, and is responsible for a number of widely diversified functions. These include the routine work of processing orders, answering inquiries, expediting shipments, preparing quotations, etc., as well as supervising advertising and sales promotion activities and catalog and instruction material. The Sales Department also has the responsibility of developing and distributing new application information needed by customers, and of keeping abreast of the trends and requirements of the industry. In addition, company Sales Engineers handle, directly, instrument sales in the New York City, New Jersey area as far South as Washington, DC.

The second subdivision of our organization is comprised of Representatives who are located throughout the country in those areas where the electronics industry is most heavily concentrated. These organizations maintain staffs of experienced sales engineers who are throughly familiar with the operation and application of BRC instruments. Each Representative is the exclusive BRC agent in his territory, and is qualified to supply complete information on our full line of equipment.

A NOTE FROM THE EDITOR

We were extremely pleased and, to tell the truth, a little amazed at the response which greeted the first edition of THE NOTEBOOK. The number of returns received has forced us to revise completely our original estimates of the size of future printings. A few figures might be interesting. We mailed roughly 40,000 copies of the first edition to a selected group of engineers, scientists, and educators. As a direct result of the reply cards returned, we are mailing this second edition to a list of 19,000 readers, and it is still growing daily. In addition, we have received several hundred notes and letters expressing interest and encouragement.

Such a vote of confidence is deeply appreciated. We will do our best to merit it in the future by maintaining, as nearly as possible, the level established in our first issue.

CORRECTION: Somewhere along the way, in preparing NOTEBOOK No. 1, an exponent was lost. The first equation in column 1, page 6, should read:

$$L_e = \frac{L}{1 - \omega^2 L C_d'}$$

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Number 3

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Signal Generator And Receiver Impedance

TO MATCH . . . OR NOT TO MATCH

W. CULLEN MOORE

Engineering Manager

The above question might logically be followed by two additional queries: "If not, why not?" and "If not, what then?" These are loaded questions. The Technician in his screened room as well as the Engineer at his desk had best tread carefully lest he bring down upon himself a barrage of counter-queries. Perhaps we can provide a few good rounds of ammunition for both sides, and may the best man win.

The purpose of a signal generator is to make available in the Laboratory a calibrated source of radio frequency signals which is equivalent to the antenna system with which the receiver eventually will operate. The signal generator, with the associated dummy antenna, must therefore behave the same as the antenna system with respect to ability of the receiver to absorb energy. Failure to meet this condition invalidates the signal generator as a substitute for the antenna.

Let us look briefly at an interesting situation. Failure to take into account the internal impedance of an antenna system can very easily lead to an error of 2:1 in the realizable sensitivity of a receiver as compared to the measured value. It would require a change of 4:1 in the antenna power of the transmitter at a fixed distance to compensate for the apparent error of 2:1 in the sensitivity of the receiver.

Other considerations, such as signal-to-noise ratio or selectivity, may impose contradictory requirements on the input impedance characteristics of a receiver as compared with the requirements dictated for maximum microvolt sensitivity alone. In such a case, what does the calibration of the signal generator tell us about the actual sensitivity of the receiver?

The necessity for making the same measurement on a receiver using different generators or comparing the results of measurements on different receivers requires an understanding of what we are about when making measurements of receiver sensitivity.

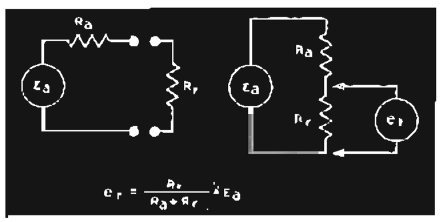

Figure 1. I.R.E. Standard for introducing signal voltage E_a through antenna impedance R_a into receiver input impedance R_r.

DUMMY ANTENNAS

The sensitivity of a radio receiver is NOT the number of microvolts applied directly to the input terminals of the receiver to produce standard output, even though this frequently is assumed to be the case.

The Institute of Radio Engineers has defined the INPUT SENSITIVITY OF A RECEIVER as the number of microvolts required to produce standard output when applied through a dummy antenna having the characteristic impedance of the antenna with which the receiver is intended to operate, to the input terminals of the receiver.[1,2]

To appreciate the logic leading to this choice let us consider the source of energy from which the combined system of the antenna and receiver is driven. Electromagnetic energy flowing in free space encounters a conductor and excites in it a voltage which acts in series with the antenna radiation resistance. Like the open-circuit electro-motive-force of a battery this voltage is available to us only in series with the internal impedance of the power source itself. The variation of this impedance with frequency may require a series-parallel combination of R, L, and C in the dummy antenna. Part of all of it may be contained in the signal generator output impedance.

POWER TRANSFER

We have mentioned above that considerations of signal-to-noise-ratio, selectivity and the like may have an effect on our choice of load for the antenna system and it is useful to see how the power transfer from the antenna to the receiver will be affected by deliberate mis-matches in load. Figure 1 is the basic circuit describing the test conditions specified by the Institute of Radio Engineers in which we have conveniently assumed that only resistances are involved. The total power intercepted by the antenna system will be dissipated in two portions of the network (1) the internal impedance of the antenna and (2) the input load presented by the receiver. The power delivered to the receiver load will equal:

$$P_r = \left[\frac{R_r}{(R_a - R_r)} E_a\right]^2 \left(\frac{1}{R_r}\right) = \frac{R_r \; E_a^2}{(R_a + R_r)^2}$$

Figure 2 indicates the efficiency

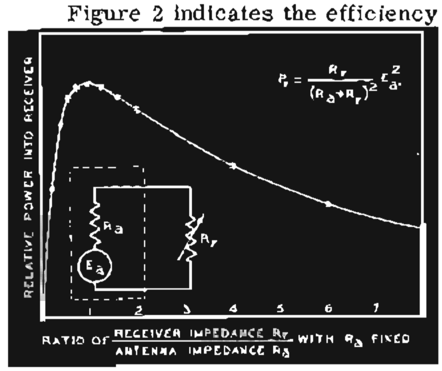

Figure 2. Effect of variation of receiver input impedance R_r on power into receiver with fixed antenna impedance R_a and voltage E_a.

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of power transfer which takes place as we vary the ratio of load to antenna resistance. This curve shows that the maximum power at the input terminals of the receiver is obtained when the receiver input impedance matches the antenna impedance. The maximum is fairly broad, but a relatively small shift downward in the input impedance of the receiver results in a very large change in the amount of power delivered to the receiver as compared to the effect of an equal increase in the receiver input impedance.

MIS-MATCHING FOR IMPROVED SELECTIVITY

Up to this point we have been discussing the title of this paper, namely: "To match or not to match". We are now faced with the second question. "If not, why not?" The curve we presented in Figure 1 shows the way in which power is transferred from one resistive circuit to another. A similar relationship exists for coupling between primaries and secondaries of tuned transformers.

As the coupling is increased, the effective Q of the resonant winding decreases until a point is reached at which the Q drops to one-half the uncoupled value. At this point (i.e. critical coupling), there is the optimum energy transfer between the circuits.

However, that value of coupling which produces optimum energy transfer has simultaneously dropped the Q of our resonant selective circuit to one-half and thereby degenerated the selectivity of the front end of the receiver. In order to obtain better selectivity characteristics, we may deliberately mis-match the receiver to the generator to reduce loading on the resonant circuit. We will shortly find out how this mis-match can be accounted for in our measurements.

MIS-MATCHING FOR IMPROVED SIGNAL-TO-NOISE RATIO

Another consideration which may lead to a deliberate mis-match of the receiver to the antenna impedance is the necessity for improving the signal-to-noise-ratio over that which would be obtained from a perfect match.[1]

The noise and the signal are intermixed in the antenna and the receiver should select them as favorably as possible. The noise voltage generated in the antenna is proportional to the square root of the antenna impedance. The noise power induced into the matched input circuit of a receiver is independent of the receiver input impedance. The signal power for a given voltage is, however, a function of the input impedance of the receiver. The ratio of signal to noise can therefore be affected by deliberately mis-matching the receiver to the antenna impedance in exchange for a loss in microvolt sensitivity.

SIGNAL GENERATOR CALIBRATION

Having seen above that there are reasons to match and reasons not to match the receiver input to the antenna impedance there remains to investigate the question: "If not matched, what then?" It should be carefully noted that the IRE standards requiring the use of a dummy antenna say nothing whatsoever about the impedance of the receiver.

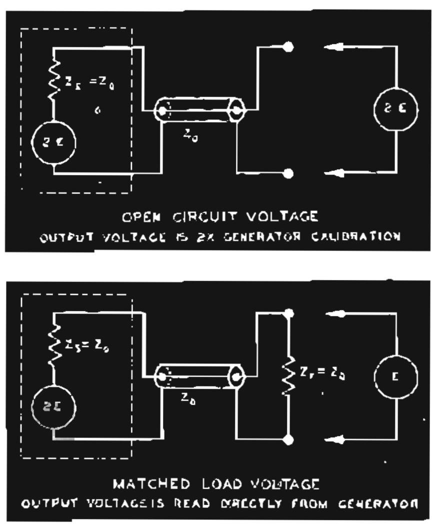

Figure 3. Output voltage calibration for a generator having its source impedance equal to the characteristic impedance of the output cable.

To use a signal generator intelligently we must understand how the output system behaves under different conditions. Signal generators can be divided roughly into two classes: a. Low source impedance, and b, matched source impedance. In low source impedance generators, the output transmission line is driven directly by a pick-up loop having the lowest possible inductance. In a matched source generator, a low inductance pick-up loop provides a very low impedance source of voltage which feeds through a matching resistor to the output cable. The value of the resistor is carefully controlled to match the characteristic impedance of the transmission line.

The output meters and dials of signal generators of both the matched and low internal impedance varieties are almost universally calibrated in terms of the voltage developed at the output jack on the front panel of the generator when this jack is terminated in a resistive load equal to the characteristic impedance of the coaxial line or the rated output impedance of the generator.

The heart of the problem lies in the length of coaxial cable commonly used to connect the point of generation of the radio frequency signals to the input to the receiver. At the frequencies generally encountered in communications and television this length of cable can approach and exceed 1/4 wavelength. A 1/4 wavelength piece of transmission line has a transforming property for both impedance and voltage which acts somewhat like a teetertotter, the midpoint of which is the characteristic impedance of the line. A low-loss 1/4 wavelength piece of line driven by a low impedance source will produce a very high voltage at the open circuit end of the line. Conversely, if the driving source impedance is high the output voltage will be low. If the source impedance equals the characteristic impedance of the line, the output voltage will equal the input voltage.

In matched signal generators the calibrated voltage is fed through the characteristic impedance of the generator and the connecting cable to the matched terminating load. In order for the dial to be calibrated in terms of the voltage, E, developed across the load alone it is necessary to deliver twice this voltage, 2E, to the input to the internal generator impedance. This means that the open-circuit voltage available at the front panel, 2E, will be twice that obtained, E, when the output jack is terminated in a matching impedance, and hence twice the dial calibration, as shown in Figure 3.

Looking back into the generator from the end of the connecting line one sees a properly terminated line. Therefore the length of line has no effect on the voltage at the output.

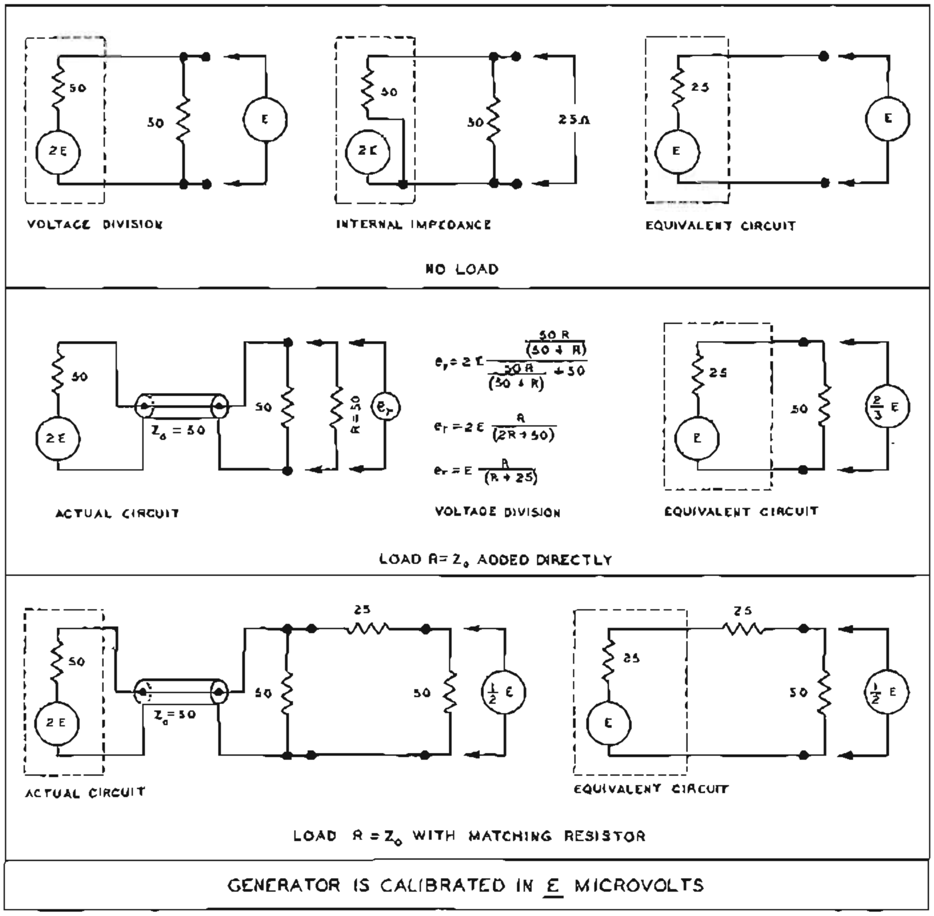

Figure 4. Application of Thevenin's Theorem to a cable terminated in Z_o.

TERMINATED CABLES

A terminating resistor frequently is used in the end of extension cables, and the voltage developed across it is exactly equal to the reading on the dials. By the application of Thevenin's theorem, the equivalent circuit of the generator now appears to be the output voltage acting in series with the paralleled matching resistor and the generating source. Thus, as we see in Figure 4, we have the equivalent of 25Ω internal impedance acting in series with the indicated number of microvolts. (Ref. 3, p. 47)

If we wish to match 50Ω it is necessary to add an additional 25Ω in series with the output cable. This gives us the equivalent of an antenna having induced in it E microvolts and having an internal impedance of 50Ω. Only half of the antenna voltage is delivered to the input terminals of the receiver as shown in Figure 4.

To the engineer struggling to meet receiver sensitivity specifications this may look like hitting oneself over the head just for the fun of it. In fact, the apparent loss of receiver sensitivity caused by accounting for the antenna impedance when making measurements with a signal generator has given rise to the expression of "hard" vs "soft" microvolts. One must work much harder to obtain sensitivity with "hard" microvolts than with "soft" ones, which pour directly out of a low impedance signal generator. Unfortunately, microvolts are hard to get out of all antenna systems.

ATTENUATOR PADS

Let us consider the impedance and voltage distribution along another commonly used system; that in which a 6 db, or 2:1, attenuator pad has been used in the line between the generator and receiver. Figure 5 shows the arrangement of the elements.

The output of the pad is equivalent to the characteristic impedance of the generator in series with E microvolts. A 20 db, or 10:1, pad is equivalent to E/5 microvolts in series with the generator impedance where E is the reading on the generator dial. A pad can be designed to match the generator output to a different value of receiver input impedance The 2:1 or 10:1 voltage division acts on the 2E supplied in series with the 50 ohm internal generator impedance

UNMATCHED LOADS

We have implied above that the output system of an internally matched generator will maintain 2E microvolts at the input to its internal impedance regardless of the termination at the end of the coaxial cable. This can be shown as follows.

In signal generators using a piston-type (or wave-guide-below-cut-off) attenuator the metered voltage is generated in a primary coil at the input end of the attenuator tube. At any particular distance down the tube a definite amount of voltage, 2E, where E is the voltage shown on the generator dials, will be induced in the secondary loop which drives the output transmission line. In an internally matched generator the matching resistor is located between the low-impedance pick-up loop and the input to the transmission line.

The arrangement described above forms a transmission line having a zero-impedance voltage source, E_g = 2E, in series with the sending end impedance, Z_s, which is matched to the characteristic impedance of the line, Z_o. The resulting voltage at the receiving end, E_r, as the receiving end impedance, Z_r, is varied can be derived from the transmission line equation for I_r (Ref. 3, p. 139., eq. 29):

$$E_r = I_r Z_r = \frac{2E_g Z_o Z_r}{(Z_o+Z_r)(Z_g+Z_o)e^{\gamma l}+(Z_o-Z_r)(Z_g-Z_o)e^{-\gamma l}} \quad \text{(e)}$$

If $Z_g = Z_o$

$$E_r = \frac{2E_g Z_o Z_r}{(Z_o + Z_r)(2Z_o)e^{\gamma l}}$$

$$E_r = \frac{E_g Z_r}{(Z_o + Z_r)e^{\gamma l}}$$

For a lossless line, the "propagation constant, γ, reduces to a "phase shift factor", $j\beta$, which usually is of no interest to us. The output voltage then is,

$$E_r = \frac{Z_r}{Z_o + Z_r} E_g$$

In the above derivation no restrictions were placed on the length of line or the receiving end impedance, provided the sending end impedance is matched to the characteristic impedance of the line. This result is logically appealing. Looking back into the generator from the receiving end of the lossless cable we see a matched load regardless of the length of cable. Even if we reduce it to zero, we are left with the matching impedance, $Z_g = Z_o$, in series with a source of voltage, E_g = 2E, where E is the dial calibration.

MULTI-FREQUENCY MEASUREMENTS

In the above discussions we have treated relatively simple configurations of resistors and cables. Actually, dummy antenna systems may become quite complicated, involving series and parallel combination of resistors, inductors, and capacitors in order to duplicate the frequency characteristics of the antennas being used with the receiving equipment.[1,2] The values obtained for intermediate frequency direct transmission and for image rejection ratios in superheterodyne receivers may be seriously affected by the accuracy of the dummy antenna since quite large ratios exist between the resonant frequency and the frequency under test. The impedance mis-match caused by off-resonance operation of the input system of the receiver, such as when taking standard selectivity curves, produces wide impedance variations which affect the calibration of the output system of a non-matched generator, and hence the validity of the selectivity curves obtained.

CONCLUSION

The function of the signal generator and dummy antenna is to reproduce in the laboratory the conditions presented by the antenna system with which the receiver is to operate.

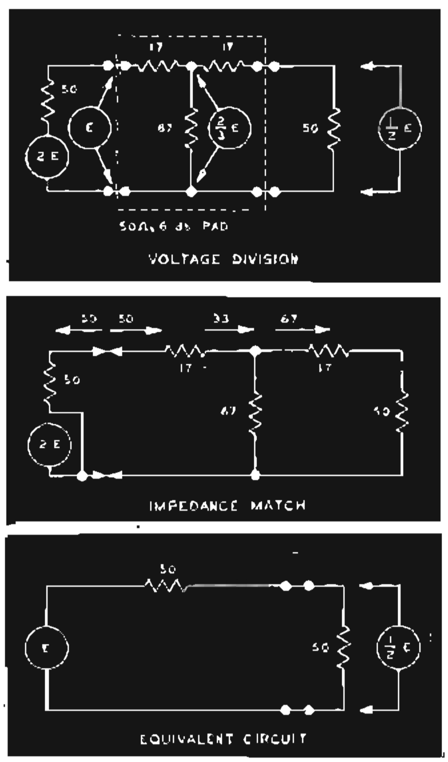

Figure 5. Equivalent circuit of a 50-ohm, 6 db symmetrical attenuating pad.

The output system should place at the end of a probe the equivalent antenna voltage and internal impedance.

The standards of measurement for RECEIVER SENSITIVITY established by the IRE reflect the physical requirements that the signal input from the antenna shall be delivered through a dummy antenna representing the antenna impedance.

The input impedance characteristics of the receiver should be designed to work with the impedance characteristics of its associated antenna system. This may or may not result in actual matching of the two impedances.

Signal generator calibrations are valid only when the output is terminated in a specified value of impedance.

The equivalent circuit for the output impedance of a matched signal generator and the impedance of the load reduces to a simple voltage divider having twice the indicated voltage across the divider regardless of length of cable or value of load impedance.

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Transmission Line Measurements WITH THE RX METER

NORMAN L. RIEMENSCHNEIDER, *Sales Engineer*

The RX Meter has found wide application in the measurement of characteristic impedance, attenuation, and velocity of propagation of transmission lines. The rapidity with which the measurements can be accomplished on relatively short pieces of cable and the accuracy realized in using the simple, direct techniques described below has made the instrument a valuable tool for the design engineer concerned with coaxial elements, as well as for the quality control engineer making spot checks during the manufacture of cable.

In this article we will describe, briefly the methods used in measuring each of the above characteristics with the RX Meter. In order to clarify the approach used in each case we have included, at the end of this article, a brief review of the basic transmission line formulas involved.

CHARACTERISTIC IMPEDANCE

Several methods are available for the measurement of characteristic impedance (Z_0) with the RX Meter. One of the most satisfactory involves the familiar relationship, $Z_0 = \sqrt{Z_i Z_L}$, where Z_i is the input impedance of a quarter-wavelength line with a given termination, and Z_L is the impedance of the termination itself. For our purpose, this relation is used in the form $Z_0 = \sqrt{R_1 R_2}$, where R_1 is a resistance measured directly on the RX Meter terminals and R_2 is the input resistance of the quarter-wave line terminated by R_1. The actual procedure used for this measurement is as follows:

The RX Meter oscillator is adjusted to the desired measuring frequency. A piece of the sample cable is cut to a length corresponding to approximately one quarter-wavelength with both ends dressed back about one-half inch to espose the center conductor and shield. If the cable dielectric is polyethylene, this length may be taken directly from Figure 1. The RX Meter is first balanced with nothing attached to the terminals and with the C_p and R_p dials set at 0 and ∞ respectively. The bridge is then rebalanced by means of the R_p and C_p dials with the cable, shorted at the far end, attached to the binding posts. If the cable length is correct, (i.e. exactly $1/4\lambda$), the C_p dial reads zero. If it reads capacitive, the frequency should be lowered or the cable shortened, while if the C_p reading is negative (indicating inductance) the frequency should be increased or a longer piece of cable used. Since the characteristic impedance does not change significantly at frequencies above approximately 20 mc., it is usually more convenient to make any needed adjustment by varying the frequency. The R_p dial must be used to null the instrument during the above procedure, but its value may be disregarded.

For the termination, select a 1/2 watt carbon resistor whose resistance is roughly equal to the estimated characteristic impedance of the cable. If the latter cannot be estimated, 50 ohms will usually suffice. Removing the short circuit from the end of the quarter wavelength line, connect the resistor in its place, keeping the leads as short as possible. Then balance the bridge and record R_p as R_1. The resistor should then be removed from the end of the cable and measured directly across the RX Meter terminals to provide the value R_2.

In a typical measurement, made on a quarter wavelength section of RG 58/U cable, a resistor which measured 47.0 ohms directly at the bridge terminals was used to terminate the line. The line with this termination measured 63.8 ohms. Then

$$Z_o = \sqrt{63.8 \times 47.0} = 54.7 \text{ ohms}.$$

An equally satisfactory method of determining Z_o (that recommended in Military Specification JAN-C-17A) is based on the relationship

$$Z_o = \frac{101{,}600}{v \times C},$$

where v is the velocity of propagation factor in percent and C is the cable capacitance in $\mu\mu f$ per foot.

The latter value is determined by attaching a very short length of the cable to the RX Meter binding posts and measuring C_p directly. The velocity of propagation may be determined as described in a later section.

A third method of measuring Z_o may be worth mentioning, although less satisfactory with respect to accuracy.

This method is implied by equation (3) at the end of this article, which indicates that the characteristic impedance of a line is equal to the absolute value of the reactance of a section 1/8 wavelength long.

To obtain the correct length, a 1/4 wave section is first established in the manner described above, at a frequency twice the desired measuring frequency. This frequency is then halved, and reactance of the section (either open or short-circuited) determined from the C_p reading at balance.

ATTENUATION

A very convenient method of measuring attenuation, using short pieces of cable, is provided by the equation

$$\alpha L = \frac{Z_o \times 8.69 \text{ db}}{Z_l},$$

where α is db per unit length and L is length. Here the value of Z_l is determined by measuring the parallel resistance (on the R_p dial) of a piece of cable 1/2 wavelength long. If the frequency is such that a half wavelength is less than approximately 4 feet, a one-or three-halves wavelength piece can be used, with no change in procedure, to minimize the effect of irregularities in the cable, etc. The attenuation in db obtained for the length of cable tested can be readily adjusted to db per 100 feet.

When the desired frequency has been selected, cut the cable to one-half wavelength, and dress one end back one-half inch. After effecting the initial balance of the bridge, connect the cable, with the far end open-circuited, to the RX Meter, making sure that the center conductor is connected to the "HI" post. The bridge should now be balanced and the values of capacity (C_p) and parallel resistance (R_p) read from the respective dials. If $C_p = 0$, the cable is the proper length and the value obtained for R_p substituted for Z_l in the above equation. If the C_p dial indicates a capacity, the cable is too long and a small amount must be cut off the far end, or the frequency must be lowered. If the C_p dial indicates a negative

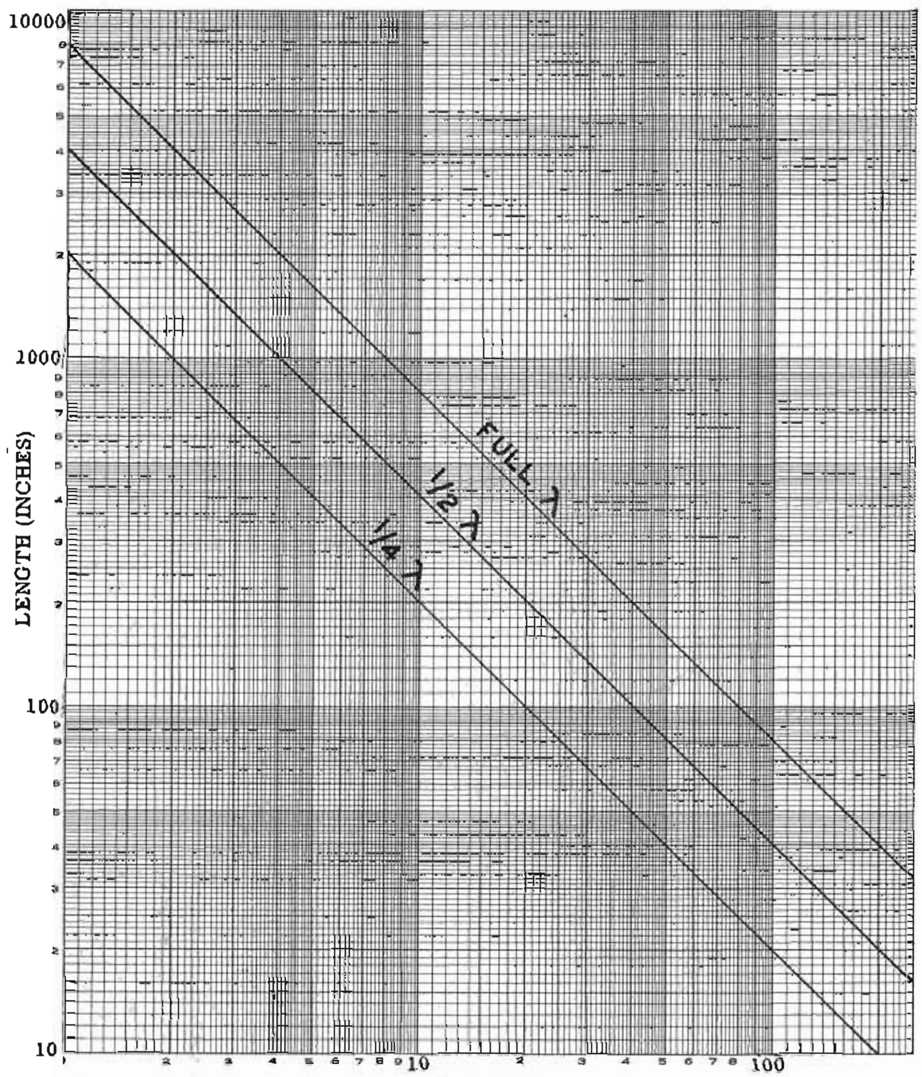

Figure 1. Wave length vs frequency for coaxial cables having polyethylene dielectric.

capacity (inductance), a longer piece must be used or the frequency raised. As an example, a one-half wavelength section of RG-58/U cable at 77 mc (52" long) was found to have an R_p of 2760 ohms. Applying this value to the formula above, together with the known characteristic impedance of 54.7 ohms, and adjusting for the length of the section, the attenuation was found to be 3.98 db/100 feet.

VELOCITY OF PROPOGATION

Since the velocity of propagation factor is equal to the physical length of a one-half wavelength section of cable divided by the length of a one-half wavelength section in air, it is merely necessary to measure the physical length of the cable involved in the preceding measurements and to compare it to the equivalent wavelength in air.

TRANSMISSION LINE EQUATIONS

The general formula for a lossless line of length l, having a characteristic impedance of Z_o, and terminated in an impedance Z_L is,

$$Z_i = Z_o\left(\frac{Z_L \cos\beta l + jZ_o \sin\beta l}{Z_o \cos\beta l + jZ_L \sin\beta l}\right) \quad (1)$$

where the phase constant $\beta = 2\pi/\lambda$, and λ = wavelength. Now, if the line is 1/4 wavelength long, $l = \lambda/4$ and

$$\beta l = \frac{2\pi}{\lambda} \times \frac{\lambda}{4} = \frac{\pi}{2} \text{ radians, or } 90^\circ.$$

Substituting in (1) above,

$$Z_i = Z_o\frac{(+jZ_o)}{(+jZ_L)} = \frac{Z_o^2}{Z_L} \text{ and}$$

$$Z_o = \sqrt{Z_i Z_L}. \quad (2)$$

If the line is 1/8 wavelength long and is short circuited, then $l = \lambda/8$, $\beta l = \pi/4$ radian or 45^o, and $Z_L = 0$. Substituting in (1), (3)

$$Z_i = Z_o\left(\frac{+jZ_o \sin 45^o}{Z_o \cos 45^o}\right) = -jZ_o = X.$$

In a similar manner it can be shown that the input impedance of a 1/8 wavelength line that is open circuited is,

$$Z_i = jZ_o. \quad (4)$$

For the purpose of deriving a means of measuring the attenuation of a transmission line, the general expression for a line with loss is given below.

The impedance, Z_i^l, looking into a line with loss, having a characteristic impedance of Z_o, and terminated in an impedance Z_R can be expressed as,

$$Z_i^l = Z_o\left(\frac{Z_R + Z_o \tanh\gamma l}{Z_o + Z_R \tanh\gamma l}\right) \quad (5)$$

where, $\gamma l = \alpha l + j\beta l$, and $\beta = 2\pi/\lambda$.

In the case of a half wavelength line, $l = \lambda/2$ and

$\beta l = 2\pi/\lambda \times \lambda/2 = \pi$ and
$\gamma l = \alpha l + j\pi$. Also,
$\tanh\gamma l = \tanh(\alpha l + j\pi) = \tanh \alpha l$,
and if αl is small, then, $\tanh\alpha l = \alpha l$,
and $\tanh \gamma l = \alpha l$.

Substituting in (5) above,

$$Z_i^l = Z_o\left(\frac{Z_R + Z_o\alpha l}{Z_o + Z_R\alpha l}\right)$$

Dividing numerator and denominator of the fraction on the right by Z_R, we obtain

$$Z_i^l = Z_o \frac{1 + \frac{Z_o\alpha l}{Z_R}}{\frac{Z_o}{Z_R} + \alpha l}. \quad (6)$$

If the half wavelength cable is open-circuited (i.e. $Z_R = \infty$), (5) will reduce to

$$Z_i^l = (Z_o)\frac{(1)}{\alpha l} = \frac{Z_o}{\alpha l} \text{ and}$$

$$\alpha l = \frac{Z_o}{Z_i^l} \text{ nepers. Then}$$

$$\alpha l = \frac{Z_o}{Z_i^l} \times 8.69 \text{ db, where } Z_i^l \text{ is}$$

resistive and is measured directly on the R_p dial of the 250-A RX Meter.

CONCLUSION

It will be observed that all the measurements described in this article are simple and direct, without involved computation and corrections. In all cases, relatively short length pieces of cable are used and measurements are made directly at the RX Meter terminals without the use of coaxial connectors. It may be of interest to note that a balanced line can be treated in the same fashion as coaxial lines when a "balun" or similar device is used in connecting it to the RX Meter.

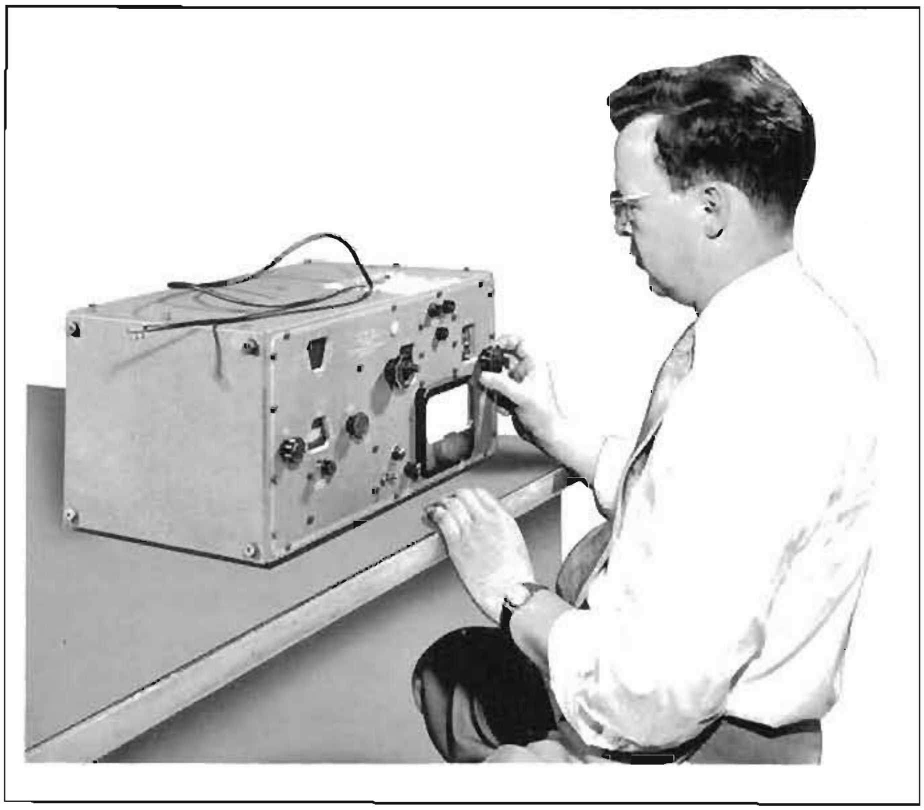

Figure 2. The author, measuring the characteristic impedance of a short length of RG-58/U cable on the RX Meter.

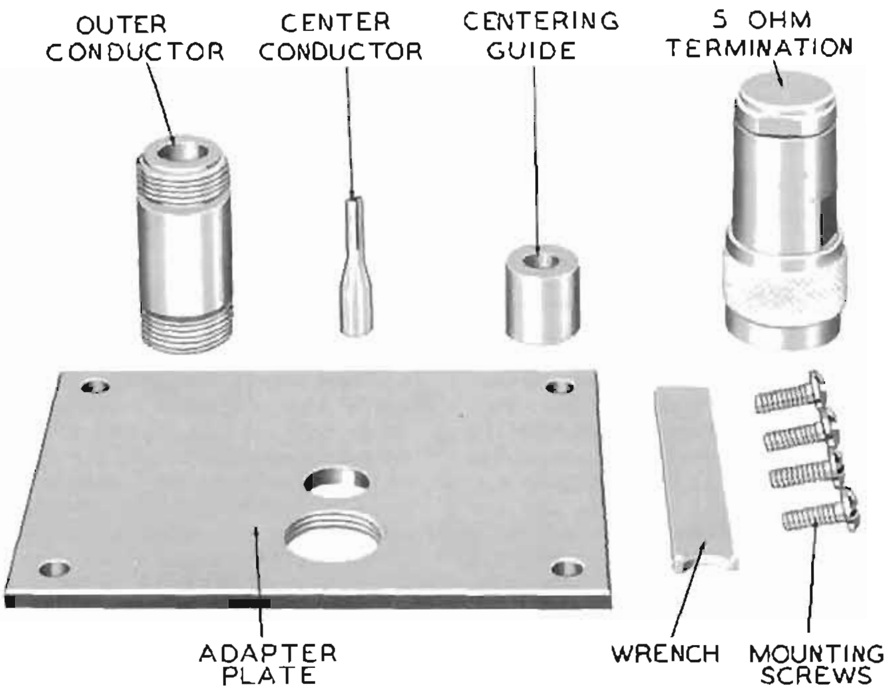

Figure 1. Components of the Co-ax Adapter Kit Type 515-A. The kit is supplied in a convenient wooden storage stand, not shown.

A Coaxial Adapter For The RX Meter

C. G. GORSS, *Development Engineer*

Soon after the RX Meter Type 250-A made its appearance in the field it became apparent that, in addition to measurement of components on the standard binding posts, many of the applications in which this new instrument was being utilized involved the use of coaxial cables and fittings. The evident need for some convenient means of coupling such components, fitted with standard coaxial connectors, to the RX Meter measuring terminals resulted in the design of a special adapter unit. This unit which, together with the necessary accessories, is now available to RX Meter owners in kit form, is designated as the "Co-ax Adapter Type 515-A."

DESIGN DETAILS

The adapter itself consists of two separate elements; a cylindrical outer conductor about 1 1/2 inches in length, the base of which is grounded to the terminal plate when mounted on the RX Meter, and a tapered center conductor which is fastened to the HI binding post stud. To mount the unit, an adapter plate, supplied with the kit, is first fastened to the terminal plate of the RX Meter. This plate has a large-diameter tapped hole which centers around the HI post stud. After the HI post clamping nut has been removed and the center conductor has been screwed firmly over the stud, the outer conductor is turned down into this hole until its base makes contact with the terminal plate. The open end of the adapter then forms a standard Type N female connector.

When the adapter is installed, ordinary measurements requiring the use of the binding posts are easily made merely by unscrewing the outer and inner conductors of the adapter and replacing the binding post nuts. The adapter plate in no way interferes with such measurements.

Along with the adapter and adapter plate, the kit includes a wrench for removing the ground binding post base nut, a centering guide for accurate positioning of the outer conductor, four screws for fastening the adapter plate, and a special 50-ohm coaxial termination.

The unit is designed to have a constant characteristic impedance of 50 ohms. All surfaces are rhodium-plated to insure good contact and to match the plating used on the RX Meter terminal plate.

The termination, which is used in obtaining preliminary balance of the RX Meter bridge, is equipped with a Type N male connector for direct connections to the adapter. Like the adapter, it is actually a short section of transmission line. Its center conductor, however, is actually a special high frequency resistor. The termination produces a voltage standing wave ratio of less than 1.10 up to 800 mc.

APPLICATION

When the adapter is installed a coaxial element may be attached and measured at any selected frequency after two minor preliminary adjustments of the RX Meter bridge circuit controls have been made to establish the correct "zero balance" condition. The first adjustment, made with nothing attached to the adapter and with the R_p dial set at ∞, consists of obtaining a null indication by alternate adjustment of the ZERO BALANCE R controls and the C_p control. This establishes the correct "resistance zero" of the circuit.

The second zero balance adjustment is made with the 50-ohm termination mounted in place on the adapter. This time a null indication is obtained by means of the R_p and ZERO BALANCE C controls, with the C_p dial at 0. This establishes the correct "reactance zero." Actually, it has the effect of adjusting to the proper value the characteristic impedance of a short internal connecting section (several centimeters in length) between the RX Meter binding post and the physical point on the bridge circuit at which the measurement is actually made. Since the characteristic impedance of this section is not, in itself, 50 ohms, that value must be established synthetically by proper adjustment of the ratio L/C. This is automatically accomplished by the setting of the ZERO BALANCE C control described above.

Although the co-ax adapter is useful in facilitating measurement of the characteristics of cables and other coaxial elements, probably its most important application is in providing for the measurement of impedances remote from the RX Meter terminals. When the proper techniques are used, it is possible to measure an impedance at the end of a section of coax line with the same accuracy with which it can be measured directly at the RX Meter terminals. Such measurements may, if desired, be made with random-length sections of 50-ohm coax, in which case the results must be transformed by means of a Smith Chart[2] or the familiar transmission line equations in order to obtain the actual impedance of the unknown. In this case, the short internal connection, mentioned above, between the binding post and bridge, becomes part of the transmission line and its effective length

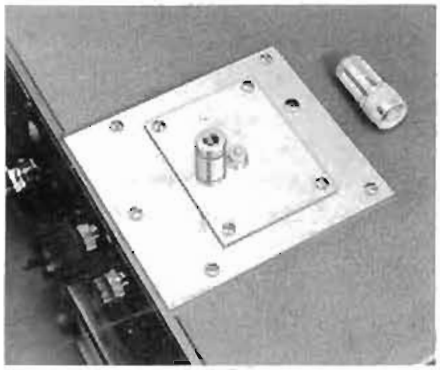

Figure 2. The Co-ax Adapter, mounted on an RX Meter. The 50-ohm termination shown beside the terminal plate is used in obtaining preliminary balance of the RX Meter bridge circuit.

must be determined and added to the physical length of the line itself in computing the results. A somewhat simpler method, when the measuring frequency is high enough for the cable length involved to be practical, is the use of a resonant half-wave section which will transfer almost exactly the value of an impedance connected at one end to the RX Meter measuring terminals connected at the other. In this case, the impedance of the section itself is not a factor, and the only correction necessary is for actual loss in the line which, frequently, is negligible.

D.R. Crosby and C.H. Pennypacker, "Radio Frequency Resistors as Uniform Transmission Lines," Proc. of the IRE, Feb., 1946.

[2] P. W. Smith, "Transmission Line Calculator," Electronics, Jan., 1939.

A NOTE FROM THE EDITOR

This is a very brief note. In fact, our authors were so enthusiastic that they left the editor only enough space to note, with considerable satisfaction, that over 22,000 people asked for the Notebook.

Univerter Signal-To-Noise Ratio

Frank G. Marble, *Sales Manager*

A simple method for extending the frequency range of a signal generator is the use of a broadband frequency converter with a gain of one. Such a converter is the Univerter Type 207-A. This instrument consists of a broadband mixer, a local oscillator and a broadband amplifier with an output impedance of fifty ohms. The output frequency differs from that of the signal generator by the frequency of the local oscillator of the Univerter. The useful frequency range is limited by the upper and lower limits of the mixer and the broadband amplifier (0.1 to 55 mc).

The mixer of the Univerter has a small amount of inherent amplitude modulation resulting from random noise generated in its input impedance. Since the pass band of the Univerter covers a broad frequency range, little selective rejection of this noise occurs. The effect is not noticeable for signal levels above approximately 2.5 microvolts. Below this level, signal level decreases of a given value will cause reductions of receiver output of a smaller value. The use of a 20 db pad, such as the 509-A, at the output of the Univerter will permit use of the signal generator and Univerter down to approximately 0.25 microvolts, since the noise is attenuated directly and the signal levels from the signal generator can be increased by 20 db to compensate for this pad. Additional attenuation can be used for lower outputs.

To allow our customers to make use of the Signal Generator (Types 202-B and 202-C) with the Univerter Type 207-A at these very low levels the Adapter Type 509-A (53 ohms unbalanced to 53 ohms unbalanced with 20 db attenuation) is now supplied with each Univerter at no increase in price.

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A Versatile Instrument - THE Q METER

What Can Be Done With a Q Meter - Besides Measure Q

LAWRENCE O. COOK, *Quality Control Engineer*

Lawrence Cook become associated with BRC in 1935, shortly after the company's founding. He started as an Electrical Inspector at the time the original Q Meter was being produced. Since then his activities have followed closely the development of subsequent Q Meter Models. Mr. Cook was graduated from Bliss Electrical School in Washington D.C. and was employed by Sparks-Withington Co. prior to BRC. He is an associate member of I.R.E. and a member of the Radio Club of America.

INTRODUCTION

A Q Meter contains in one instrument a frequency-calibrated RF oscillator, a system for RF voltage injection and measurement, a vacuum tube voltmeter, a calibrated variable capacitor, and four terminals for the connection of components to be measured. All of these elements (but two terminals) are employed in measuring the Q of a coil, the normal connections being shown in Figure 1. But these elements, in part or in combination, may be employed in the performance of various other functions. It is the purpose of this article to describe some of these applications of the versatile Q Meter.

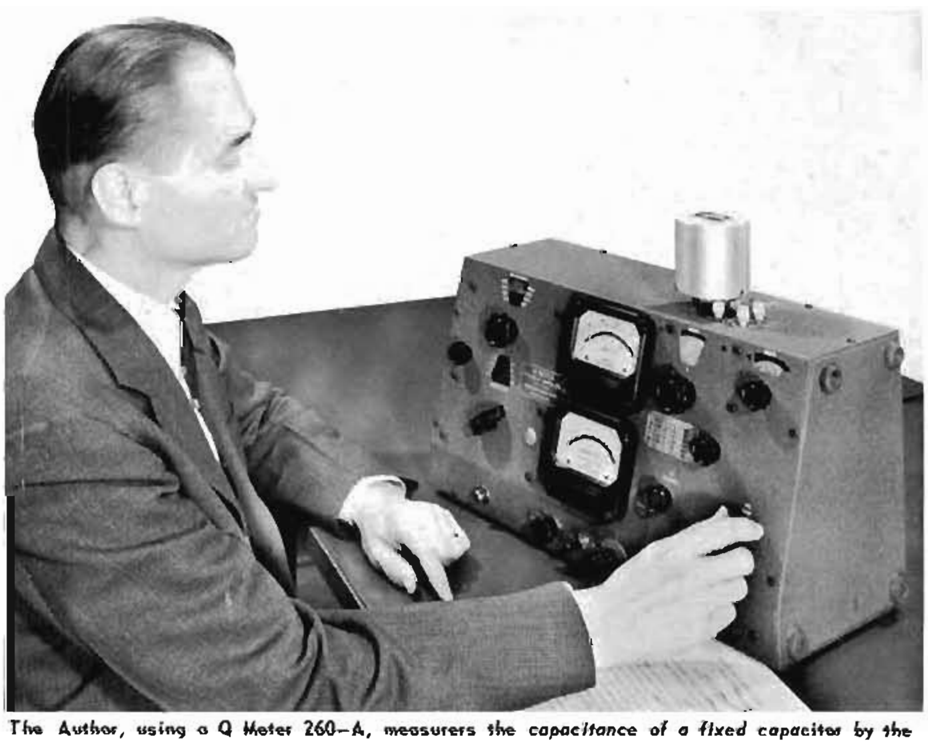
CAPACITANCE MEASUREMENT

For capacitance measurements convenient use can be readily made of the calibrated variable capacitor just mentioned. Let us assume that a coil is connected to the COIL posts and resonated to the oscillator frequency (see Figure 2). Note the reading of the calibrated variable capacitor dial C_c, calling this C_1. If an unknown capacitor C_x is now connected to the Q Meter CAP terminals (*), it will be placed in parallel with the variable capacitor. Next, if the Q Meter is re-resonated by adjustment of the variable capacitor to a lower value, calling the reading C_2, the quantity $C_1 - C_2$ represents the effective value of the unknown capacitor C_x.

The above measurement is preferably made at a low frequency, e.g. 1 mc., thus avoiding lead inductance effects. However, by employing a higher frequency, e.g. 50 mc. or 100 mc., the rise in effective capacitance of the unknown as a result of lead inductance may be readily observed. Fractional-inch variations in lead length will be found to have a pronounced effect on the measured quantity $C_1 - C_2$.

The parallel type of measurement just described can be extended in capacitance range as follows: first measure a capacitor by the method described, calling the measured capacitance C_a. Then, to measure a

(*)These terminals are designated CAP on Q Meters Type 190–A and 260–A and COND on 160–A and 170–A.



The Author, using a Q Meter 260–A, measurers the capacitance of a fixed capacitor by the Parallel Method.

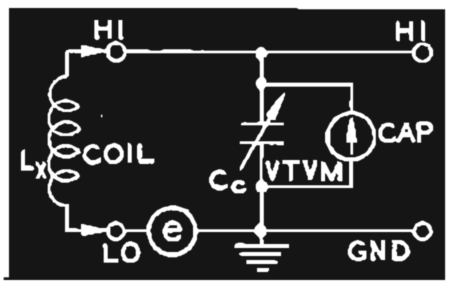


Figure 1. Q Meter Simplified Diagram. e--Calibrated injection voltage derived from RF oscillator, C_c--Calibrated variable capacitor, VTVM -- Vacuum-tube-voltmeter measuring voltage across C_c, Terminals for connection of components to be measured are indicated as Hi, Lo, GND, and L_x -- coil being measured.

YOU WILL FIND. . .

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second but larger capacitor C_x, use the previous method but connect C_a in parallel with the Q Meter CAP terminals (*) when determining C_1; then disconnect C_a, connect C_x, and readjust the calibrated variable capacitor to determine C_2.

Then $C_x = C_1 - C_2 + C_a$.

Alternatively, a series type of measurement may be employed to extend the range upward by approximately 10:1 (see Figure 3). The unknown capacitor is connected in series with

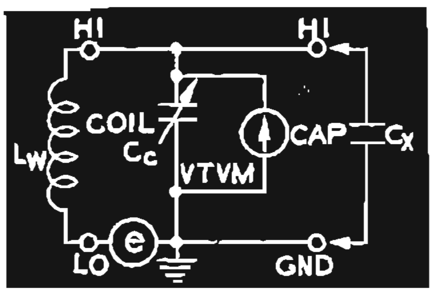


Figure 2. Capacitance Measurement, Parallel Method. L_w -- workcoil (such as Type 103-A or 590-A), C_x -- unknown capacitor (or dielectric sample) to be measured.

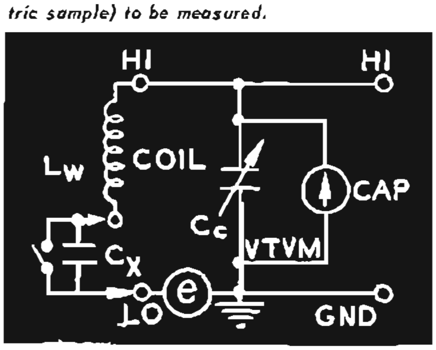


Figure 3. Capacitance Measurement, Series Method.

the measuring circuit between the low potential side of the coil and the Q Meter LO terminal; the circuit is resonated, preferably at a high reading on the calibrated variable capacitor, this reading being called C_2. The unknown capacitor is then removed from the circuit, or preferably left in position but short-circuited to minimize changes in circuit inductance, the variable capacitor is readjusted for resonance, the reading being called C_1. The effective value of the unknown capacitor is then

$$C_x = \frac{C_1 \; C_2}{C_2 - C_1}$$

MEASUREMENT OF DIELECTRICS

Dielectric samples (insulating material samples) for electrical measurement purposes are usually in sheet form and of 1/16 inch to 1/4 inch in thickness. If, to the sample surfaces, are attached metal foil electrodes, a capacitor is formed whose electrical properties depend largely upon the sample material.

The metal foil electrodes are usually of lead or aluminum, of 0.00075 inch to 0.0015 inch thickness, and are attached to the sample surfaces by means of a thin film of vaseline (petroleum) jelly). The foil area is usually adjusted to provide a sample capacitance of 20 to 80 $\mu\mu$f depending upon the Q or power factor of the material to be measured and the type of Q Meter to be used. Sample discs of 2 inch diameter with foils extending to the sample edges are frequently used.

A parallel measurement is used on the Q Meter (see Figure 2). An inductor is connected to the Q Meter COIL terminals and resonated at the measurement frequency. The Q and tuning capacitance, Q_1 and C_1, are read. The sample foils are then connected to the Q Meter CAP terminals (*), the tuning capacitor C_c is readjusted for resonance, the Q and tuning capacitance, Q_2 and C_2, are read.

Neglecting edge and stray effects the following formulas apply.

The Q of the sample is

$$Q_x = \frac{(C_1 - C_2) \; Q_1 \; Q_2}{C_1 \; (Q_1 - Q_2)}$$

The power factor of the sample (for values less than 10%) is

$$\text{P. F. } (\%) = \frac{100}{Q_x} = \frac{100 \; C_1 \; (Q_1 - Q_2)}{(C_1 - C_2) \; Q_1 Q_2}.$$

The dielectric constant of the sample is

$$\epsilon = \frac{C_1 - C_2}{C_a} = \frac{4.45 \; (C_1 - C_2) \; t}{A}$$

where C_a = Calculated capacitance (micro-micro-farads) of equivalent capacitor with dielectric replaced by air.

t = Dielectric material thickness in inches.

A = Dielectric active area between electrodes in square inches.

C_1 and C_2 are in micro-micro-farads.

At frequencies above 10 mc the measurement method just described is sometimes unsatisfactory because of the effects of lead inductance, foil resistance, etc. For such conditions a somewhat different measuring technique is used with a specially designed sample holder which provides a constant circuit inductance for the "sample in" and "sample out" readings. This holder also permits, for certain measurements, elimination of the foil electrodes and their associated resistance. The General Radio Company Dielectric Sample Holder Type 1690-A is suitable for this work. Fabrication drawings of a mounting plate for attaching this Dielectric Sample Holder to Q Meters Types 160-A, 170-A, 190-A and 260-A are available from Boonton Radio Corporation.

Oils and other fluids require a cell or container with suitable electrodes between which the fluid to be measured may be placed.

INDUCTANCE

The calibrated oscillator frequency and calibrated variable capacitance scales of the Q Meter provide a convenient means of determining coil inductance. For this purpose the unknown coil is connected to the Q Meter COIL terminals and resonated as for reading Q (see Figure 1). The frequency, f, and tuning capacitance, C_1, are read and inserted in the following formula for inductance.

$$L_s = \frac{1}{\omega^2 \; C_1}$$

where L is in henries
ω is 2π times the frequency in cycles
C is in farads
or,

$$L_s = \frac{2.53 \times 10^4}{f^2 \; C_1}$$

where L is in microhenries
f is in megacyles
C is in micro-micro-farads

In either instance the inductance value obtained is the effective inductance of the coil including the effect of distributed capacitance.

All Q Meters Type 260-A and 160-A of late manufacture, include a calibrated inductance scale on the calibrated variable capacitor dial. (An LC Dial Kit Type 560-A is available for adding this feature to early Q Meters Type 160-A). A chart provided on the Q Meter front panel permits use of any one of six inductance ranges by selection of the proper oscillator frequency. The inductance scale then reads directly in terms of effective inductance.

If the distributed capacitance of the inductor is known, the true inductance can be readily determined. With the variable capacitor dial adjusted to the effective inductance value, note the corresponding reading on the capacitance scale in micro-micro-farads immediately above. Add the distributed capacitance to this reading; adjust the dial to indicate the sum just obtained. Although the measuring circuit is now detuned from resonance, the true inductance of the coil may be read on the inductance scale immediately below.

DISTRIBUTED CAPACITANCE

All coils have distributed capacitance and a measurement of this quantity is often required. The measurement may be made on the Q Meter by the following method:

Connect the coil to be measured to the Q Meter COIL terminals(Figure 1). Resonate the Q Meter, calling the oscillator frequency dial reading f_1 and the calibrated variable capacitor dial reading C_1 (C_1 should be preferably in the lower part of the scale).

Readjust the oscillator to a considerably lower frequency f_2, equal to f_1/n. Restore resonance by readjusting the variable capacitor, calling the new reading C_2. The distributed capacitance is then

$$C_d = \frac{C_2 - n^2 C_1}{n^2 - 1}$$

If f_2 is made exactly equal to $f_1/2$, then

$$C_d = \frac{C_2 - 4\ C_1}{3}$$

An average of several measurements employing different values of C_1 and C_2 will improve the accuracy of the results.

SELF-RESONANCE

The self-resonant frequency of an inductor, i.e. the resonant frequency with nothing connected externally to the inductor terminals, can be readily determined with the Q Meter. Looking into the terminals of the inductor, reactance conditions vs frequency will be seen as in Table 1, columns 2 and 3. The Q Meter distinguishes readily between conditions a, b, c, thus providing an accurate determination of the self-resonant frequency, as will be explained in detail with the aid of column 4.

In making the measurement the first step is the determination of leads required to connect the unknown inductor to the Q Meter CAP terminals (1). These leads are then permanently connected to the CAP terminals and the inductor is disconnected. This procedure minimizes the effects of lead capacitance on the self-resonant frequency.

Next, the Q Meter is resonated with a work or accessory coil (preferably shielded, such as the Types 103-A or 590-A) connected to the Q Meter COIL terminals (see Figure 4). The frequency chosen should be in the region

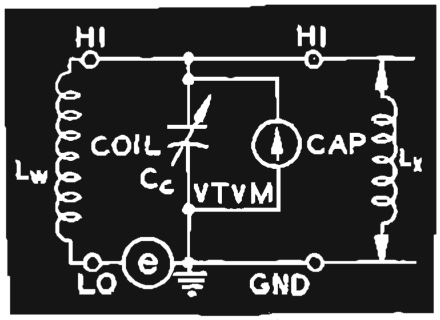


Figure 4. Inductor or Choke Measurement, Parallel Method.

of the estimated self-resonant frequency of the inductor to be tested. Now connect the unknown inductor L_X to the CAP terminal leads previously established. Re-resonate the Q meter by means of the capacitance dial C_c, noting the direction of movement of this dial as referred to the original setting.

Unless the unknown inductor is found to be non-reactive at the measurement frequency first chosen, the test procedure is now to be repeated at a somewhat higher or lower frequency as determined by reference to Table 1, columns 2 and 4. Successive frequency adjustments will eventually achieve the desired condition where the capacitance dial reading for resonance is unchanged as a result of connecting the unknown inductor to the Q Meter CAP terminal leads. The unknown inductor is then non-reactive and self-resonant at the frequency indicated by the oscillator dial.

CHOKE COILS

A choke coil, to provide proper isolation characteristics, must exhibit a high impedance throughout its operating frequency range. Failure to meet this requirement may result in low operating efficiency, frequency error in calibrated circuits, etc.

The Q Meter provides an ideal means for the measurement of choke coil characteristics. A work coil, preferably shielded, is connected to the Q Meter COIL terminals (Figure 4). Leads of short length may be used, if required, to connect the unknown choke coil to the Q Meter; these leads are now to be attached to the Q Meter CAP terminals (1) but the choke coil is to be disconnected. The work coil is resonated at the frequency of measurement, called f, the Q reading being called Q_1, and the calibrated variable capacitor reading being called C_1. Temporarily remove the Q Meter CAP terminal leads, if used, and note the increase required in the calibrated variable capacitor reading for resonance; call the increase C_L; re-connect the leads. Next the unknown choke coil L_X is connected to the Q Meter CAP terminals leads and the calibrated variable capacitor is readjusted for resonance. Call the Q reading Q_2 and the capacitor reading C_2.

The above procedure should be repeated at other frequencies within the

TABLE 1

INDUCTOR SELF-RESONANCE DATA

(1) Condition	(2) If Frequency Is	(3) Inductor Reactance Will Be	(4) Q Meter Capacitance Dial Test Reading
a	Below self-resonance	Inductive	Increases
b	At self-resonance	Non-reactive	No change
c	Above self-resonance	Capacitive	Decreases

operating range of the choke coil (C_L may be assumed to be constant and need not be re-checked when the frequency is changed).

The effective parallel resistance, R_p, and effective parallel reactance, X_p, of the choke are

$$R_p = \frac{Q_1 Q_2}{\omega(C_1 + C_L)(Q_1 - Q_2)}$$

and

$$X_p = \frac{1}{\omega(C_2 - C_1)}$$

where R_p, X_p are in ohms
$\omega = 2\pi f$ (cycles)
C is in farads.

$$\text{Or, } R_p = \frac{1.59\text{x}10^5 \text{ x } Q_1 Q_2}{f(C_1 + C_L)(Q_1 - Q_2)}$$

$$\text{and } X_p = \frac{1.59 \times 10^5}{f(C_2 - C_1)}$$

where R_p, X_p are in ohms
f is in megacycles
C is in micro-micro-farads.

NOTE: The sign of the quantity (C_2-C_1) indicates the type of effective reactance. A positive quantity indicates inductive reactance. A negative quantity indicates capacitive reactance.

MUTUAL INDUCTANCE AND CRITICAL COUPLING

(a) The mutual inductance and coefficient of coupling of RF coils may be measured on the Q Meter at high frequencies by the familiar method often employed at low frequencies with audio frequency bridges.

This method is used for large coupling coefficients, i.e. 0.5 or greater. Four measurements are made (on the Q Meter COIL terminals) at or near the operating frequency and preferably at one frequency.

Measure L_1 and L_2 separately. Then measure L_a (mutual aiding) with L_1 and L_2 connected series aiding, and L_b (mutual bucking) with L_1 and L_2 connected series bucking (Figure 5).

The mutual inductance then is

$$M = \frac{L_a - L_b}{4}.$$

The coefficient of coupling is

$$K = \frac{L_a - L_b}{4\sqrt{L_1 L_2}} = \frac{M}{\sqrt{L_1 L_2}}.$$

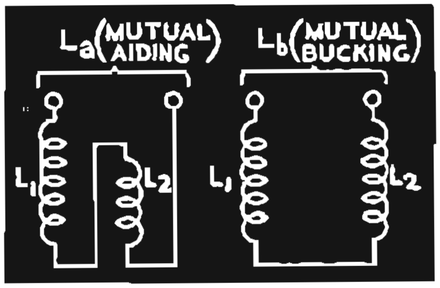


Figure 5. Mutual Inductance Connections, L_1 L_2 -- first and second coupled coils, respectively.

If the measurements are made at one frequency

$$K = \frac{\left(\frac{1}{C_a} - \frac{1}{C_b}\right)\sqrt{C_1 C_2}}{4}$$

where C = Q Meter tuning capacitance necessary for resonance with mutual aiding connection, mutual bucking connection, and single inductors respectively.

(b) With RF coils most commonly used the critical coefficient of coupling (i.e., the condition where the resistance that the secondary circuit at resonance couples into the primary circuit is equal to the resistance of the primary circuit) occurs at a low value of coupling coefficient. Design-wise, the critical coupling condition is important because it yields the maximum value of secondary current and it may be readily determined as follows.

Connect one of the two coils to the Q Meter COIL terminals with the second coil open-circuited (see Figure 6) and adjust the Q Meter for resonance. Read Q_1. Now complete the secondary circuit and, by means of its trimmer, resonate it to the same frequency as indicated by a minimum Q reading, Q_2.

If Q_2/Q_1 equals 0.5 the coils are critically coupled; if greater than 0.5 the coupling is less than critical, and if less than 0.5 the coupling is greater than critical.

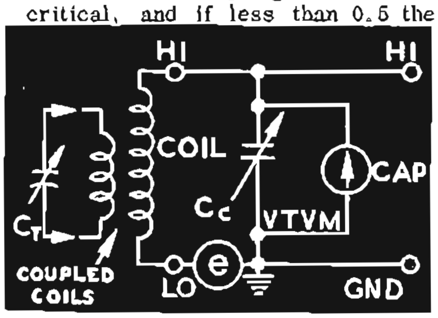


Figure 6. Critical Coupling. C_t -- Secondary coil Trimmer.

These results will be with respect to the coils only. If it is desired to include the effects of tube and circuit loading, resistors duplicating these loading effects should be added to the coils before making the measurements.

GAIN OF COUPLED COILS

The Q Meter is essentially a gain measuring device, i.e., Q is measured by determining the ratio of two voltages. This instrument is thus readily adaptable to the gain measurement of coupled coils within its range.

For example, a transformer used to couple a low-impedance loop antenna to a receiver input may be measured. Referring to Figure 7, the transformer primary circuit including the loop (a coil may be used to simulate the loop) is connected to the Q Meter GND and LO terminals. The transformer secondary is connected to the Q Meter GND and HI terminals. The Q Meter injection voltage thus excites the transformer primary circuit and the transformer secondary voltage is fed to the Q voltmeter. Adjustment of the calibrated variable capacitor C_c for resonance will now yield a "Q reading" which is numerically equivalent to the transformer circuit gain. The Q scale reads gain directly when the "Multiply Q By" meter is set to x1.

By completion of the proper connections the above method can obviously be extended to include stage gain measurements.

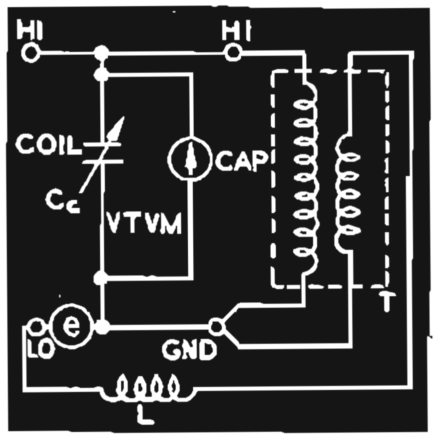


Figure 7. Transformer Gain Measurement.

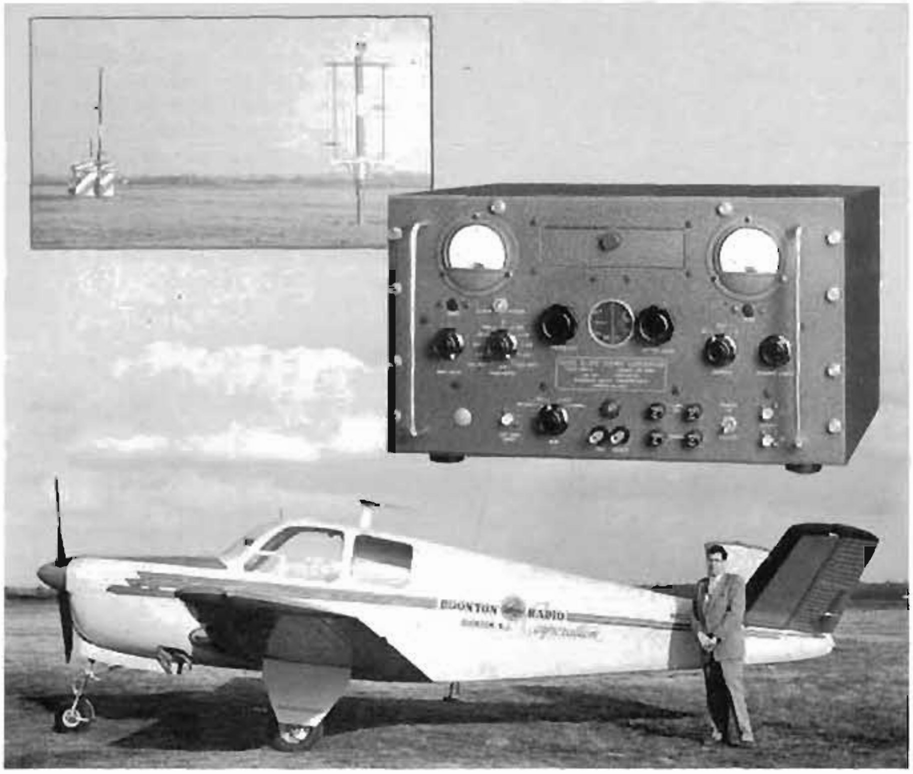
SUMMARY

We have described some of the "extra-curricular" uses to which a Q Meter may be put. No attempt was made to write an exhaustive article and we are sure that our many customers have devised other ways of utilizing this versatile instrument.

May we draw this conclusion: that, when an RF measurement problem is at hand, the Q Meter may do the job.

BAD WEATHER FLYING

EDSON W. BEATTY, *Chief Pilot*



The Author standing beside the BRC Beechcraft Bonanza. The aircraft is fully equipped for Instrument flight and provides a valuable source of information relevant to actual aircraft operation under all flight conditions. Inset upper left is a typical Instrument Landing System Glide Slope field installation. The BRC Glide Slope Signal Generator Type 232-A (center) provides calibrated RF signals and modulation on all Glide Slope frequencies -- thus providing manufacturers and service organizations with an instrument capable of simulating any or all signals transmitted in the Glide Slope Section of the Instrument Landing System.

Ever been delayed at the airport by weather or arrive several hours behind schedule? Most of us have, but have you noticed recently these delays are not occuring as often? We here at BRC feel that with our test equipment we are helping to contribute a part towards reducing these occurances.

Today, on board a modern high speed air transport, we give little thought of the problems brought about by the increase in speed, traffic density and type of weather now considered flyable. Just as an example, an aircraft traveling at 150 MPH with a course error of 5^o will be approximately 4 miles off course after 30 minutes time. However, at 350 MPH with the same course error for 30 minutes, the aircraft would be over 8-1/2 miles off course. Now commercial flights are being made daily with little more than part of the airport runway being visible. Experimental flights are being made under visibility conditions where great difficulty in ground or taxiing handling is encountered. The writer had an experience recently when, rolling down the runway after landing in a heavy fog, the control tower called by radio from 1/4 mile away and requested "Have you landed yet ?"

These improvements in air navigation and communications have all been brought about through electronic equipment. The primary aids now used are Omnirange for navigation, Instrument Landing System for blind landing and VHF Communications.

THE VOR SYSTEM

Omnirange, sometimes referred to as VOR (Visual, Omni-directional Range), is a recent development in radio navigation aids. The Omnirange is designed, as are other air navigation systems, to furnish directional guidance to an aircraft in space. It is the primary aid in point-to-point air navigation. The word Omni is derived from Latin "Omnis" meaning all. These stations are so named because they have an infinite number of courses, whereas, the facility they replace has only four courses.

An Omnirange station might be described as a very large wheel with 360 spokes (theoretically an infinite number) with the station being the hub. Any one of the spokes might be chosen as a guide in space. This is accomplished electronically by comparison of the phase difference between the audio modulation of two radiated radio frequency signals, the difference in phase varying with change in azimuth. The modulation on one of these signals is non-directional and has a constant phase throughout 360 degrees of azimuth. This is called the reference phase. In order to separate the two signals for comparison in the receiver and converter, a 10KC FM subcarrier is used to carry the reference signal. The phase of other signal rotates at a speed of 1800 RPM and varies in phase with azimuth. This is called the variable phase and is produced by a motor driven goniometer feeding an RF voltage into four antennae (two at a time). As the goniometer revolves, the RF voltage fed into the antennae (180 degrees apart) varies sinesoidally at the rate of 30 cps to produce a rotating field. The system is set up so at magnetic north the reference and variable signals are exactly in phase.

From a pilot's standpoint the operation is quite simple. Other than receiver, converter and antennae, there are four basic units in the aircraft. Although some manufacturers combine these units for simplicity, we shall discuss the primary type.

The pilot's controls are illustrated in Figure 1 and consists of:

1. Frequency Selector (conventional receiver control)
2. Azimuth Selector
3. Deviation Indicator (with signal strength indicating flag)
4. Sense Indicator

Further describing each unit:

1. Frequency Selector: tunes receiver to desired station frequency which is identified aurally with code (recorded voice also on some stations).
2. Azimuth Selector (Course Selector): selects the desired azimuth (or track) to control direction in space.
3. Deviation Indicator: indicates the difference between the selected azimuth and present position or,

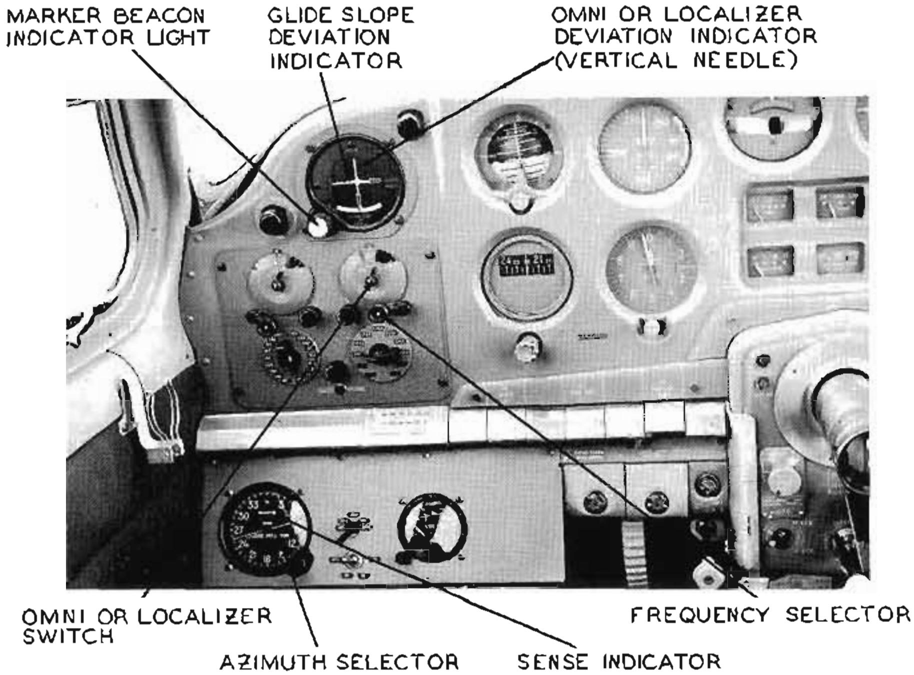


Figure 1. The Pilot's VHF Radio Controls installed in the Instrument panel of the BRC Beechcraft Bonanza.

conversely, the azimuth from a station to the aircraft's position. Two or more stations may be used to establish position. Also included is an alarm indicator which indicates when a usable signal is being received.

4. Sense Indicator (To-From Meter): determines the phase comparison to establish the quadrant (i.e. north azimuth or south azimuth).

At the present time there are 392 omni stations now in use throughout the United States.

THE ILS SYSTEM

We have discussed point-to-point air navigation and although the omni-range may be used as a landing aid it is not the primary type. The Instrument Landing System (ILS) is the more effective type and considered the most practical from a cost and operational standpoint. Its purpose is to provide a predetermined, precise path to a landing runway without visual reference to the ground.

The system employs three elements:

1. Localizer
2. Glide Slope
3. Outer and Middle Marker

To explain each element:

1. Localizer: provides the directional guidance to and down the landing runway.
2. Glide Slope: provides the altitude guidance while approaching on the Localizer.
3. Outer and Middle Markers: provide fixes or locations on Localizer and Glide Slope.

In order to describe the operation of ILS, we shall consider each element separately.

The Localizer provides the directional guidance by radiating a field pattern directly down the center line of the instrument runway. The carrier is modulated at two frequencies, 90 and 150 cps, with each modulated carrier applied to a separate antenna system. They are arranged so that while on the approach end of the instrument runway facing the antenna, the 90 cps signal predominates on the left and the 150 cps on the right. With this arrangement, an equal signal ratio of 90 to 150 cps is projected down the instrument runway and continuing off into the approach area. The equal signal zone is designed to be approximately 5° wide.

The Glide Slope provides altitude guidance while approaching on the localizer. This is accomplished in much the same manner as the localizer with the exception of the direction of equal signal zone. The carrier is modulated at two frequencies, 90 and 150 cps with each modulated carrier supplied to a separate antenna system. These systems are arranged so an equal signal zone, or tone ratio, is 2-1/2° to 3° from parallel to the earths surface and is approximately 1° wide.

The ILS markers, there are two, called outer and middle, serve as radio fixes to check progress on Localizer and Glide Slope. Both are vertically-radiated, low-power signals (always 75MC) elliptical in shape and directed so the center is directly under the localizer on-course signal. The Outer Marker is located between 4 and 7 miles from runway threshold. The carrier is modulated at 400 cps and keyed to emit continuous dashes. The middle marker is located between 1250 ft. and 3500 ft. from runway threshold. It s carrier is modulated at 1300 cps and keyed to emit alternate dots and dashes.

The equipment aboard the aircraft (other than receivers and antenna) consists of the following:

1. Frequency Selectors: (one each for Localizer and Glide Slope).
2. Deviation Indicators: (two meter movements in same instrument each with signal strength indicator alarm).
3. Marker Beacon Indicator Lights: (fixed frequency receiver).

From the pilot standpoint, the controls serve as follows:

1. Frequency Selector for Localizer (108-112MC): tunes the proper localizer which is identified aurally. Glide Slope (329.3 to 335MC) is tuned with a separate control and may only be identified by flag alarm opposite deviation indicator.
2. Deviation Indicators: provide guide for following Localizer and Glide Slope, the vertical indicator for Localizer and horizontal for Glide Slope.

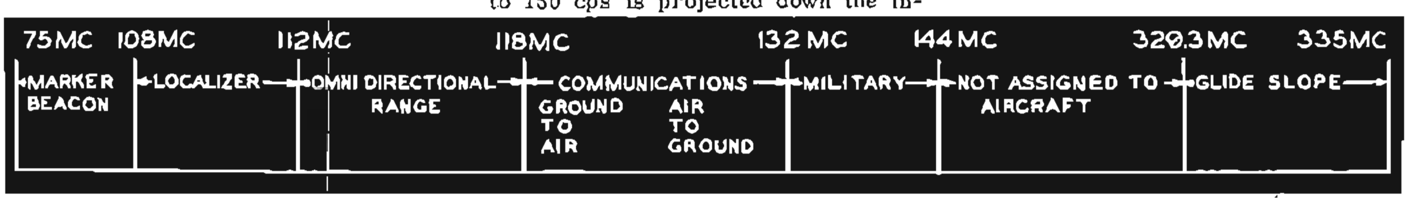


Figure 2. F.C.C. Frequency Assignments -- Aircraft Navigation and Communications.

3. Marker Beacon Lights: indicates by flashing signal which marker is being passed over, also, these signals may be identified aurally.

At the present time there are 146 ILS systems in operation in the United States not including military installations.

With the assignment of new frequencies following World War II (See Figure 2) there was tremendous need for a stable Signal Generator between 88 and 140MC. The BRC Signal Generator Type 202-B has been accepted and purchased by The Civil Aeronautics Authority and widely used throughout the industry. During the development of the Omnirange system, a phase shift was encountered in the Type 202-B which was not desirable. In 1948, a completely new Signal Generator, the Type 211-A was announced, eliminating this problem. Due to increasing demand for a crystal controlled stable Glide Slope Generator, the Type 232-A was placed on the market in 1953. All these units are approved by The Civil Aeronautics Authority as part of the necessary equipment to obtain a CAA licensed Radio Repair Station.

Already the CAA is making additions to the Omnirange and ILS systems by equipping them with DME (Distance Measuring Equipment); at present only in high traffic density areas. This together with Radar Monitoring of air traffic, bring closer the day of no-weather delays.

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CHECK YOUR Q READINGS By the Delta C Method

JAMES E. WACHTER, *Project Engineer*

There are instances not covered by the Q Standard Type 513-A in which the Q Meter user may question the Q values indicated by his instrument and, lacking a quick cross-check, believes he must content himself with questionable information. This is not necessarily so, since in many cases the "Delta C" method is both convenient and reassuring. Convenient in that the check can be conducted relatively quickly and at any frequency within the Q accuracy specification of the instrument; reassuring when it substantiates the Q Meter.

The check is based upon the equation:

$$Q_c = \frac{2C_r}{\Delta C}\sqrt{\left(\frac{V_r}{V_1}\right)^2 - 1} \qquad (1)$$

the derivation of which is too lengthy to include here.* The quantities involved in the equation are from the following Q Meter equivalent circuit including the external inductor and its associated voltage-capacitance curve:

C_r = capacitance to resonate the circuit.

V_r = voltage across the Q Meter capacitor at resonance.

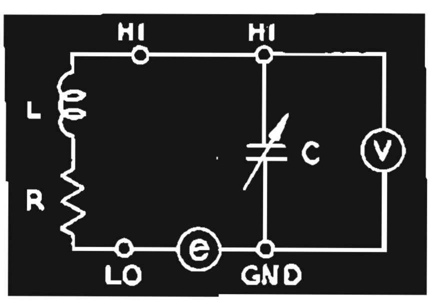

Figure 1. Equivalent Circuit of Q Meter.

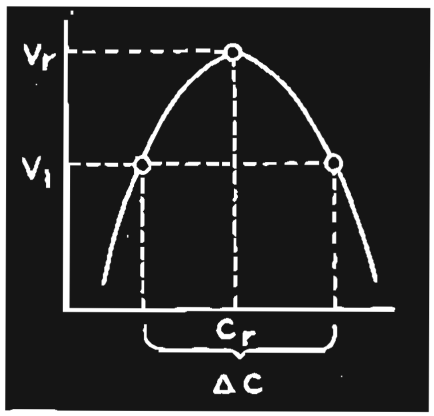

Figure 2. Capacitance Curve of the Q Meter Circuit.

V_1 = voltage across the Q Meter capacitor at a point other than resonance.

ΔC = capacitance between two points of equal voltage (V_1), one on either side of resonance.

Q_c = circuit Q = $\omega L/R$ where R includes all losses in the coil and the Q Meter circuit.

It is worthwhile to note here that since the Q- voltmeters of all BRC Q Meters are linear with respect to voltage and Q, equivalent values of indicated Q may be substituted in the ratio V_r/V_1. Equation 1 contains an approximation which is negligible when Q is greater than 100.

An easy level at which to make the ΔC measurement is at the half voltage or half Q points ($V_r/V_1 = 2.0$), in which case the preceding equation becomes

$$Q_c = 3.4641 \frac{C_r}{\Delta C} \qquad (2)$$

Another frequently used level is at the 0.707 voltage or Q points where:

$$Q_c = \frac{2\,C_r}{\Delta C} \qquad (3)$$

An outline of the procedure using equation (2) and applicable to all Q Meters manufactured by Boonton Radio Corporation is:

1. Set the Q Meter oscillator to the desired frequency.
2. Adjust the XQ control for unity.
3. Connect a shielded inductor requiring a capacitance setting near the maximum available reading for a Q Reading near full scale.
4. If Q Meters Type 160-A or 260-A are being checked, set the vernier scale to zero.
5. Resonate the circuit with the internal resonating capacitor.
   a. Record the resonating capacitance indicated on the Q capacitor dial as C_r.
   b. Record the Q at resonance as indicated on the Q voltmeter as Q_r.
6. With the internal resonating capacitor (vernier capacitor on Q Meters Type 160-A and 260-A) detune the circuit on either side of resonance to the point where the Q indicated by the Q voltmeter is equal to $Q_r/2$. Record the capacity between these two points as ΔC.
7. To avoid errors due to mechanical and electrical backlash all settings of the Q condenser should be approached with the same direction of rotation. To minimize errors in reading all settings and readings should

be made several times and then averaged.

8. Insert the values of C_r and ΔC in equation (2) and calculate Q_C.

Now, if the value of Q_C calculated in step 8 agrees with the value of Q_r recorded in step 5.b. within ± 15 percent all is well and good and the Q Meter can be assumed to be performing satisfactorily.

The method of Q Meter checking discussed here does not take into account variations in Q indication resulting from changes in loading across the measuring terminals (see " Q Meter Comparison ", Notebook 2, Summer 1954.) Difficulties (if indicated) lie elsewhere. For methods of isolating the problem see the Maintenance Section of the applicable Instruction Book.

* Hartshorn, L., and Ward, W.H., Institute of Electrical Engineers, (London, 1936), equation 6. - pp. 79, 597, 609.

EDITOR'S NOTE....

THE Q CLUB OF BRC

A BRC Employee had a bright idea back in 1942. The idea was born of the Shop and Office collection problem-- a problem that probably is common to any firm employing three or more persons. The idea provided a club for the purpose of remembering fellow workers on special occasions and to assist in arranging picnics etc.. Club dues would supply the needed funds and individual employees would no longer face the Shop and Office collections. The Company's best known instrument was the Q Meter and the new organization was befittingly named the Q Club of BRC.

Today the Q Club is a thriving organization sponsoring activities in which the great majority participate and enjoy and remembering fellow members of the BRC Family on special occasions. But we are still bothered by collections. The Q Club has gone a long way in controlling the problem but never has been fully able to meet the goal set in the original idea.

Some of the Club's inability to eradicate collections can be traced to an aspect of the Club's existence which was not fully apparent to the founders. This important aspect is the Club's healthy influence on employee relations. When a new employee starts work at BRC, he or she is soon greeted by a Q Club Representative--there is a representative for each ten employees. The Representative explains the Club's history, purpose and informs the new employee that he or she will be eligible for membership after the probation period. Fellow workers are introduced and the new employee soon has a feeling of friendship and belonging to a group. In short, employee relations are off to a good start and where friendly relations flourish, the ideas and desires for collections also flourish. The Q Club treasury always falls a little short of the good will it has created and fostered.

The good will emanating from the Q Club activities reflects itself in the BRC operations. Few people take more pains and pride in their work than members of the BRC Family. They are understanding of company production and engineering problems. On the other hand, supervisors are more understanding of the individual's problems.

Yes, Shop and Office collections have greatly benefited Boonton Radio Corporation-- they brought about the Q CLUB OF BRC!

BRC is proud of several other employee organizations. The BRC Men's Bowling League is ABC sanctioned and its weekly "Bowling Nights" regularly draw 35% of all male employees. The BRC Camera Club was recently organized and is very popular The photos on pages 5 and 6 were taken by our Camera Club President.

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Number 5

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Sweep Frequency Signal Generator Design Techniques

JOHN H. MENNIE and CHI LUNG KANG, *Development Engineers*

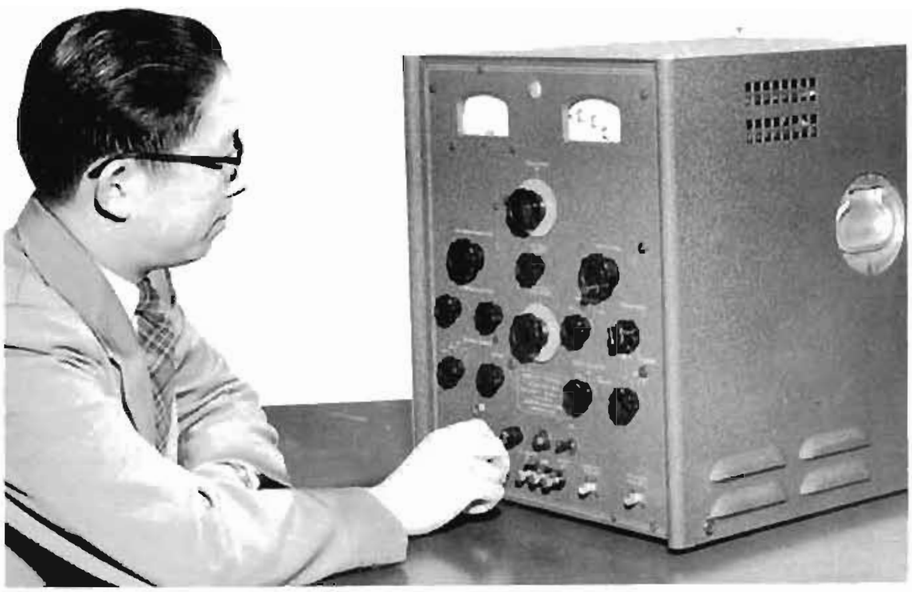

Figure 1. Dr. C. L. Kang, co-author, adjusts the Sweep Signal Generator Type 240-A.

Sweep frequency testing techniques have been used in one form or another for a number of years. Even before the cathode ray oscilloscope became available as a laboratory instrument, very slow sweep frequency devices were used in conjunction with mechanical recording systems to provide a graphic representation of frequency response. The commercial development of the cathode ray oscilloscope plus the increasing need for a rapid, accurate and convenient technique for aligning multi-stage wide-band IF amplifiers used in radar, FM and TV systems created the need for the sweeping oscillator as a laboratory and production tool.

Although many sweeping oscillators have been developed for this purpose, it has been felt that in general no one instrument incorporated all of the desirable features, such as good center frequency stability, adequate shielding, flat high-level output, precision-calibrated attenuator for operation in the low microvolt region, low sweep rate, satisfactory marking system, and ability to operate as a CW signal generator.

NATURE OF SWEEP FREQUENCY METHOD

To get a physical picture of how the sweep method works, consider a signal whose frequency is periodically varied (or swept) about a center frequency at a slow repetition rate. At a certain instant of each sweep cycle, the instantaneous frequency, i.e. the rate of change of phase angle, passes a certain value. The signal can be considered to be a single frequency equal to the instantaneous frequency over a short interval of time around that particular instant of each cycle.

Now the response of the network under test will be the steady state response at the instantaneous frequency plus a transient term. If the network is not highly frequency sensitive, i.e., low Q, its steady-state response changes gradually with frequency. Hence, the transient term will have small amplitude and, furthermore, it dies down quickly.

When the rate of frequency change is low, the time interval over which the frequency can be considered constant is appreciable and enough time will be allowed for the transients to die down. In short, in low Q circuits with low rates of frequency change, the transient term will be small and therefore the response while sweeping is a close approximation of the steady state response. This is the basis of the sweep frequency method.

SYSTEM REQUIREMENTS

Before beginning the detailed development of a sweeping signal generator it is wise to carefully consider each of several fundamental and inter-related problems inherent in all such devices. The first, of course, is to decide on the most practical means for sweeping frequency in a linear fashion with sufficient controllability and stability to satisfy both wide and narrow band requirements.

Providing adequate isolation between the generator and the load with constant amplitude high-level output and low distortion are inter-related problems made difficult by the desirability of a broadband, untuned output-buffer system.

It is often erroneously assumed that a high order of frequency stability is not essential in a sweeping device, because the frequency is constantly changing and some form of precision marker is always used to determine the exact frequency location on the display pattern. Good frequency stability has been found to be quite essential, however, to avoid jitter in the display of narrow band sweeps, for zero sweep or CW applications, and for proper identification of markers by means of an accurately calibrated center-frequency dial. A fundamental problem exists in attempting to attain good frequency stability in a sweep generator since the parameter used for sweeping is usually unstable by nature and therefore must be protected from extraneous effects

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such as temperature, voltage and current variations, magnetic fields, vibration, hysteresis, etc.

Another problem is that of providing adequate identification of all significant frequencies under test. A continuously variable marker frequency cannot be generated having the crystal-controlled accuracy often required. Although it is sometime possible to use special crystals ground to the specific frequencies required for a particular test, this procedure is impractical for a general purpose, continuously-tuned sweep generator. It is also quite difficult to set up a system of multiple crystal-controlled markers that permits identification of marker frequencies anywhere in the spectrum. The number and spacing of markers is of extreme importance as too many will be hard to identify at high frequencies and too few will not provide sufficient accuracy at low frequencies.

METHODS FOR GENERATING BROAD - BAND FREQUENCY DEVIATIONS

There are several methods commonly used for generating wide frequency deviations in sweeping oscillator circuits. These may be divided into two groups; one consisting of the mechanical methods of driving a capacitor or inductor by means of a motor, vibrator or speaker coil, and the other group consisting of all electronic variable-frequency devices such as reactance tubes, klystrons, saturable reactors, and ferroelectric capacitors. None of these is ideal in every respect, each method having definite advantages and disadvantages as listed in Table I.

IDENTIFICATION OR MARKING FREQUENCY

The fundamental problem confronted in frequency identification or marking is to indicate or mark the instant at which some varying frequency signal passes a certain value, f_k.

The two cases 1. $f_k = 0$
2. $f_k \neq 0$

are distinctly different and will be discussed separately.

1. $f_k = 0$ This is the zero beat case. The varying or swept frequency is beat with a fixed reference frequency to obtain a difference frequency as observed on an oscillo- coincidental with the reference frequency, the difference frequency becomes zero. The difference frequency as observed on an oscilloscope has a characteristic notch, when the difference is zero, which unfortunately is quite wide and jumps up and down due to random phase relations. Thus it is rather difficult to derive some triggering signal at the instant of zero beat to initiate a mark. It has been done by using the sum of squared integrals of quadrature zero beat wave forms.[1] The circuitry used is by no means simple. However, if the zero beat wave form is displayed on an oscilloscope, its center, i.e. the instant of zero beat, can be located quite accurately with the minimum amount of circuitry.

2. $f_k \neq 0$ In this case, we do not use the characteristic zero beat notch. However, the response of a frequency-sensitive network to the varying, or swept, frequency signal can be used to initiate a mark. The sweeping frequency will produce a given difference frequency twice; once as it approaches the reference frequency and again as it recedes from the reference frequency after going through zero beat. The problem is of the same general nature as detection of an FM signal, but the rate of frequency change encountered in sweep frequency technique is usually much higher and FM detection concerns chiefly change of frequency while here the actual value of the frequency is of importance.

Different methods fall into two general categories:

(a) Frequency marked by a maximum or minimum response.—The absolute level of response is therefore not important. The main difficulties with this are:

(1) Shift of envelope peak due to rate of frequency sweep.—The network used must be highly frequency sensitive in order to get a sharp peak. This very feature, however, leads to a shift of the peak when swept, which results in inaccuracy. Furthermore, this shift is not constant. It depends not only on the rate of sweep and sweep width, but also on the direction of frequency change. The process of deriving a trigger from the peak response also introduces some additional inaccuracy, especially if the peak is not very sharp.

(2) It would be difficult to design a trigger circuit which would handle a wide range of amplitudes from the frequency sensitive circuit.

(3 At a low rate of frequency change or for a narrow sweep width it may be difficult to get a high enough Q to give a sharp peak.

(4) Any envelope detector will add an additional delay in the marker display.

(b) Frequency marked by a reference amplitude of response of a frequency sensitive network:

The problems encountered are:

(1) A network of high selectivity is needed to obtain sensititivy and resolution; this feature introduces undesirable effects due to the sweeping rate.

(2) The performance or accuracy of the

TABLE I

SWEEP METHOD	ADVANTAGES	DISADVANTAGES
Mechanical Devices	High Q at all frequencies. High output possible (without buffer stage). Workable over wide range of output frequencies. Wide sweep range.	Microphonism causing frequency jitter. Non-linear sweep. Mechanical maintenance problems.
Reactance Tube	Good stability and accuracy. Non-microphonic. Linear sweep.	Limited to narrow sweep at low frequencies.
Saturable Reactor	Wide sweep range. Good stability and accuracy. Non-microphonic. Linear sweep.	Low Q at high frequencies. Susceptable to AC magnetic fields. Hysteresis effects.
Klystron Beat Method	Wide sweep range. Workable over wide range of output frequencies. Linear sweep. Non-microphonic.	Frequency jitter. Low output. Poor accuracy at low freqs.
Ferroelectric Capacitor	Non-microphonic. Linear sweep.	Excessive temperature Coeff. Low Q. Hysteresis effects.

1. D Sunstein and J. Teller, "Automatic Calibrator for Frequency Meters". Electronics, vol. 17, May 1944.

system will depend very much on the stability of the frequency sensitive network and also on the performance of the amplitude comparison circuit used to initiate the marker.

The frequency markers may form a fixed scale with a mark appearing at 5 mc or 10 mc spacings or there may be a variable marker which can be set at any specific value of frequency. For a fixed scale, it is desirable to have a choice of marker spacings to fit different sweep widths. The output indications super-imposed on the response curve may be zero beat signals between sweep and reference signals (usually referred to as "birdies"), short pulses, or intensity modulation of the scope. In this respect, the "birdie" markers have the drawback that they tend to confuse the display more than the other methods do.

The signal whose frequency is marked can either be the input or the output from the circuit under test. In the latter case, the possible wide range of variation of output amplitude poses a difficult problem for mixing. Also the R.F. output may not be easily accessible without disturbing the system under test. It is therefore generally desirable to sample the input signal for marking.

Markers are not labeled directly in frequency, and therefore a positive and convenient way to identify a mark is of importance.

SYSTEM DISCUSSION OF SWEEP SIGNAL GENERATOR TYPE 240-A

Having discussed the various aspects of sweep frequency measurements we will describe the BRC Sweep Signal Generator Type 240-A and indicate various techniques employed to satisfy the essential requirements.

Referring to Table I, it becomes apparent that the saturable reactor system offers many advantages for attaining the required wide linear sweep with good frequency stability and accuracy at frequencies below 150 mc. Although the low Q limitation of the best commercially available ferrite material suitable for high frequency use was found to be quite severe, a satisfactory oscillator was developed utilizing two high Gm triodes (type 5718) in a push-pull Colpitts circuit. Sufficient output with good waveform was thus attained under this low Q condition without exceeding plate dissipation ratings.

The R.F. Section of the Generator is shown in the photograph in Figure 2 and the block diagram in Figure 3. Center frequency tuning of the Sweep Oscillator is accomplished by means of a split stator capacitor in conjunction with a band selector switch for connecting any one of five saturable reactors. These reactors are driven by a specially shaped saw-tooth current developed and stabilized by the sweep circuitry. The problem of 60 cycle field modulation of the ferrite reactor by the power supply was solved by use of double magnetic shielding plus proper location and phasing of the power transformers. The resulting hum modulation was thus reduced to less than 0.001% of the carrier frequency.

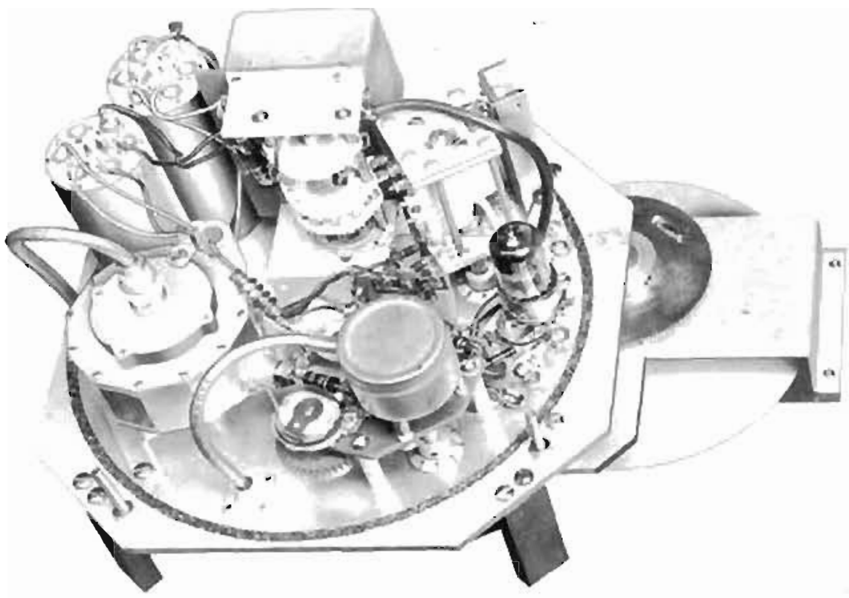

Figure 2. The Type 240—A RF Head. Switched saturable reactor windings, tuned by a variable air capacitor, control the 4.5-120 MC frequency of the oscillator. The swept frequency signal is monitored for flatness at the output of the buffer and for level at the continuously variable and the step attenuators. Special shielding, filtering, and grounding keep stray leakage at a low level.

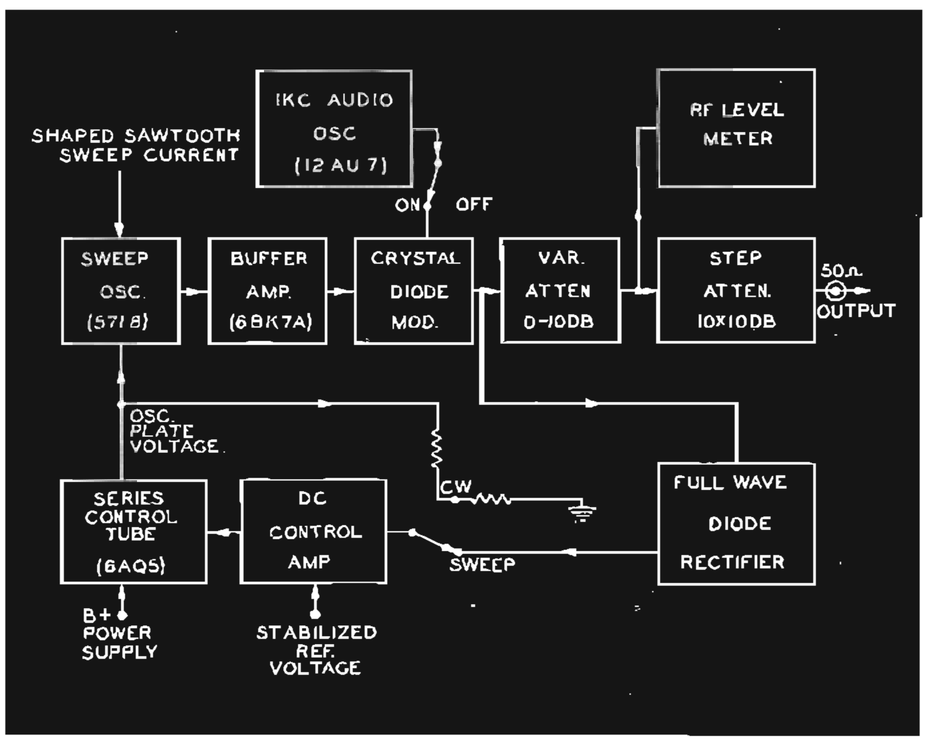

Figure 3. Block Diagram - RF Unit

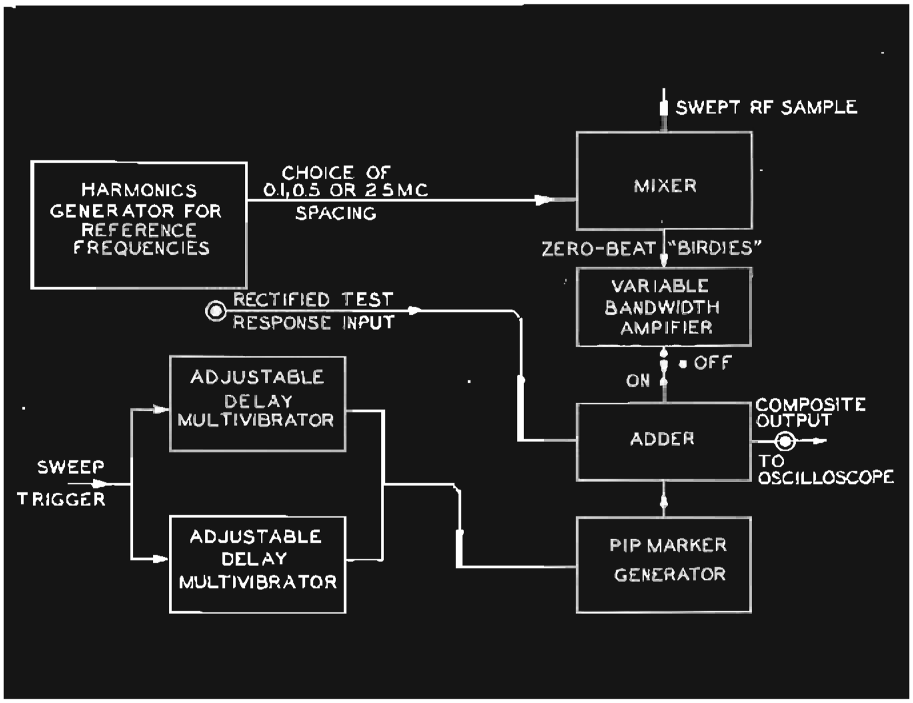

Figure 4. Block Diagram — Marker System.

The hysteresis effect in saturable reactors can be extremely serious in some designs, causing erratic frequency changes by as much as 50% each time the bias current and sweep voltage is applied to the system. It was found that this effect could be reduced to tolerable levels of less than 1% by proper choice of magnetic material in the reactor yoke assembly as well as the ferrite material used for the reactor itself. Proper design of switching circuits to eliminate transients also proved effective in reducing erratic hysteresis effects.

A grounded-grid 6BK7A buffer-amplifier follows the oscillator to reduce the effects of output impedance variations and AM modulation on oscillator frequency. Following the buffer stage is a shunt crystal diode modulator capable of providing 30% audio modulation from a low power 12AU7 RC oscillator.

The swept RF output is monitored by means of a full wave crystal diode peak-to-peak rectifier circuit that delivers to the control amplifier a DC voltage proportional to the fundamental RF level. The difference voltage between this rectifier and a stabilized DC reference voltage is amplified by a high gain DC amplifier to provide control of the oscillator plate voltage through the 6AQ5 series control tube. The resulting sweep flatness does not vary more than 0.7db as measured on a bolometer bridge. To permit audio modulation under CW conditions the AGC loop is opened up as indicated in the block diagram Figure 3.

The output attenuator system consists of a continuously variable 0-10 db pad for adjusting the RF voltage at the input to a ladder type step attenuator having 10 steps of 10 db each. The output level is thus continuously adjustable by utilizing the R.F. Level Meter to interpolate between the 10 db steps of the step attenuator. The attenuator dial is calibrated to show the RF output in terms of full scale meter reading in either volts, millivolts or microvolts as attenuation is increased.

The Sweep Signal Generator Type 240-A has a linear frequency sweep of triangular wave form. The Sweep Width Control changes the frequency excursion symmetrically above and below the center frequency which remains constant at the value set on the continuously tuned dial. In order to extend the use of the signal generator to more frequency-sensitive networks, the sweep rate is made variable from 70 cps down to 20 cps.

A variable-frequency multivibrator produces a square wave which is integrated to give a triangular wave. This triangular voltage waveform is made available through the sweep circuit amplifier for producing deflection on the oscilloscope horizontal axis. The same triangular voltage is amplified and shaped by three diodes to develop the required current waveform necessary to obtain a linear frequency sweep. During the decreasing frequency portion of the sweep, the r.f. oscillator output is reduced to zero to provide a zero level reference line. Provisions are also made so that the sweep circuit can be driven from an external source.

For frequency marking, the zero beat type marker system was adopted not only because of circuit simplicity, but also because of its good accuracy. (See Figure 4.) The problem of avoiding confusion of display due to the presence of too many "birdie" (zero beat type) markers is uniquely solved by providing two movable pips (short rectangular pulses) which can be set anywhere on the pattern in reference to the birdie markers which can then be turned off leaving only two clean pip markers.

The harmonic generator generates a fence of crystal-controlled reference frequencies with a choice of spacings; 2.5 mc., 0.5 mc or 0.1 mc. The sweep signal beats with the reference fence, giving birdie markers at the same spacing as the reference fence. The swept signal sample is taken at constant level from the buffer stage following the oscillator.

The movable pip markers are generated by a monostable multivibrator which is triggered by two adjustable delay multivibrators. The two delay circuits are themselves triggered by a pulse from the multivibrator in the sweep circuit. The time delays, and hence the pip marker positions, therefore, both have the begining of each sweep as reference. The calibration of the CW frequency to an accuracy of ±1% provides a satisfactory way of identifying the frequency markers.

CONCLUSION

A sweep frequency signal generator has been developed possessing many desirable features heretofore unavailable in a commercial instrument of this type. These include a high order of frequency stability, continuous center frequency adjustment from 4.5 to 120 mc, wide range of sweep width in conjunction with variable sweep rate, good linearity and constant amplitude high level output with good isolation between oscillator and load.

Additional features include ability to operate as a well-shielded CW signal generator with AM modulation and precision output voltage calibration down to 1.0 microvolt. Last, but not least, is a self-contained marker system providing a versatile display of crystal-controlled beat type markers and two movable pip markers, thus offering the unique combination of very accurate marking, a clean display, and easy identification.

THE AUTHORS

Chi Lung Kang was graduated from Chiaotung University in Shanghai in 1945 with a BS degree in ME and was awarded degrees of MS in ME, MS in EE, and Ph D in EE in 1948, 1949 and 1951 respectively from the University of Illinois. Dr. Kang joined Boonton Radio Corporation in 1951 as a Development Engineer. He has been active in analysis and experimental work in connection with Glide Slope Generators, the Q Standard and the Sweep Signal Generator Type 240-A and other technical development work. He was particularly concerned with the frequency identification system used in the Sweep Signal Generator described here.

For biographical material on John H. Mennie see page 4 of the Summer 1954 issue Number 2 of The Notebook. Mr. Mennie was particularly concerned with sweep circuits and the RF section of the Sweep Signal Generator Type 240-A.

Audio Frequency Measurements Using The Coupling Unit Type 564-A

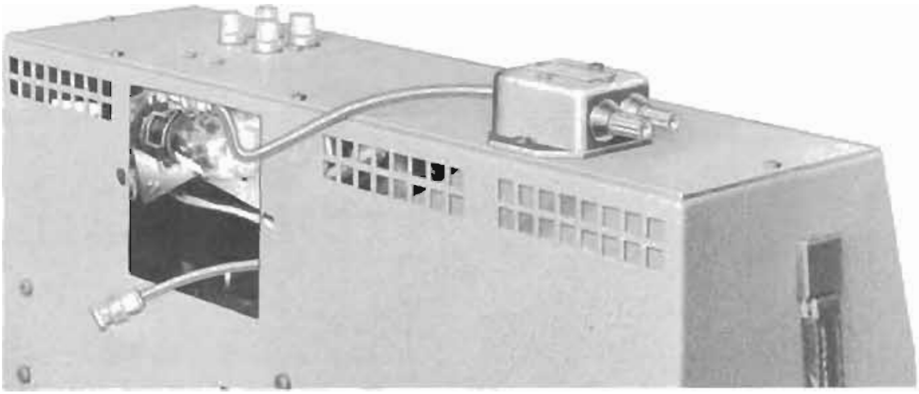

Figure 1. The Coupling Unit Type 564–A mounted on a Q Meter Type 260–A.

The voltage injection and Q-measuring circuits of the Q Meters Type 160-A and Type 260-A will perform satisfactorily at frequencies well below the lower limit of 50 kc provided by the self-contained oscillator.

Since the impedance of the thermocouple circuit is quite low (0.3 ohms in the Q Meter Type 260-A), it is often inconvenient to locate a source of audio-frequency voltage having a low enough output impedance to supply the current required to drive the Q Meter injection circuit.

To overcome this restriction BRC now offers the Coupling Unit Type 564-A which makes feasible Q measurements in the frequency range of 1 kc to 50 kc. This unit contains an impedance-matching transformer whose response varies less than 2 db over the above frequency range.

When working into the 0.3 ohm impedance of the Q Meter Type 260-A injection circuit, it presents an impedance of 500 ohms to the output of the auxiliary oscillator. Any suitable commercial audio oscillator having a variable output of up to 22 volts may be used.

The coupling transformer is mounted in a steel housing which may be fastened directly to the top of the Q Meter cabinet. Two binding posts, one insulated from ground (red insulator) and one grounded inside the housing (black insulator), provide for the connection of the auxiliary oscillator output. A 12 inch coaxial cable with a UG-88/U plug serves to connect the transformer secondary to the Q Meter injection circuit.

A. Mounting

The Coupling Unit may be firmly mounted by fastening it to the top of the Q Meter cabinet. This may be done in the following manner. Facing the Q Meter, remove the left-hand Phillips screw from the row at the top rear of the cabinet. Using the longer screw supplied with the Coupling Unit, fasten the unit to the Q Meter at this point by means of the drilled base flange, as shown in Fig. I.

B. Connection

1. Set the frequency range indicator midway between any two ranges (i.e. between detents).
2. Remove the small panel on the rear of the Q Meter cabinet.
3. Disconnect the local oscillator by removing the BNC type plug from the injection circuit receptacle, located near the top of the opening.
4. Attach the Coupling Unit coaxial connecting cable to the injection circuit receptacle.
5. To avoid measurement errors from leakage into the measuring circuit, connect the output of the auxiliary oscillator to the binding posts of the Coupling Unit with a length of *SHIELDED CABLE*, connecting the center conductor to the red and the shield to the black insulated post. (CAUTION: Turn voltage output of oscillator to zero before making this connection.)
6. Select the desired frequency on the auxiliary oscillator, and increase the output voltage slowly until the Multiply Q By meter indicates X1.

C. Measurement Procedure

For low frequency measurement instructions, refer to the Q Meter Type 260-A instruction manual.

D. Input Voltage Requirements

Because the response of the Coupling Unit varies slightly with frequency (largely because of the capacitance of the output cable), the oscillator output voltage required to produce a X1 reading on the Multiply Q By meter will increase at frequencies approaching 50 kc. Fig. II indicates the signal voltage requirements within the frequency range of the Coupling Unit.

E. Low Frequency Q Voltmeter Correction

The Q-indicating voltmeter circuit of the Q Meter has been bypassed in order to provide optimum performance at frequencies between 50 kc and 50 mc. For this reason the response of the voltmeter to low audio frequencies is not flat, and a correction, indicated in Fig. III, must be applied to the observed Q or LO Q reading to obtain the corrected value of indicated Q.

F. Use With Q Meter Type 160–A

Although designed primarily for operation with the Q Meter Type 260-A, the Coupling Unit may be used with the older Q Meter Type 160-A if the following requirements are observed:

1. The auxiliary oscillator used must have a maximum output of at least 30 volts. Input voltages higher than those indicated in Fig. II will be required.
2. An adapter must be prepared to permit connection of the Coupling Unit output to the phone jack on the top of the Q Meter cabinet. Such an adapter may consist of a standard phone plug and a UG-291/U receptacle, connected by a short length of RG-58/U cable.

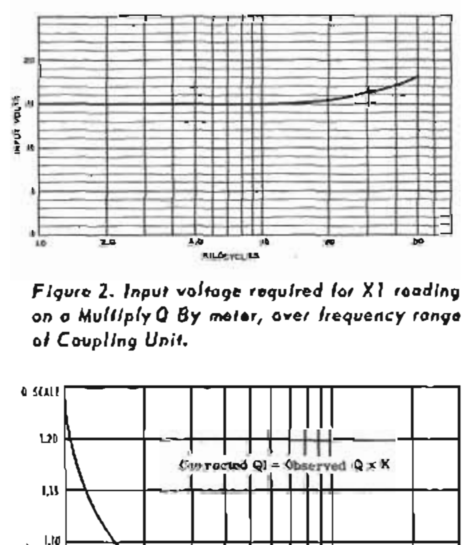

Figure 2. Input voltage required for X1 reading on a Multiply Q By meter, over frequency range of Coupling Unit.

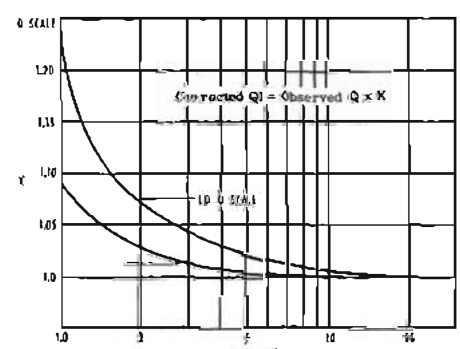

Figure 3. Low frequency corrections for Indicated Q observed on Q Meter Type 260–A.

An RF Voltage Standard Supplies A Standard Signal At A Level Of One Microvolt

CHARLES G. GORSS, *Development Engineer*

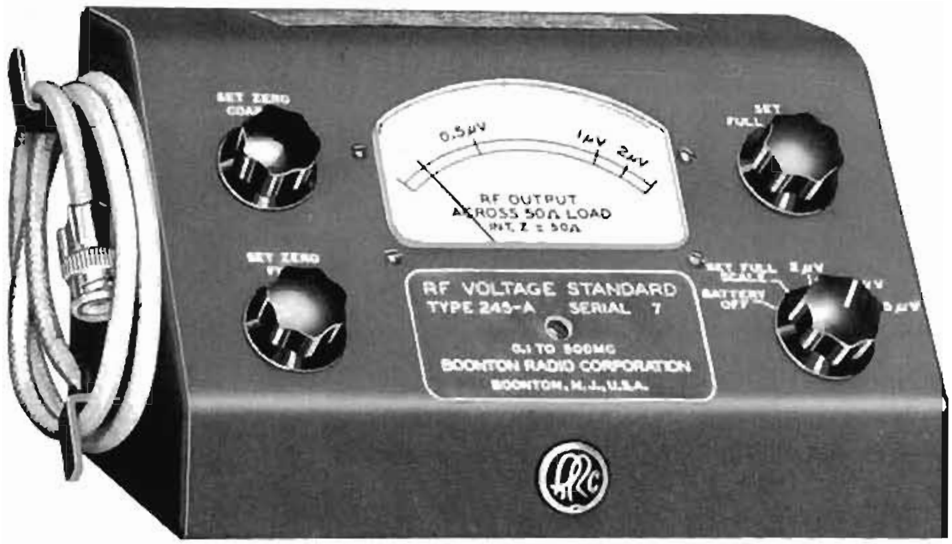

Figure 1. The RF Voltage Standard Type 245–A.

The sensitivity of a radio receiver is a well advertised feature, and properly so since it is one of the most important attributes of a receiver. This sensitivity often is in the order of one microvolt or less, and the direct measurement of rf voltages at this level is seldom possible and never accurate.

The generation of a level of voltage which cannot be accurately measured presents a problem since the receivers under test are the only devices capable of even detecting the presence of these diminuative voltages. To use a receiver of unknown sensitivity to measure these levels would not be an accurate process. A solution to the problem is a source of rf voltages, at microvolt levels, which can be established with a definite and reasonable accuracy without actually measuring it at the low levels.

GENERATOR OUTPUT SYSTEMS

Numerous signal generators are in existence with an output range including 1 microvolt. They usually depend on a voltmeter which monitors the voltage input to an attenuator system having a very large attenuation ratio and whose internal impedance generally varies with frequency and often is not accurately known. Both the voltmeter accuracy and the coupling to the attenuator may vary with frequency.

The piston, or mode cutoff, attenuator is regarded as one of the best types of attenuators commonly used in signal generators today. However, there are serious drawbacks to high precision over a broad band of frequencies. The fact that there are spurious modes generated in the attenuator detracts from its accuracy. Of course, the effect can be calculated for a mechanically perfect unit.[1] However, the dimensions, the roundness of the tube or lack thereof, and the angular alignment of the input and output loops all influence the accuracy of such an attenuator to a great degree. Besides this, the apparent diameter of the tube changes as the frequency drops, due to an increasing penetration of the metal of the tube by the field. One last and serious drawback is that no check with precise direct current instruments is possible.

VOLTAGE TRANSFER PROBLEMS

Even if one were capable of producing an accurate output of one microvolt level from a system, there must be a way to accurately transfer it to a receiver. The device should therefore have a resistive output impedance equal to the characteristic impedance of the cable to be used to connect the standard source to the receiver. The subject of connecting signal generators to receivers has been adequately covered in a previous article[2] in The Notebook.

A device which could produce an accurate level of one microvolt through a known impedance would be a powerful tool for standardizing signal generators being used to check sensitivity of receivers. In the development laboratory, or on the manufacturing line, or in the calibration laboratories of a large receiver user, such as an airline which wants to be sure of receiver performance, there is an immediate need for an instrument of this kind. The RF Voltage Standard Type 245-A is such a device.

CALIBRATED RF VOLTAGE SOURCE

The RF Voltage Standard Type 245-A, shown in Figure 1, takes approximately the full output of an average signal generator over the frequency range of 1-500 mc, precisely monitors the high level input to an attenuator system, and accurately attenuates this voltage to the microvolt region. The now-accurately-known signal at microvolt level appears in the output circuit in series with a 50 ohm resistive impedance.

The use of this device to calibrate a signal generator is relatively simple as illustrated in Figures 2 and 3. The signal generator is connected to the RF Voltage Standard by the integral input cable, and the generator rf output increased until the RF Voltage Standard monitor meter reaches the reference line corresponding to the desired output level. The output is connected to the receiver antenna terminals by an output cable with a 50 ohm termination such as a Type 501-B.

Once the receiver output indication has been recorded, the Type 501-B cable is connected to the signal generator output in place of the RF Voltage Standard as shown in Figure 3. The signal generator attenuator is turned down till the same output indication is observed on the receiver as had been previously recorded for the output of the RF Voltage Standard. The signal generator is now furnishing the same level as the RF Voltage Standard had been and the signal generator attenuator setting is recorded as a standardized point.

ACCURACY REQUIREMENTS

Of course, no comparison Standard can correct for errors arising from voltage standing waves on the output induced by incorrect values of the generator output or load input impedance and this impedance must be close to 50 ohms for maximum accuracy.

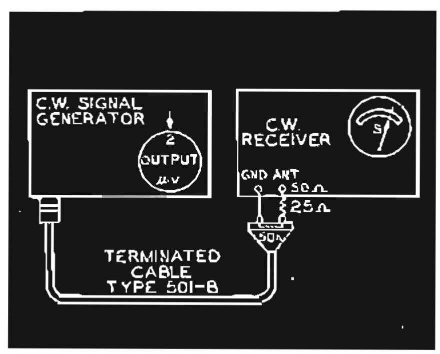

Figure 2. Step No. 1 RF voltage at high level is obtained from the signal generator and measured. The attenuated output is detected at low level on a receiver and a reference reading noted. Calibration applies to the end of the terminated output cable.

ATTENUATOR SYSTEM

The accurate attenuator system, which is the heart of the RF Voltage Standard, is a variation of the Micropotentiometer principle described by M. C. Selby of the National Bureau of Standards.[3] The output voltage is generated by causing a monitored current to flow through a 2 milliohm concentric disc resistor. This disc resistor is thinner than the penetration depth of the current by several times at all useable frequencies. The consequence of this fact is that the resistance can be determined very accurately by DC means and will not vary from this appreciably in the frequency range of this device. By the very nature of the symetrical annular design the inductance of the disc is of a negligible magnitude.

The current into the disc is controlled by placing the disc in a coaxial system at the terminus of a resistive concentric line which is designed to have no frequency dependance as shown in Figure 4. The resistance of the line is 50 ohms, and the line appears at its input as a non-reactive 50 ohm termination.[4] As far as the input is concerned, the 2 milliohm disc is a short circuit.

A diode voltmeter monitors the input to the termination. The VSWR of the resistive line is very low to well above 500 mc and therefore the current flowing in it will remain constant with frequency changes if the voltage at the input is constant. The output impedance of the device is controlled by a 50 ohm resistive section on the output side of the disc resistor. This resistive line is designed exactly like the input section. The output connector is joined to this section by a fifty ohm loss-less air line having negligible discontinuities.

MONITORING SYSTEM

The crystal voltmeter system uses a new UHF crystal diode in a specially designed coaxial mounting. The crystal is operated in a rather unique fashion. A 100 microampere current flows through the crystal at zero rf input. This bias stabilizes the characteristics of the diode to a much greater degree than they would be in a no bias state. The DC output current of the diode is taken from the rf assembly through an rf filter which prevents any leakage from the high level side of the attenuator.

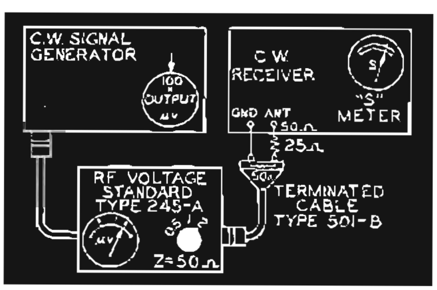

Figure 3. Step No. 2. The low level output of the signal generator is adjusted to produce the same reference level reading on the receiver as was produced by the unknown low level output of the RF Voltage Standard.

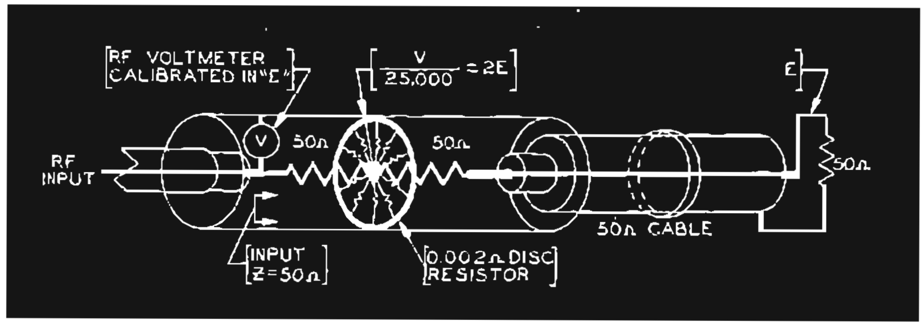

Figure 4. RF Voltage Monitor and Control Attenuator System of the RF Voltage Standard Type 245–A.

In order to keep the low impedance of the crystal detector from seriously damping the 20 microampere dc output meter, a common base transitor amplifier is used. The output impedance of this configuration is high enough to leave the meter just about critically damped. The current amplification of such a device is essentially α or about 0.98 times the input current for the RD2521A transistor which is used in this instrument. This represents no serious loss in output, and since it is a very stable property of a transistor there is no serious loss of stability.

Power to operate the transistor and the crystal bias circuitry comes from a small self-contained battery used at low current drain. Battery life is essentially "shelf life".

CONCLUSION

In conclusion, the RF Voltage Standard Type.245-A is a simple, self-contained, portable and accurate solution to the problem of low rf level measurements. It offers outputs of 0.5, 1, and 2 microvolts with significant accuracy and has an output impedance of 50 ohms with a low Voltage Standing Wave Ratio. This instrument can offer great assistance toward resolving the existing confusion in the comparison of signal generators and receiver sensitivity measurements.

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NOTE FROM THE EDITOR

Every Spring your editor hears the noise of unusual activity in the Sales Department. This Spring the noises have reached an unusual volume. We were finally approached with a request for permission to include some descriptive information on new equipment in the envelope with this issue of The Notebook. It is our policy to include in these pages information of the most general technical interest possible. We are not, however, insensitive to the value of really new information on new equipment. The new equipment will be on display at the IRE Convention in New York City at Kingsbridge Armory. However, many of our readers may not be able to attend the Convention. For them especially we have included some extra information in our envelope. We believe you'll profit by reviewing our enclosure.

We published our first issue of the Notebook one year ago this month. In the first issue we indicated our intention of distributing information of value on the theory and practice of radio frequency and measurement. We believe that an article on the last page of the current issue listing on "INDEX" to previous issues of The Notebook is worth reviewing to determine how we have implemented our intention. Our editorial plan for The Notebook requires the publication of five articles in each issue; one article in each of the following categories:

1. Articles of broad technical interest of general and lasting usefulness in the measurement field.

2. Articles of more popular technical interest and practical nature.

3. Articles covering instrument application or service material.

4. Non-technical articles of general interest.

5. Notes from the Editor.

IF YOU FIND MATERIAL OF INTEREST IN THE INDEX WHICH IS NOT IN YOUR HANDS, PLEASE, LET US HEAR FROM YOU. We still have a few extra copies of past issues.

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The NOTEBOOK

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Some VHF Bridge Applications

NORMAN L. RIEMENSCHNEIDER, *Sales Engineer*

Recognizing the gap that inevitably exists between the basic information offered in the instruction manual and the inherent potentialities of the instrument involved, this article is written with the intent of sharing our knowledge of the RX Meter Type 250-A and offering the benefit of our field experience.

For those not having a speaking acquaintance with the RX meter, we might briefly mention that it consists of a V.H.F. Schering bridge, signal generator, and a detector-indicator completely self-contained in one unit which measures impedances from 500 KC to 250 MC in terms of equivalent parallel resistance and parallel capacitance (or inductance). It will be shown that with these basic elements and some understanding of the principles involved, the bridge lends itself to many applications not immediately obvious.

The Author off on an engineering field problem with the bridge.

Measurement of Transistors, Vacuum Tubes, Diodes and Biased Circuits

The RX Type 250-A Meter by virtue of its design lends itself very conveniently to the measurement of circuits under conditions of normal bias and operating voltages. Since there is a d.c. path between the bridge terminals with a d.c. resistance in the order of 66 ohms, biasing currents up to 50 ma can be introduced through the bridge and the component in the manner shown in Figure 1. In using this particular circuit the capacitor C should be made large enough so as to have negligible reactance at the operating frequency. It is desirable to use a d.c. source having a much higher voltage than required, and reducing the voltage through resistor R. By keeping R high the effect of the D.C. supply source is minimized.

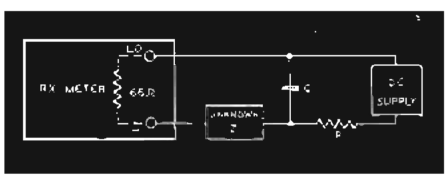

Figure 1. Method of Applying DC Biasing Current Less Than 50 ma.

For measurements requiring biasing currents in excess of 50 ma the circuit shown as Figure 2 is suggested. In this arrangement the biasing current circumvents the bridge and therefore there are no restrictions imposed upon its magnitude. Here again it is desirable to keep C large enough to have negligible reactance. Since the purpose of the R.F. choke is to isolate the d.c. voltage source its inductance should be high. To preclude the possibility of any error due to the power supply, certain precautionary steps can be taken. Before attaching the test specimen, balance the instrument with the D.C. supply circuit connected. If the range of the balance controls is insufficient, balance the instrument by itself and measure the Cp and Rp of the D.C. supply circuit alone. The subsequent readings of the test specimen can then be corrected (for the effect of the D.C. supply circuit) as follows:

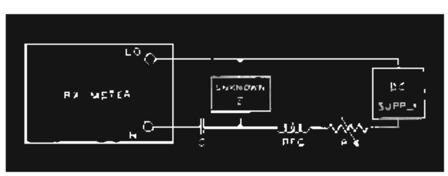

Figure 2. Method of Applying Biasing Current Greater Than 50 ma.

$$C_{px} = C_{xo} - C_o$$

where

$C_{px} = C_p$ of specimen

$C_{xo} = C_p$ of specimen & dc supply

$C_o = C_p$ of dc supply alone

$$R_{px} = \frac{R_o \times R_{ox}}{R_o - R_{ox}}$$

where

$R_{px} = R_p$ of specimen

$R_o = R_p$ of dc supply alone

$R_{ox} = R_p$ of dc supply & specimen

The basic circuits shown in Figures 1 & 2 can be modified or expanded as required to allow measurements to be made of the desired parameters in many types of circuits. Some typical arrangements are given in Figures 3 to 9. In Figures 3 (tube input impedance) and 4 (crystal diode impedance) it should be noted the ground shown by dotted line is the circuit ground and is not connected to the instrument ground. Figures 5 to 9, for which the author is indebted to the "Radio Development & Research Company" of Jersey City, New Jersey, are examples of some typical circuits employed in the measurement of transistors.

One factor that bears consideration in the measurement of vacuum tubes, transistors, and diodes is the level of the r.f. voltage applied to the component under test. With the RX Meter this is in the order of from 0.1 to 0.5 volts r.m.s. Should

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the constants of the circuit be such that this level cannot be tolerated, the addition of a 2 watt, 50,000 ohm potentiometer, mounted in a louvre hole in the rear of the instrument and wired according to Figure 10 to reduce the oscillator B+ voltage, makes it possible to lower the applied r.f. voltage down to 0.02 volts with satisfactory sensitivity during the measurement.

Measurement of the Self-Resonant Frequency of a Coil

This procedure consists of finding the frequency at which the inductive and capacitive reactances of the coil are equal (and opposite), and the coil itself looks like a pure resistance. The logical way to do this would be to set the Cp dial equal to zero and vary the frequency to obtain a null on the indicating meter. However, since it is necessary on the RX Meter 250-A to peak the "detector tuning" when operating the instrument at the higher frequencies, the self-resonant frequency is found by making two or three rapid preliminary measurements, balancing the instrument, and obtaining the final value.

The actual modus operandi is as follows:

1. Set up the bridge in accordance with the preliminary instructions given in the instruction manual. For the first approximation select a frequency higher than the estimated self-resonant frequency.
2. Mount the coil on the instrument and measure it in a normal manner. If the Cp dial reads in the capacitive range, the frequency is too high and should be lowered for the next approximation. Conversely, if the Cp dial indicates an inductance the frequency should be increased. The frequency change necessitated can be sensed from the effect on the Cp dial reading of the previous change.
3. When the Cp dial reads within 1 or 2 mmfds of the zero line, remove the coil and balance the bridge before making the final reading.

It should be noted that while some manipulation of the Rp dial and some adjustment of the "detector tuning" knob might be needed in making the rough approximations, it is necessary to adjust the initial balance of the bridge *only* for the final measurement.

With those coils having multiple resonances, the alternate parallel and series

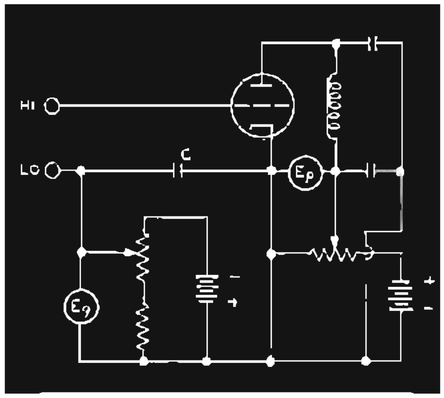


Figure 3. Typical Circuit for Measuring Vacuum Tube Input Impedance on RX Meter.

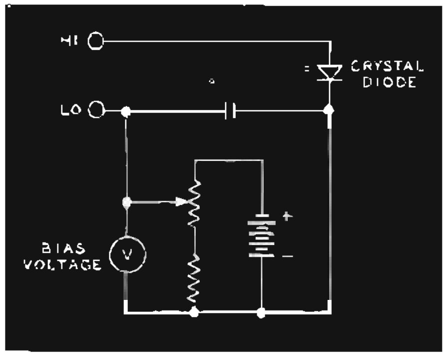


Figure 4. Typical Circuit for Measuring Crystal Diodes on RX Meter.

points can be identified by the accompanying respective high or low values of parallel resistance indicated on the Rp dial. The lowest parallel self-resonant frequency is the one used in the determination of distributed capacity discussed below.

Measurement of the Distributed Capacitance (Cd) of a Coil

To consider an old friend very briefly, the expression for the resonant frequency of a series tuned circuit is:

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad (1)$$

This same relationship obtains with a parallel tuned circuit, for all practical purposes, if the Q is ten or greater. For the purpose of our discussion we shall consider coils falling in this category.

Referring to (1) above, if the inductance is held constant but the capacitance changed to a new value, C_2, then:

$$\frac{f_1}{f_2} = \frac{\dfrac{1}{2\pi\sqrt{LC_1}}}{\dfrac{1}{2\pi\sqrt{LC_2}}},$$

or

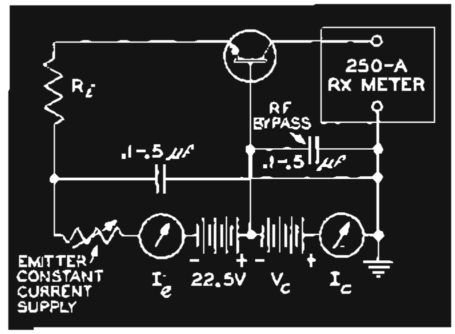


Figure 5. Collector Characteristics -- Grounded Base Tetrode.

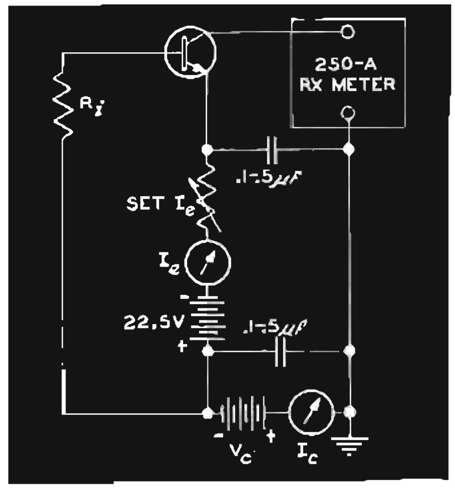


Figure 6. Collector Characteristics -- Grounded Emitter Triode.

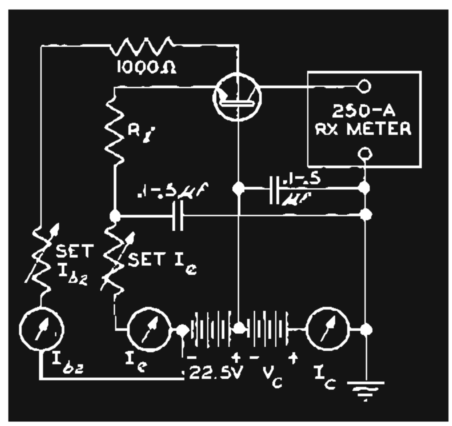


Figure 7. Collector Characteristics -- Grounded Base Tetrode.

$$\frac{f_1}{f_2} = \frac{\sqrt{C_2}}{\sqrt{C_1}} \quad \text{and}$$

$$\frac{C_2}{C_1} = \left(\frac{f_1}{f_2}\right)^2 \qquad (2)$$

If, in the case of a coil, f_1 is the self-resonant frequency of f_o, and C_1 is the distributed capacity of the coil, C_d, the relationship then becomes,

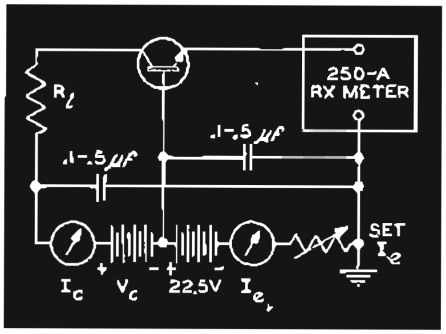


Figure 8. Emitter Characteristics Grounded Base Triode.

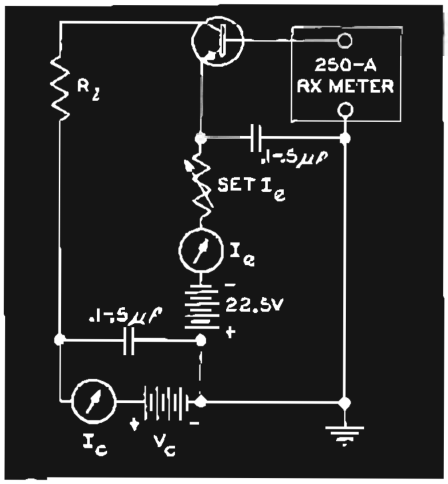


Figure 9. Emitter Characteristics - - Grounded Emitter Triode.

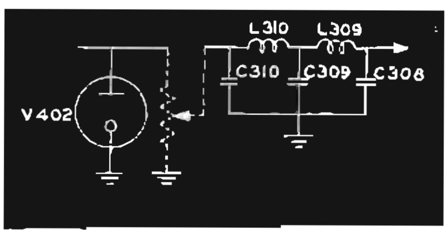


Figure 10A. Terminal Voltage Reduction-Circuit Alteration.

$$\frac{C_2}{C_d} = \left(\frac{f_0}{f_2}\right)^2, \tag{3}$$

and if frequency f_2 is made equal to 70.7% of f_0,

$$\frac{C_2}{C_d} = \left(\frac{f_0}{.707\, f_0}\right)^2 = 2\,. \tag{4}$$

For the purpose of our discussion this can be written

$$\frac{C_d + C_p}{C_d} = 2 \text{ and} \tag{5}$$

$$C_d = C_p. \tag{6}$$

Where Cp is the amount of capacity needed in addition to the C_d of the coil to resonate the coil at a frequency equal to 0.707 times its self-resonant frequency. This value is equal to the C_d of the coil and is

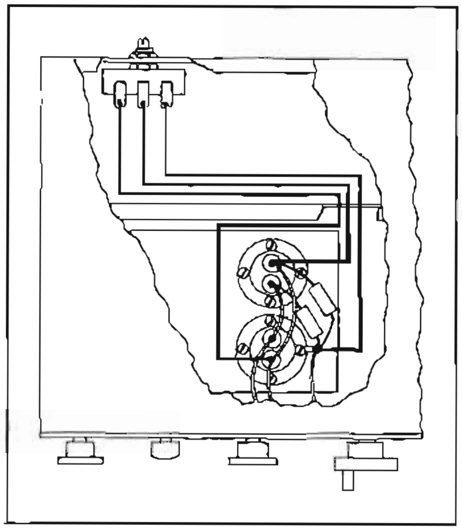


Figure. 10B. Diagram of Circuit Alteration.

read directly off the silver portion of the C_p dial on the RX meter when the bridge is balanced.

Thus to measure the C_d of a coil on the RX Meter it is only necessary to determine its self-resonant frequency, set the bridge at 0.707 times this frequency, and measure the C_d directly on the C_p dial.

Measurement of Tuned Circuits

Very often it is necessary to measure the R_p or Q of a tuned circuit at its resonant frequency. Since under these conditions the circuit is essentially resistive, it merely necessitates operating the RX Meter at the resonant frequency, verifying the resonant condition by insuring the C_p dial reads zero, and reading the dynamic resistance directly off the R_p dial. Conversely, the RX Meter can be set at the desired frequency and the tuned circuit adjusted for resonance as evidenced by a null on the indicating meter. Having the R_p and being able to measure or compute either reactance, the Q of the tuned circuit can be determined from the relationship, $Q = R_p/X_p$.

By setting the RX Meter 250-A to the desired frequency, and the R_p and C_p dials to the required values, the instrument adapts itself to the tuning of pi-coupled matching networks used on transmitters and to the adjustment of line terminations, equalizers, and filters.

Extension of Ranges

The RX Meter 250-A has a parallel resistance range of 15 to 100,000 ohms and a parallel capacitance range from +20 mmfd to —100 mmfd. The negative symbol denotes an inductance and the quantity is the amount of capacitance required to resonate it. These ranges can be extended by the following means which are described for the particular extension desired.

A. Manner of Extending Capacitance Range

Referring to Figure 11, which is a graphical representation of the Cp dial on the RX Meter 250-A, it can be seen that when the dial is set at the zero point, for the purpose of obtaining the initial balance, there actually is a net capacity of 40 mmfd across the terminals. Regardless of what measurement is being made, it is always necessary to have this net capacity of 40 mmfd for the bridge to balance; whether it is obtained by setting the Cp dial to zero, or setting it at +20 mmfd as would be done when actually measuring a capacity in this order, or measuring an L/C combination exhibiting a net capacity of 40 mmfd is of no consequence. As long as this net amount is present, the bridge will balance.

Therefore, to extend the range of capacitance measurement—say to something greater than 20 mmfd but not over 120—it is only necessary to add a coil during the initial balance that will cause the bridge to balance with the C_p dial at —100 mmfd instead of at zero. For a balance point at —100 mmfd, this coil would have the amount of inductance necessary to resonate with 100 mmfd at the chosen frequency. Using the —100 mmfd value as the reference point in this manner, and leaving the coil intact during the measurement, the capacity range now is from —100 to +20 mmfd—or a total of 120 mmfd. The inductance of the coil does not have to be known and the reference balance can be established at any desired point from zero to —100. This allows quite a latitude in the selection of the coil. There are no restrictions imposed upon the coil with respect to Q, since if the latter is not sufficiently high the equivalent R_p of the coil itself can be read in the initial balance, and proper allowances made for it. From experience it has been found that in most cases any reasonable hand-formed coil will suffice without requiring corrections.

Referring once again to Figure 11, it can be seen that by using a smaller coil B and adding a 100 mmfd. capacitor externally across the bridge binding posts to be used in conjunction with the 100 mmfds on the Cp dial, it is possible to displace the original balance by 200 mmfd, and in this manner increase the capacitance range to 220 mmfd. The capacitance measurement thus becomes essentially one of substitution, since after the unknown is added to the terminals, balance is sought by removing capacity on the Cp dial, 0.1 mmfd at a time. If the measurement of the unknown is not made by the time the C_p dial reaches the +20 point, the 100 mmfd capacitor used for extending the range is removed, the Cp dial returned to —100 and the process repeated until balance is obtained. In this way although the range-extending capacity is added in an increment of 100 mmfd, it is removed by using the Cp dial in increments of 0.1 mmfd. In a like manner, other 100 mmfd capacitors, in conjunction with smaller coils, can be used to extend the range of capacitance measurement still further. A variable pre-

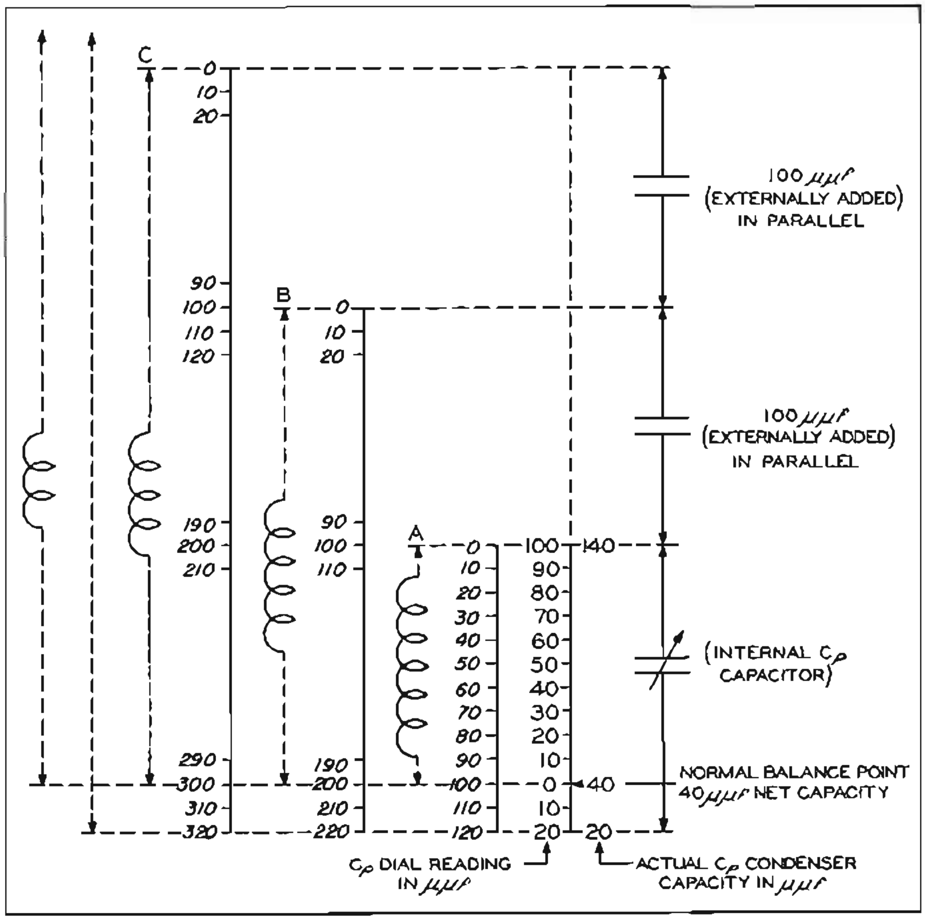


Figure 11. Graphic Illustration of Manner of Extending Capacitive Range of RX Meter 250-A

cision capacitor is a very convenient device in these extensions since it increases the inductance latitude of the coils used for the extension. In all cases the value of capacity employed must be known since it is part of the calibration.

Since the bridge has an inherent residual inductance of 0.003 microhenries (in series with the impedance being measured), the following restrictions are imposed to prevent the residual inductance from becoming a sizeable portion of the reading:

1. Do not use auxiliary coils having less than 0.1 uh inductance.
2. If necessary to use a coil of less than 0.6 uh it must be adjusted to a reactance value within ±20% of the capacitance reactance under test and the following correction formula applied to the RX Meter readings:

$$\text{True Cap} = \Delta C \left(1 + \frac{.003}{L}\right)$$

Where:

C = Difference reading when test is connected in uuf

L = Auxiliary inductance in uh

3. The above limitations and corrections apply only when the junction between test capacitor and auxiliary coil is directly under the knurled binding post nuts.

For those cases where the frequency of operation does not allow the size of the coil used to conform to the above, it is possible to correct for the residual inductance as follows:

$$\text{True } C_p \text{ dial reading} = C_p' \left(\frac{1}{\omega^2 L_1 C_p' - 1}\right)$$

Where C_p' = Capacity as read on the C_p dial

$$\omega^2 = 4\pi^2 f^2$$

$$L_1 = 0.003 \text{ microhenries}$$

Correcting the dial reading as shown above for the Cp readings with and without the addition of the capacitance under test will make the required allowances for the residual inductance and the capacitance range can be extended in the normal manner. Care must be exercised with respect to minimizing lead lengths, placing components, etc.

Manner of Extending the Inductance Range

For the measurement of the inductance of coils requiring more than the 100 mmfds. available in the 250-A RX Meter to resonate them at the operating frequency, additional known capacity can be added in parallel to the coil to obtain balance. The inductance of the coil would then be

$$L = \frac{1}{\omega^2 (C_p \text{ dial rdg} + \text{ext. } C)}$$

or the reactance

$$X_L = \frac{1}{\omega (C_p \text{ dial rdg} + \text{ext. } C)}$$

In the above, external capacity is considered as having a negative sign as has Cp in this case. It has been found the range of measurement can be extended to include much smaller value of inductance by using an auxiliary resistor, connected in series with the "high" side of the coil. In this manner the overall Q is reduced which in turn allows smaller values of resonating capacitances in accordance with the relationship.

$$L_s = \frac{C_p R_p^2}{1 + Q^2}$$

The value of the auxiliary resistor, which is not critical, depends upon the inductance to be measured and some idea of the required value can be obtained from the following table.

INDUCTANCE RANGE (Microhenries)	REQUIRED RESISTOR (Ohms)
10 to 1000	1000
1 to 10	300
0.1 to 1	100
0.001 to 0.1	33

The value and accuracy of the auxiliary resistor is not critical and need only be of the correct order. The following procedure is suggested for such measurements:

a. Connect the unknown inductance in series with the auxiliary resistor across the RX Meter binding post. Using a minimum length of heavy, conducting strap, short the terminals of the inductance to remove it temporarily from the circuit.

b. Balance the bridge circuit and note the values of C_{p1} obtained for the series resistor alone.

c. Remove the shorting strap from the inductive component, restoring the latter to the circuit, and rebalance the bridge. Note the values of R_{p2} and C_{p2} for the series combination. Then the unknown inductance is obtained by

$$L_s = \Delta C \, (R_{p2})^2$$

$$\text{where } \Delta C = C_{p1} - C_{p2}$$

It should be noted that the inductance is shorted out rather than removed to avoid alteration of the physical configuration of the components which might otherwise affect the results. In dealing with extremely small inductance values, the inductance of the shorting strap itself will become significant and must be considered in interpreting the results.

Extension of the Rp Dial Below 15 Ohms

In dealing with low Q devices it is sometimes desirable to be able to measure resistance values below 15 ohms which is the lower limit of the direct-reading Rp scale.

At higher frequencies (in the neighborhood of 200 mc and above) the residual inductance of most components having series resistance values below 15 ohms such as low-value resistors, may be sufficient to increase the equivalent parallel resistance value above 15 ohms so that it may be measured directly. If not, a small inductance (having negligible series resistance) connected in series with the unknown will be sufficient to increase the Rp of the combination to the range of direct measurement.

At lower frequencies, the Rp of the unknown may be effectively increased for measurement by adding in series a small auxiliary resistor having a value preferably between 15 and 25 ohms. The series combination is measured and the values R_{p1} and C_{p1} are noted. The auxiliary resistor is then measured alone to obtain R_{p2} and C_{p2}.

Then

$$R_S = R_{S1} - R_{S2} \quad \text{and}$$

$$L_S = L_{S1} - L_{S2}$$

Series resistance in each case would be

$$R_S = \frac{R_P}{1 + Q^2}.$$

C_{p1} and C_{p2} can be changed to their respective values of inductance L_{s1} and L_{s2} in accordance with

$$L_S = \frac{C_P R_P^2}{1 + Q^2}.$$

Conclusion

The foregoing discussion has not been presented in an effort to define the limits of the bridge, but rather to give some indication of its application in regions not readily apparent. As with any tool its ultimate potentialities are dependent upon the skill and ingenuity of the person using it.

THE AUTHOR

Norman L. Riemenschneider has been intensively engaged in field work on customers' problems and special applications of Boonton Radio Corporation's equipment since his association with the Company. His background includes eight years with Western Electric in various engineering capacities, and substantial experience in allied fields with other engineering firms. Mr. Riemenschneider was graduated from the evening section of the Newark College of Engineering with a B.S. in E.E. and is a member of the IRE. He is also very active in Amateur Radio circles.

MECHANICAL DESIGN REQUIREMENTS OF ELECTRONIC INSTRUMENTS

DAVID S. WAHLBERG, *Mechanical Design Engineer*

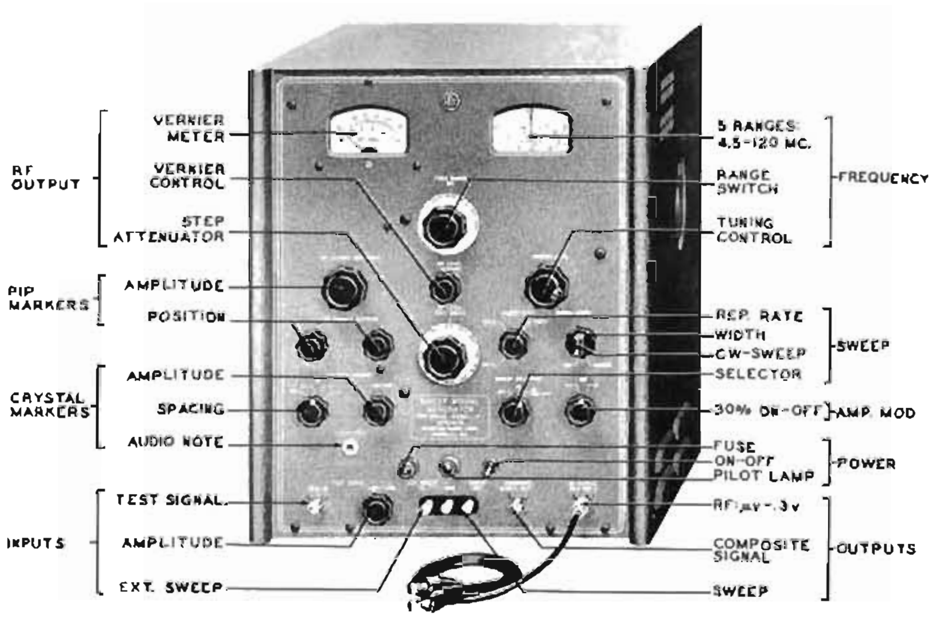


Figure 1. The Front Panel of the BRC Sweep Signal Generator Type, Type 240-A illustrates the final solution of a mechanical design problem involving mechanical placement, electronic function, nomenclature and esthetic factors.

One of the most interesting and challenging phases in the design of an electronic instrument is that concerned with converting a set of specifications and a breadboard model into a practical, economical and useful product. It is at this stage of the endeavor that many of the problems become mechanical in nature.

The Mechanical Designer's Problem

From the moment the mechanical design of an instrument begins, a myriad of other considerations arise to confront what might otherwise seem a straightforward piece of electronic equipment. The mechanical designer must consider the electronic requirements of the Development and Project engineers, the functional and saleable appearance, weight and price insisted upon by Sales, and the mechanical urgencies of simple, rigid designs and drives using the proper materials. In addition the Shop must be allowed reasonable tolerances within the limitations of available processes and equipment. Assembly should have units adapted to smooth work flow, and Inspection (and the user!) needs easy, accessible adjustments. Among many other factors are Purchasing's and Accounting's hopes that standard parts will be used, and Shipping's plea for enough unobstructed cabinet area to allow proper bracing in the packaging.

Thus, as those in industry know, any design comes about as the result of many compromises, all measured against the ultimate goal.

The Front Panel

Almost always, one of the first operations in the design program is the preparation of a front panel layout. However, this drawing also usually turns out to be the last one finished. A typical instance is the Sweep Signal Generator Type 240 A panel of Figure (2). The final symmetrical and functional grouping of the controls was arrived at after eleven distinct drawing revisions. Included were several major and many more minor changes, each one the result of discussion and action as the various problems of mechanical placement, electronic function, and clear titling were solved. As the give-and-take proceeds between esthetic and functional requirements versus circuit and control mechanism considerations, the front panel layout also starts to include the influences of framework and cabinet design.

The Heat Dissipation Problem

At this point the need for ventilation must be reckoned with. Close-accuracy instruments usually require heavy duty, constant voltage power supplies with much attendant heat. Therefore, such a power supply is usually separated from other

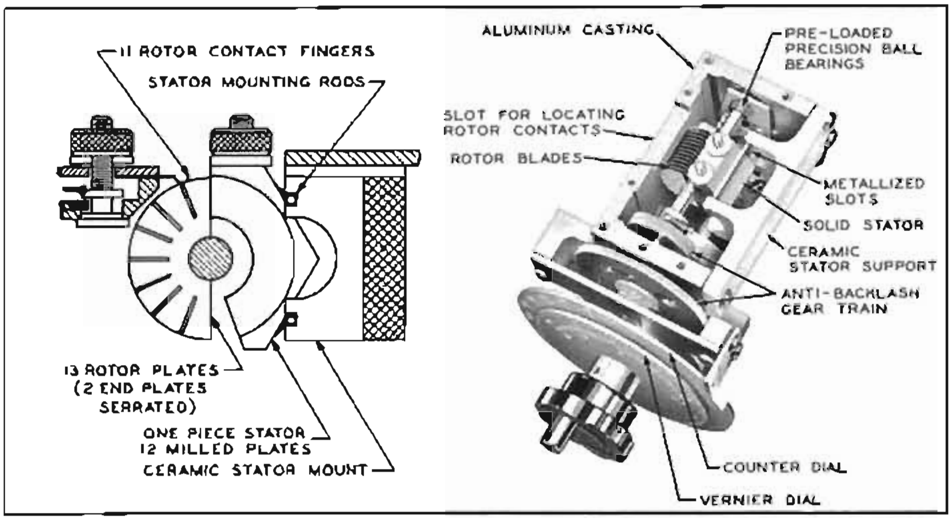


Figure 2. The 190-A Q Unit solves the problem of rigidity, low capacitance and constant inductance.

circuits as far as possible, either physically or by shielding, since it is the largest single source of heat. Once the power supply and other elements are located, and an estimate made of the heat to be dissipated, the ventilation paths and louvers are roughed out. Although most heat is carried off by convection if possible, the complete electrical shielding often needed around RF circuits prevent this. In these cases, heat must be conducted away by providing shields or heavy conductors leading to a large heat sink. Radiant heat is also a factor under these conditions. Bright metal shields can serve to "bottle up" a good deal of heat energy, whereas proper treatment of the surfaces will cause them to absorb and help dissipate heat. Final word as to the adequacy of the cooling provided in an instrument usually awaits the completion of final temperature runs on the prototypes. Convection currents occasionally cause uneven heating of critical components, and such problems must be resolved by baffles or circuit adjustments before the design is completely "frozen"

Dealing with RF Leakage

The means of preventing RF leakage have a major influence in dictating the overall design layout. All circuits carrying heavy RF currents must be carefully bottled up with leakage paths kept to a minimum. Control shaft holes particularly must be bridged by efficient shorting devices. The shafting necessary to operate an RF unit may often be just the right length to perform as an antenna at UHF frequencies. Gaskets or other sealing means are necessary to seal all covers and all elements projecting into an RF field.

RF leakage often does not rear its ugly head until near the end of a design project. As the final tests are being run several instruments may pass well within the limits, only to be followed by another seemingly incurable "leaker". This may need only a slight increase in contact pressure at a crucial point for a cure. But occasionally the mechanical designer finds to his chagrin that a scarely measurable amount of RF has found a configuration of supports and surfaces that needs only a little RF energy to prove itself a resonant circuit of inductance and capacitance. Last minute changes are then in order to tie down this last loose end.

The Need for Rigidity

Mechanical rigidity is a necessity in any accurate tool and is, in fact, often a measure of the accuracy attainable. A brief description of the Rp dial drive of the RX Meter Type 250-A will underline the reasons for the massive structures occasionally required in an electronic instrument. One arm of the 250-A bridge network is adjusted by means of a small variable capacitor, C2, to which is geared the Rp dial, measuring parallel resistance. Although the capacitance range is only about 18 mmf., the dial scale length is expanded to about 28 inches with the graduations spread out approximately logarithmically. Therefore, at the end of the dial indicating infinite resistance, where the capacitance is near maximum, the increments must be extremely small.

This effect is achieved by a special shape of the rotor plates, but the practical realization of such small increases in capacity depends largely on the accuracy and rigidity of the framework and drive. In the 250-A these requirements are met by supporting the rotor on preloaded ball bearings carried in a massive housing. Anti-backlash gears are of course a necessity and the entire system is mounted in a rigid casting. These provisions make the rotor capable of resisting any deflection except the deliberate rotational movements needed in adjusting capacity. At the 500K ohm point on the dial, where the capacitance stability must be of the order of 0.025%, an extraneous rotor movement as small as 0.0001 inch will result in inaccuracies beyond the specification limits.

The internal resonating capacitor might well be called the heart of the Q-Meter, and is an excellent example of the interdependence of mechanical and electrical design. Taking the 190-A Q unit as an example, the electronic requirements are low minimum capacitance, together with low and constant values of inductance and resistance all of which are difficult to attain in conventional designs. Mechanically the design must be extremely rigid to assure constant and accurate re-setability. The massive structure ordinarily needed to attain the last-named end is in direct opposition to the minimum capacity requirement.

Reference to Figure (2) will show how satisfactory solutions were found for those conflicting needs.

The stator is mounted by means of two rods soldered into the metallized slots of a high quality ceramic support. Insofar as possible it floats in air dielectric. In addition, the rotor travel is restricted to less than 180°, with the included angle of the stator reduced by a proportionate amount to result in the largest possible angular gap between the two at the minimum setting. By these means the capacity at minimum was limited to the lowest value compatable with sufficient mechanical strength.

Low and constant inductance between the stator plates is achieved by milling out a solid bar to leave only the outer shell and the plates, solidly connected with each other along their entire peripheries. A secondary result of this method is the "built-in" shielding the shell provides from extraneous fields.

The tandem edge wipers, contacting all the rotor blades in parallel, serve to reduce the associated inductance and resistance to a very low and nearly constant value and are rhodium plated to provide good wearing qualities.

Where other considerations do not enter, rigidity is attained in the complete unit by mounting all the parts on a rigid cast frame. All shafts are carried on preloaded ball bearings.

Conclusion

It is hoped that this brief description of instrument building from the mechanical point of view has shown how absorbing this part of the design effort can be, and that it is not separate unto itself but as much a part of the program as is the electronic design. Nevertheless a Mechanical Design Department would be something less than truthful for not admitting to the occasional feeling that it was also the focal point of almost everyone's problems. However, closer to the general truth is the plain fact that the Mechanical Designer has need for and receives much help from many people in the course of a design. A well-designed electronic instrument is the result of the co-operative teamwork of all those concerned with its inception, design and manufacture.

SERVICE NOTES FOR THE RX METER TYPE 250-A

BRUNO BARTH, *Inspection Department Foreman*

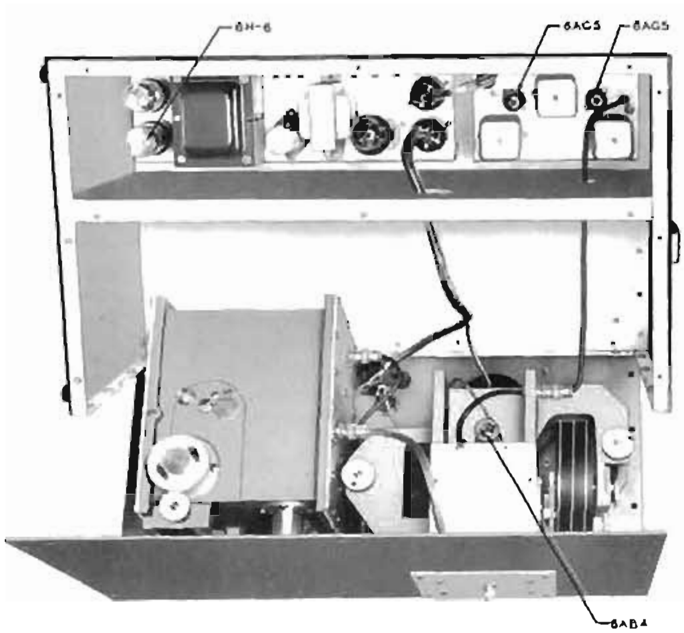


Figure 1. RX Meter showing location of Mixer, Ballast, and I.F. Amplifier tubes.

Operational problems with the RX Meter often can be traced to one of two sources:

Problem No. 1 concerns the mixer tube 6AB4.

Problem No. 2 concerns the ballast tube 6H-6.

Both problems are easily identified and corrected in the field. In most cases less time is needed to make the necessary repair than going through the formality of having a repair order issued!

Symptoms of a noisy 6AB4 mixer tube:

1. Difficult to balance the Rp dial.
2. Impossible to balance above 100 MC.
3. Insufficient range of fine and coarse balance controls.
4. Instability of the null indicator.
5. Increased null reading.

A quick check of the 6AB4 mixer tube is to tune the oscillator for maximum reading on the null indicator meter with the bridge unbalanced. Balance for minimum indication of null indicator. The reading may be 1 to 3 divisions or so. Switch the range switch out of detent. This will disable the oscillator.

If the needle does not fall to zero the 6AB4 is noisy. The amount indicated will reduce the sensitivity and accuracy of the Rp dial accordingly. The mixer tube 6AB4 is found in the bottom rear of the bridge casting which contains the Cp dial assembly. (See Figure 1.) It is advisable to use the same brand tube as found in the instrument and it may also be necessary to try several tubes before satisfactory results are obtained. The company supplies a specially selected 6AB4 tube, BRC #301637.

Incidentally, an increased null reading can also be caused by one or both 6AG5 tubes in the I.F. amplifier—V201 and V202—trying a few tubes and checking the null indication will tell the story.

Be sure to turn the instrument off when changing the I.F. tubes. V201 is in the regulated filament supply. Removal of this tube in operation will overload the 6H-6 ballast tube and also the two oscillator tubes.

Mention of the 6H-6 ballast tube recalls to mind that some instruments were returned for repair because someone replaced the 6H-6 with a 6H6 tube. In all cases replacement of the 6H6 with a 6H-6 was all that was necessary to put the RX Meter back in operation.

Symptoms of a faulty 6H-6 ballast tube are:

1. Oscillator ranges 6-7-8 intermittent or inoperative.
2. Oscillator output low as indicated by the peak reading of less than 30 or 35 on the null indicator.
3. Oscillator output will decrease at the low frequency end of the higher ranges.
4. Visual—any noticable kink in the 6H-6 filament.
5. Oscillator is completely inoperative.

To check voltages use voltage check points of figure (2) below.

Removal from Cabinet: The bridge and oscillator assemblies of the instrument are permanently fastened to the front panel and are removed from the cabinet as a unit. The power supply and amplifier are constructed on a separate chassis, located end-to-end in the rear section of the cabinet and fastened to the bottom of the cabinet by four screws each. All four major sub-assemblies are interconnected by cables with removable plugs.

A large portion of any required maintenance, such as replacement of tubes, may be accomplished by removing the front panel (with bridge and oscillator) and top panel together. This may be done as follows:

1. Remove all 12 black Phillips screws from the top panel. (Do not remove or loosen any of the screws on the terminal plate.)
2. Remove the four Phillips screws from each side of the front panel and the three from along the bottom edge.
3. The top and front panels may now be tilted forward from the cabinet to provide access to the interior of the instrument. If it is desired to remove them entirely, the plug connections on the internal cables to the power supply and amplifier must first be disconnected.

Be sure to replace at least two screws to fasten top panel and two screws to fasten front panel when rechecking operation of instrument after insertion of any new tube.

Two years have gone by since our first delivery of RX Meters. It is about time for some of the older instruments to develop the above symptoms, as the tubes by this time have weakened.

It is hoped that the above information may serve to cut down costly repair and shipping charges. The loss of time and valued use of the RX Meter 250-A in production and in projects is well recognized.

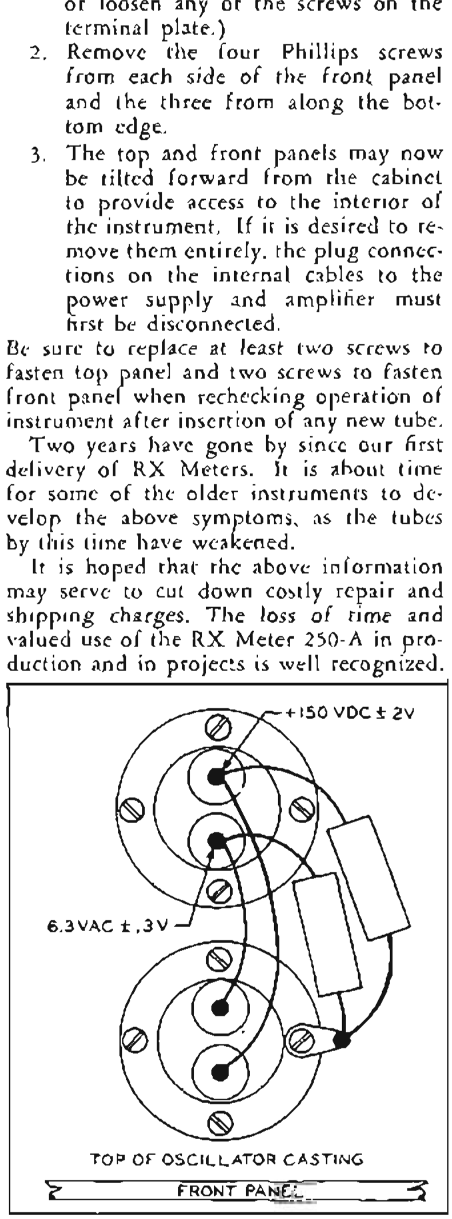
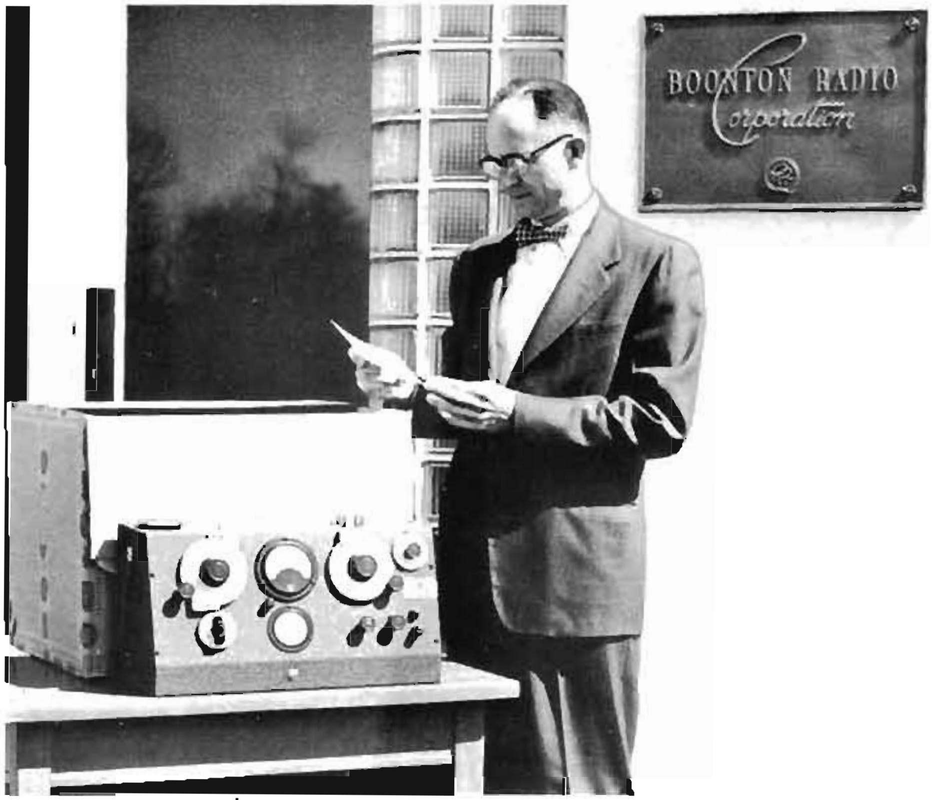


Figure 2. Voltage check chart

EDITOR'S NOTE

Due to a mechanical error Fig.2 and Fig.3 in the previous issue (Spring, 1955) were interchanged. Also, the text in the caption at the foot of column I on page 7 should read: "... was produced by the known low level output of the RF Voltage Standard." We shall try to avoid a repetition of this situation in the future.

Q METER AWARD



Dr. G. A. Downsbrough, President and General Manager of Boonton Radio Corporation, is shown above reviewing the card which he has just drawn to determine the winner of the Q Meter also shown in the picture. This Q Meter, displayed at our booth in the IRE Exhibits at Kingsbridge Armory in March, was to be awarded to one of our guests who completed a registration card at the booth.

The winning card was drawn from several thousand cards completed by our friends at the show and was signed by Mr. Waldemar Horizny, Technical Supervisor and Assistant Director of the Home Study Department, of RCA Institutes, Inc., of New York City. Mr. Horizny was born in Detroit, Michigan in 1921 and was awarded a BA degree from New York University in 1943. For a good number of years, he has been employed by RCA Institutes, Inc., a Service Company of Radio Corporation of America. Initially, he served as an instructor, and is now part of the Administrative Staff. His responsibilities entail supervising all technical operations of the Home Study Department, setting up courses of study, etc. His Department has prepared a Home Study Course in Television Servicing, and recently completed a Color Television Course The department is now preparing a general radio and electronics course.

Mr. Horizny, in accepting the award, commented "The Boonton Q Meter is no stranger, as it is no stranger to many other workers in the field. I have used the instrument in classroom instruction and for work related to the preparation of our courses."

Mr. Horizny lives at 138 Cypress Street in Floral Park, New York. He expects to use his Q Meter in his work for a period of time after which he will move it to his home laboratory.

We wish to thank all of the people who came by to see our exhibit at the IRE. We'll welcome you at future shows.

Printed in U.S.A.

FALL, 1955

Number 7

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Applications of a Sweep Signal Generator

FRANK G. MARBLE, *Vice President—Sales*

The design techniques used in the development of a new Sweep Frequency Signal Generator were discussed in the Spring 1955 Number 5 Issue of THE NOTEBOOK.* That discussion covered the methods used to obtain the performance required of a precision AM modulated Signal Generator, sweep and marker system in a single instrument. This article continues the discussion by considering some of the various methods by which such an instrument can be used.

One advantage of a sweep frequency signal generator lies in its ability to save time and thus economize engineering manpower, freeing it for other constructive work. One might, for instance, use an adjustable frequency, adjustable level cw signal generator to obtain output-vs-input data for an if amplifier at several discrete frequency points. This data can then be plotted on a graph showing response-vs-frequency to obtain the pass-band of the circuit. For each circuit readjustment this procedure must be repeated. For a narrow pass-band circuit this process is at best tedious, but for a broad-band circuit its time requirements are virtually prohibitive.

The simultaneous display of the response-vs-frequency curve of a circuit on the screen of a cathode ray oscilloscope by a sweep frequency signal generator system and the instantaneous indication of changes caused by adjustments expedites the engineer's work enormously. Another advantage of the sweep method is the practical fact that some of the time so saved will be used to obtain refinements which would have been overlooked using the slower single frequency methods.

"Sweep Frequency Signal Generator Design Techniques" John H. Mennie and Chi Lung Kang—The Notebook Spring 1955 Number 5.

YOU WILL FIND

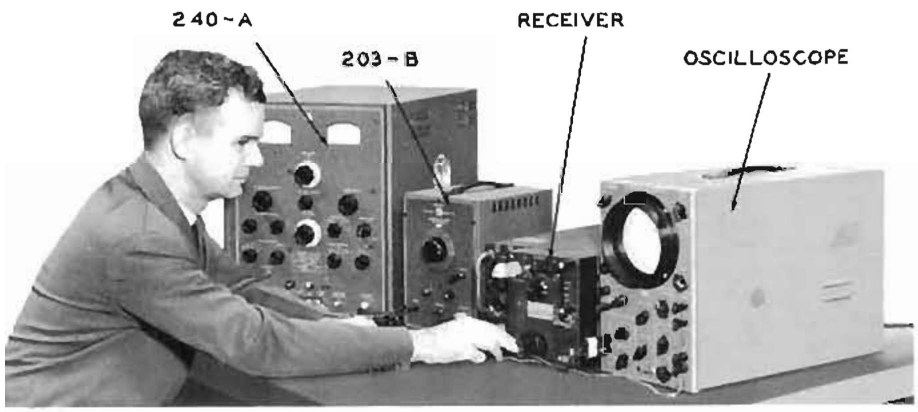


Figure 1. The author adjusts the pass-band of the IF amplifier below the 4.5 mc lower limit of the Sweep Signal Generator 240-A by using the Univerter Type 203-B. This combination permits cw and sweep measurements from 0.1 to 120 megacycles per second.

Besides the savings in time and the greater practical refinement obtained, some information is immediately observed by sweep methods which can be easily overlooked in the point by point method. Regeneration effects and "suck-outs" may cause changes in the response curve which persist over only a very narrow range of frequencies. Since cw measurements are made at discrete frequency points only, it is possible to obtain a smooth response curve excluding these effects if they happen to lie between the selected measurement points. A Sweep Signal Generator presents data which is continuous with frequency. This removes the possibility of missing important information.

The Basic Measuring System

Fig 2 shows the 240-A, a broad band detector, and an oscilloscope interconnected. The resultant information which appears on the screen of the oscilloscope is shown in enlarged form in the photograph in Figure 3. The display is a graph with abscissa proportional to frequency and ordinate proportional to the amplitude response of the detector circuit.

The interconnections in Figure 2 required to obtain the display include the connection of the rf output of the signal generator to the detector whose output connects to a marker adder circuit "Test In" in the Signal Generator and hence to the vertical deflection amplifier of the oscilloscope. The sawtooth of

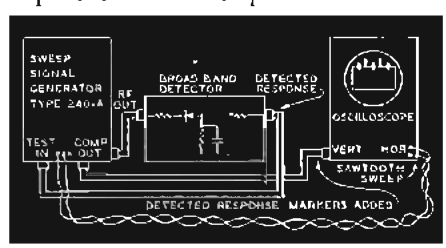


Figure 2. Typical interconnections of Sweep Signal Generator Type 240-A, Test Circuit (Broad Band Detector) and Oscilloscope.

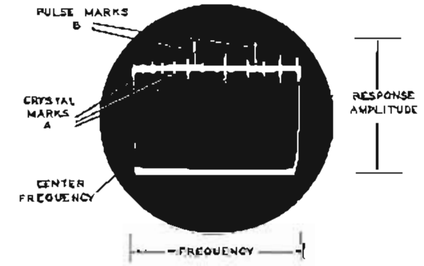


Figure 3. Enlarged photograph of the display appearing on the screen of the oscilloscope in figure 2.

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voltage which frequency modulates the constant-amplitude RF output of the Signal Generator is connected to the horizontal deflection amplifier of the oscilloscope.

The sawtooth voltage, while increasing in amplitude, modulates the constant amplitude RF output from a minimum to a maximum frequency. Simultaneously it deflects the oscilloscope from left to right. At the maximum frequency point the sawtooth starts decreasing in amplitude, the rf output of the signal generator is reduced to zero, the oscilloscope deflected from right to left and the tuning mechanism of the signal generator returned to the minimum frequency point. The constant amplitude rf output reappears and the cycle is repeated.

The lower line in the display in Figure 3 represents the reference line of the graph or line of zero input and the upper line the detector response curve.

The frequencies along the Abscissa must be identified if the response curve is to have meaning. In the Sweep Signal Generator Type 240-A, frequency identification is accomplished by two types of marks. The marks at (a) in Figure 3 appear at the harmonics of a crystal oscillating at 2.5, 0.5, or 0.1 megacycles. The marks indicated by the arrows marked (A) have separation of 2.5 megacycles. The center frequency can be identified from the tuning dial of the Signal Generator. With the center frequency identified the frequency of each of the other marks can be deduced since the frequency spacing is known.

The marks at (B) on Figure 3 are sharp narrow pulses. The position of these pulses can be adjusted along the frequency axis by

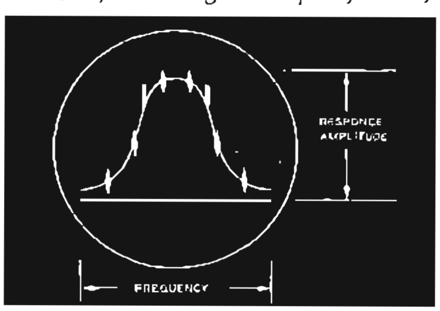

Figure 4. Oscilloscope display resulting from insertion of a selective circuit and associated detector in Sweep Signal Generator test circuit.

front panel controls. The crystal marks can be switched off after the pulses have been positioned to coincide with any two of these marks. This leaves these two frequencies marked in a manner which causes minimum interference with the reference curve. The pulses (B) can also be positioned between two crystal marks (A) since the frequency changes linearly with distance. The crystal marks can then be switched off. In this way, any two frequencies along the frequency axis can be marked.

The marks shown at (A) and (B) on Figure 3 are added to the display in the marker-adder circuit through which the signal from the detector (shown in Figure 2) passes before it is connected to the vertical deflection amplifier of the oscilloscope.

Determination of Selectivity and Sensitivity

The elements of a Sweep Signal Generator system for measuring selectivity and sensitivity of a test circuit are the same as shown in Figure 2 with the test circuit inserted between the RF output and the detector. If the test circuit contains a detector, the detector in Figure 2 can obviously be omitted. The resultant display appears in Figure 4. The constant amplitude signal source is frequency modulated or swept from a low to a higher frequency at a slow rate compared to the signal frequency. When the maximum frequency of the sweep is reached the signal source output is turned off and the generator returned to the low frequency point for a subsequent sweep from low to high frequency.

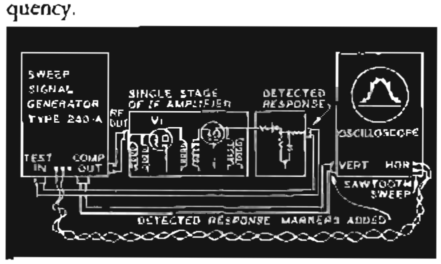

Figure 5. Interconnections for observation of pass band of a single stage within an IF amplifier.

The test circuit detector provides response curve of attenuation vs. frequency and frequency identification marks are added to the varying signal from the test circuit.

The horizontal deflection connections of the oscilloscope are connected to the same voltage that sweeps the signal source. The display on the oscilloscope, Figure 4, includes the response vs. frequency response of the test circuit, the frequency identification marks and a base or zero reference line indicating the level out of the test circuit with no input. The selectivity of the test circuit is apparent from a comparison of the change in response vs. the number of megacycles or kilocycles per inch along the horizontal axis of the display. This frequency calibration of the horizontal axis is deduced from the markers shown.

Selectivity usually varies with signal level as a result of AGC, limiters, non-linear amplifier, etc. Therefore it is important to test it at various operating signal levels. The Sweep Signal Generator Type 240-A, mentioned in the article cited in the first paragraph of this article, provides calibrated output level from 1.0 to 300,000 microvolts while sweeping. Its leakage is sufficiently low to permit use of an external 20 db attenuator to obtain outputs down to 0.1 microvolt.

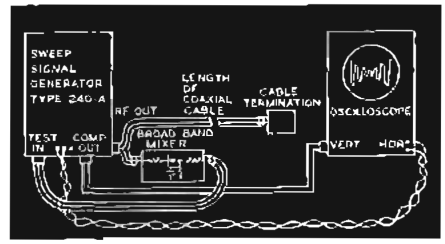

Figure 6. Interconnections for study of cable and cable termination characteristics.

Selectivity of Single Stages

The system of connection in Figure 2 is suitable for receivers, filters or amplifiers. The terminated rf cable (a 50 ohm system) is connected into the input of the test circuit. The detector of the test circuit is connected to the marker adder circuit in the sweep signal generator (input impedance 1 megohm). The use of a sweep signal source is not limited to complete receivers or amplifiers, however. So long as arrangements are made to avoid any effect on the sensitivity or selectivity of the circuit under test by the impedance of the rf output of the Sweep Signal Generator or of the detector the selectivity-sensitivity characteristics of any circuit may be observed within the sensitivity limits of the oscilloscope being used.

A convenient method of observing the pass band of a single stage appears in Figure 5. The output of the Sweep Signal Generator connects to the grid of the tube which contains the test stage in its plate circuit. A broad band detector is connected to the plate of the following stage through a coupling condenser. The low input impedance of the detector lowers the Q of the circuit in the plate of this tube so materially as to make its effect on the final result insignificant. The tuned circuit of tube V1 is operating under its normal condition and its sensitivity-selectivity characteristic can be observed on the oscilloscope.

Study of Pass Band Characteristics

The Q of the pass band of a test circuit can be approximately deduced by use of a sweep signal generator. As discussed above the response curve of a circuit can be displayed on an amplitude vs. frequency graph on the face of a cathode ray tube. The marking system of the Sweep Frequency Signal Generator makes it possible to identify any frequency along the horizontal axis. Since the response in the vertical direction on the oscilloscope is linear, a point 0.707 times the distance from the zero or reference line to the peak of the response curve can be located on each side of the peak. From the frequency marking system the frequency difference

(Δf) between these two points and the frequency of the peak can be obtained. Q can then be obtained from the following formula: $Q = \frac{f_o}{\Delta f}$

Adjustment of Stagger Tuned Circuits

Broad pass bands are often obtained by adjusting the resonant frequencies of successive single tuned circuits to slightly different frequencies within the desired pass band. The overall result is a relatively flat pass band broader in frequency than any one of the individual tuned circuits.

To adjust this type of amplifier, it is normally quicker to first resonate each individual circuit to the proper frequency with a cw signal generator. After completion of this procedure, the overall pass band configuration can be investigated and "touch up" adjustment made with a sweep signal generator. The Sweep Signal Generator Type 240-A is excellently suited to this procedure since it operates as a cw (with or without AM) or sweep signal generator, without the necessity of disturbing the input or output connections to the test circuit. A Vacuum Tube Voltmeter can be bridged across the input to the vertical deflection amplifier input of the oscilloscope for the single frequency work. The oscilloscope of course is used for the overall investigation and "touch up." Since the output monitoring and attenuation system is equally applicable to cw and sweep work the sensitivity can easily be checked under either condition.

Study of Cable Characteristics

The characteristics of high frequency cables may be investigated by use of a sweep frequency signal generator. In Figure 6 a sweep signal generator is shown connected to the input of a length of high frequency cable. Also connected to the input of the cable is a wide band detector. The low frequency sweep voltage from the sweep signal generator is connected to the horizontal deflection input of the oscilloscope. The RF signal, swept or frequency modulated at a low rate of 60 times per second, is fed into the cable. Reflected signal from imperfections in the cable or the termination arrives back at the

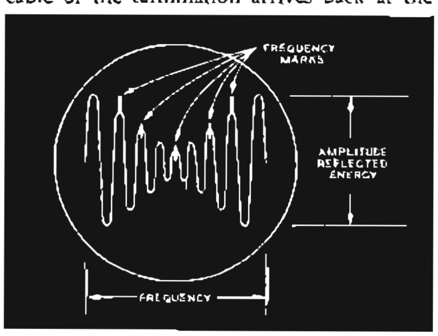

Figure 7. Oscilloscope display indicating amplitude of reflected energy from a termination coaxial cable.

input a finite time later. Since during this finite time the input signal has changed to a new frequency, an audio difference frequency (input frequency minus reflected frequency) appears across the output of the detector. The amplitude of the input signal is great and constant and the reflected frequency amplitude for a near match is small and variable. The amount of energy reflected from the end of the line depends on the correctness of the termination and varies from zero for a perfect match to a finite value proportional to the mismatch for mismatched lines. Since the termination impedance will, in general, vary with frequency, the amount of energy reflected will also vary. The audio frequency from the detector appears on the oscilloscope The envelope amplitude of the display is proportional to the instantaneous reflected signal and the abscissa is proportional to frequency of RF input as shown in Figure 7.

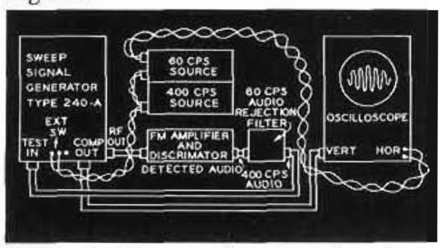

Figure 8. Diagram of equipment and connections for measurement of linearity of FM discriminator.

With a perfect termination over the frequency range in question, various cables can be observed for imperfections in construction. A periodic variation in dielectric constant of the cable insulation will exhibit itself on the oscilloscope display.

Adjustable resistance load will permit quick determination of the Z_0 for long cable lengths.

The Linearity of FM Discriminators

The Sweep Signal Generator Type 240-A provides a powerful method of determining the linearity of an FM discriminator. The method is indicated in Figure 8. A low frequency (60 cps) sweep, adjusted to sweep the full frequency range of the discriminator, and a higher frequency sweep (400 cps) is fed into the EXT sweep input of the Sweep Signal Generator. The high frequency voltage is adjusted to sweep only a small fraction of the frequency range of the discriminator. In effect the high frequency sweep explores the slope of each section of the discriminator while it is slowly moved from section to section by the low frequency sweep. The output is detected and passed through a high pass filter which passes only the resulting 400 cps note. The display of the amplitude of this note vs. the low frequency sweep affords a visual display in which the slope of the amplitude of the envelope of the 400 cps note is proportional to FM discriminator linearity. A constant amplitude indicates a linear discriminator whereas a varying amplitude indicates a variation in linearity.

The Study of Crystal Modes

The rapid location of the several frequency modes at which a crystal oscillates is important but tedious by discrete frequency methods. The Sweep Signal Generator Type 240-A provides a frequency sweep on which individual frequencies can be identified by the marker system included in the generator. A crystal, however, has such a high Q that sweep rates must be very low to prevent ringing and spurious responses. By using an oscilloscope with low frequency sawtooth sweep available, the 240-A can be swept at frequencies of 1 or 2 cps by connecting the oscilloscope sweep output to the "External

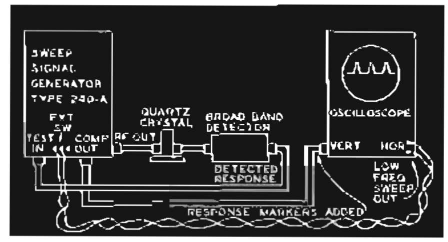

Figure 9. Equipment arrangement for measurement of quartz crystal characteristics .

Sweep" of the 240-A. The system is then connected as shown in Figure 9. By varying the center frequency of the 240-A and its sweep width the crystal can be explored for responses over a considerable frequency range.

Extension of the Frequency Range

The lowest center frequency of the Sweep Frequency Signal Generator Type 240-A is 4.5 megacycles. At this frequency the sweep frequency capabilities of the instrument are $\pm 1\%$ to $\pm 30\%$ of center frequency or ± 45 KC to ± 1.35 MC. For applications in television video amplifiers both for color and black and white, and aircraft navigation receiver intermediate frequency amplifiers, lower center frequencies and/or broader sweeps are required. Both these requirements can be met by use of the Univerter Type 203-B with the Sweep Frequency Signal Generator Type 240-A. The 203-B consists of a broad band mixer with local oscillator at 70 MC. followed by a broad band amplifier with a 50 ohm output. The gain of the 203-B is set at unity. Figure 1 shows the 240-A, 203-B in a measuring set-up. In use the 240-A is tuned to a frequency equal to 70 MC plus the desired output center frequency from the system. Sweeps from ± 0.7 MC to ± 15 MC are available. Single frequency outputs unmodulated or with AM modulation can be obtained. Thus single frequency or sweep outputs are made available over the band width of the 203-B which is 0.1 to 25 MC.

Summary

The Sweep Signal Generator is a powerful tool of considerable flexibility. It not only saves considerable time but makes refinements possible in circuit adjustment and development which would not normally be possible.

THE AUTHOR

Frank G. Marble's career covers a broad field of engineering experience: design & development work for Philco; coordinator on various government projects; two years with Western Electric's electrical research division & engineering administrative posts with Pratt & Whitney Aircraft & Kay Electrical Co. Mr. Marble has been with Boonton Radio since 1951 & a Vice-president—sales since 1954. Mr. Marble has a BS in EE (Mississippi State College 1934) & an MS in EE (M.I.T. 1935).

Use of the RF Voltage Standard Type 245-A

W. C. MOORE, *Engineering Manager*

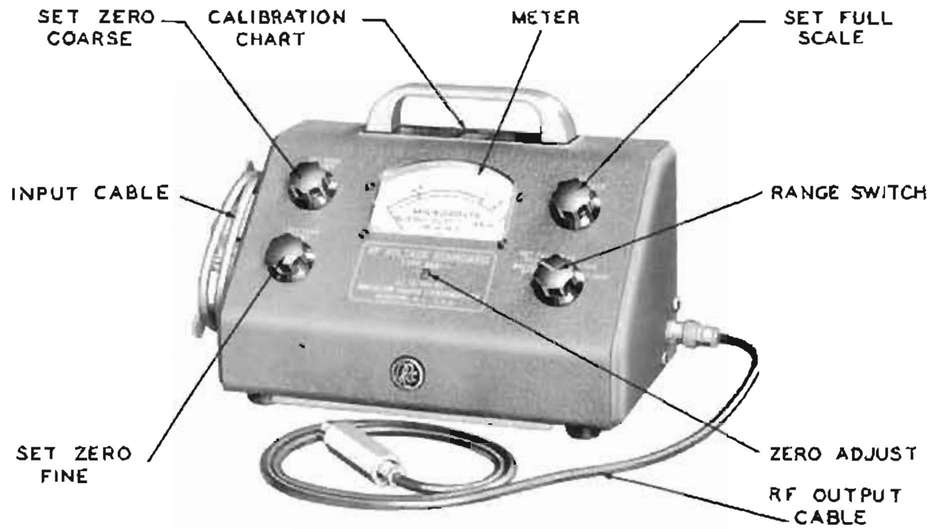

Figure 1. RF Voltage Standard Type 245–A.

When discrepancies exist among measurements made with different signal generators on the same radio receiver, it is often very difficult to determine just which of the instruments is performing correctly. The introduction of a reference standard usually will resolve the dilemma so that effort can profitably be applied to the offending units. However, care must be exercised in the use of such a standard and understanding applied to the interpretation of the results. This article discusses the use of a source of standardized voltage at a known impedance.

The R.F. Voltage Standard Type 245-A, shown in Figure 1, is designed to deliver, across the BNC output jack at the end of its Output Cable, open-circuit radio frequency voltages of ½, 1, and 2 microvolts through a source impedance of 50 ohms over the frequency range of 0.1 mc to 500 mc. It can be used in conjunction with a signal generator as a source of known voltage and impedance for determining receiver sensitivity performance. Using this source of voltage as a point of reference, it is also possible to perform relative comparisons with other sources of radio frequency voltage whose frequencies lie within the specified range. In addition, the input system is calibrated for use as a 50 ohm rf voltmeter at 0.05 volts over a wide frequency range.

Principle of Operation

A description of the RF Voltage Standard is given in the Spring, 1955 issue of the BRC Notebook[1]. The system block diagram of the RF Voltage Standard is shown in Figure 2. An external source is used to supply rf voltage to the Input Cable. The voltage at the output end of this cable is indicated by an RF Voltmeter at the point where the cable is terminated by the input to the Coaxial Attenuator[2]. The low voltage output of the Coaxial Attenuator appears in series with an impedance matching resistor.

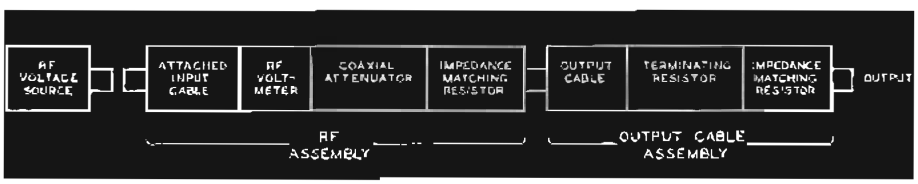

Figure 2. RF Voltage Standard System Block Diagram.

The low voltage output from the RF Assembly, which presents a source impedance of 50 ohms, drives the 50 ohm Output Cable which is terminated by a 50 ohm coaxial terminating resistor. The terminating resistor is followed by a 25 ohm impedance matching resistor to raise the equivalent source impedance at the end of the output cable to 50 ohms.

Output Voltage Calibration

Figure 3 shows the distribution of voltages throughout the instrument when the meter is set at the 1 microvolt level and there is no external load connected to the output cable. An input level of 0.05 volts is established across the input to the coaxial attenuator by adjusting the voltage output of the external rf voltage source until the indicating meter on the RF Voltage Standard reads at the 1 microvolt calibration point on the meter scale with the range switch set to 1 microvolt. The 25,000:1 coaxial attenuator divides the 0.05 volts down to 2 microvolts which appear across the 0.0024 ohm resistor in series with the 50 ohm impedance matching resistor in the rf assembly.

Since the 50 ohm characteristic impedance of the Output Cable is matched by the 50 ohm resistor, its length is electrically indeterminant. In fact, its length may even be considered zero, and the 50 ohm terminating resistor is effectively connected to ground directly from the 50 ohm impedance matching resistor in the RF Assembly. This divides down the 2 microvolts delivered by the Coxial Attenuator to 1 microvolt across the 50 ohm terminating resistor.

The meter on the RF Voltage Standard is calibrated in terms of the open circuit voltage appearing across the BNC output jack on the output cable, with no load connected to the cable.

Output Impedance

The output system of the RF Voltage Standard is based on a 50 ohm characteristic impedance. The optimum conditions for power transfer and control of voltage standing waves on the cable as the load impedance is varied are described in the Fall 1954 issue of the BRC Notebook[3].

The output impedance of the RF Assembly is 50 ohms, determined by the 50 ohm impedance matching resistor, which is the termination of a specially designed section of coaxial transmission line.[1, 2] Looking back along the coaxial output cable from the 50 ohm terminating resistor toward the RF Assembly, one sees the 50 ohm characteristic impedance of the cable in shunt across the 50 ohm terminating resistor. The net result of this parallel combination is 25 ohms of resistance which is then built up to the desired 50 ohms by the 25 ohm series Impedance matching resistor located between the terminating resistor and the BNC output jack.

The open circuit output impedance at the output jack on the Output Cable is 50 ohms.

Measuring Receiver Sensitivity

The sensitivity of a radio receiver has been defined by the Institute of Radio Engineers[4] as the number of microvolts required to produce standard output when applied to the dummy antenna in series with the input impedance of the receiver. For a system consisting of a 50 ohm transmission line system and a 50 ohm receiver, this means that a "1 microvolt receiver" will produce standard output when 1 microvolt is applied across the series combination of the 50 ohms antenna impedance and the 50 ohm input impedance of the receiver. This yields ½ microvolt across the receiver input terminals.

Figure 4 shows how this condition is met by the voltage calibration and output impedance characteristics of the RF Voltage Standard Type 245-A. The actual circuit can be reduced to a schematic circuit because the characteristic impedance of the cable is matched at the voltage source as described above. The diagrams show the distribution of voltages and impedances along the circuits for the loaded and open circuit conditions when the meter indicates 1 microvolt. At this level setting, the circuit is being driven by 2 microvolts out of the Coaxial Attenuator.

The equivalent circuit diagrams show that the same loaded and open circuit characteristics of voltage and impedance will be presented to the load if we assume a simple series circuit consisting of a 1 microvolt generator in series with 50 ohms. This result could have been obtained directly by an application of Thevenin's Theorum to the original circuit. Additional diagrams and explanatory information can be found in the Instruction Manuals for BRC Signal Generator Types 202-B and 211-A, and Univerter Type 207-A.

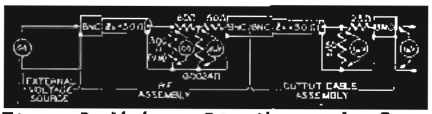

Figure 3. Voltage Distribution for Open Circuit Voltage of 1μV at End of Output Cable.

The sensitivity of a receiver designed to work with a 50 ohm antenna line impedance can therefore be read directly from the meter at ½, 1, and 2 microvolts because the equivalent source impedance of the RF Voltage Standard provides the 50 ohms to which ½, 1, or 2 microvolts are applied.

If higher values of antenna resistance are involved, direct readings of receiver sensitivity can be obtained by merely adding in series with the output cable a suitably-mounted, non-reactive resistor whose resistance is equal to the desired antenna resistance minus the 50 ohms already presented by the RF Voltage Standard. For example: to read directly the sensitivity of a receiver designed to work from a 75 ohm line, such as RG-11/U, a 25 ohm resistor must be added in series with the inner conductor at the BNC output jack on the output cable to obtain the correct impedance match. If values of antenna resistance less than 50 ohms are involved, it is necessary to use an impedance matching pad and allow for its insertion loss.

Checking Signal Generator Output

The use of the RF Voltage Standard to check the output from a signal generator is based on using a receiver as an un-calibrated transfer indicator to compare the outputs from the two sources at a fixed signal level. Figure 5 shows the steps for the case of a signal generator having 50 ohms output impedance at the output jack.

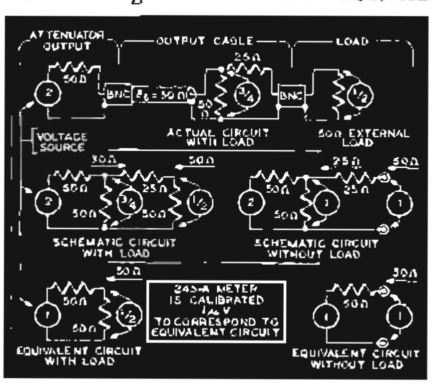

Figure 4. Derivation of Equivalent Circuit of RF Voltage Standard Output System Assuming a Matched Load & 1μV Setting.

The method shown in Figure 5, in which the same Output Cable is switched from the RF Voltage Standard to the signal generator output jack is valid *only* for signal generators having a 50 ohm source impedance at the panel output jack.

Some signal generators, however, present 50 ohms only at the output end of their own special 50 ohm terminated output cable. In this case, the receiver input must be transferred between the terminals of the output cable on the signal generator and the output cable on the RF Voltage Standard. Only in this way will the comparison show up standing wave errors in the signal generator output system.

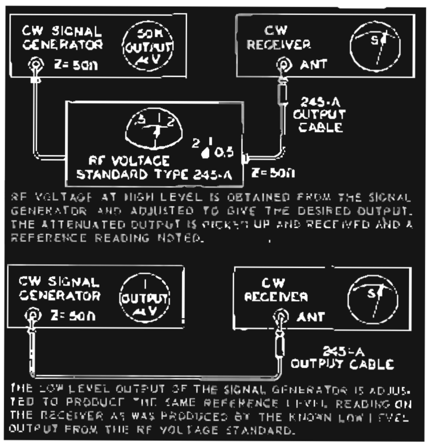

Figure 5. Comparison of Voltage Output from a 50 Ohm Signal Generator with the RF Voltage Standard, Using a Receiver as an Uncalibrated Transfer Indicator.

In case the signal generator has a source impedance of 50 ohms, it is not necessary that the receiver input impedance be matched to the signal generator output impedance to obtain a valid comparative reading. Since the two sources of voltage present the same impedance, it is necessary only that the receiver input impedance remains constant, at whatever value it may have, throughout the comparison process. For this reason, *only* the signal generator frequency can be changed to peak the receiver response, since small changes in receiver tuning may result in appreciable changes in input impedance.

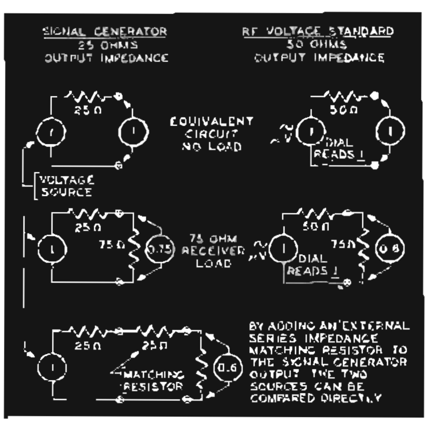

Figure 6. Comparison of Voltage Output from Un-equal Source Impedances by Addition of an External Impedance Matching Resistor.

An amplitude modulated signal can be used with an AM receiver and an audio voltmeter, provided the amplitude modulation is kept below 30%.

Unequal Source Impedance

The problems of interpreting signal generator output readings increase when checking the calibration accuracy of a signal generator whose output impedance cannot be made the same as the reference standard by suitable resistive pads, as shown in Figure 6, or whose output cable system sets up standing waves at critical frequencies. These same problems arise in the use of such a signal generator for receiver sensitivity measurements. The necessary information to make these corrections is given in some detail in catalogues and instruction manuals by the major manufacturers of signal generators.

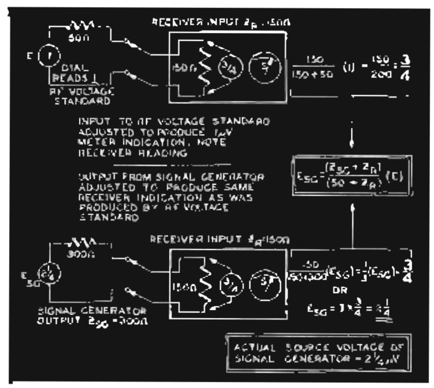

Figure 7. Signal Generator Calibration when all Three Impedances are Different.

Figure 7 shows a case in which the impedances of the RF Voltage Standard, the signal generator and the receiver are 50 ohms, 300 ohms and 150 ohms respectively. The general equation shown in the figure gives the number of microvolts actually delivered by the signal generator for any combination of impedances in terms of the indicated output level of the RF Voltage Standard.

The presence of standing waves in the output system of a signal generator which is not matched internally will produce errors in calibration which must be corrected by using data supplied in the signal generator instruction manual. These errors are a function of frequency and must be taken into account at each frequency setting.

In summary:

1. Determine the output impedance characteristics of the signal generator being calibrated.
2. Attempt to modify it to 50 ohms by the use of pads or dummy antenna systems, taking into account their effect on the calibration dues to attenuation characteristics.

3. If the output impedance cannot be made 50 ohms, determine the complex impedance of *both* the receiver and the signal generator and calculate the resulting voltage divider. Also calculate the voltage divider consisting of the receiver and the 50 ohm impedance of the RF Voltage Standard.

4. Since the outputs of the two voltage dividers are equal when the signal generator output is adjusted to give the same receiver reading as the RF Voltage Standard, we can equate the two expressions as follows: $\frac{Z_r}{Z_r + Z_{sg}}(E_{sg}) = \frac{Z_r}{Z_r + 50}(E)$, where

Z_r = receiver input impedance
Z_{sg} = signal generator output impedance
50 = RF Voltage Standard output impedance
E_{sg} = signal generator open circuit voltage
E = RF Voltage Standard open circuit voltage

Then the signal generator setting, Esg, which will produce the same receiver response as the output of the RF Voltage Standard, E, can be determined from the equation: $E_{sg} = \frac{Z_r + Z_{sg}}{Z_r + 50}(E)$

Use As A 50 OHM RF Voltmeter

The input system of the RF Voltage Standard is shown in Figure 8. It contains a length of coaxial cable which connects the source of power to the coaxial "head," which consists of a diode voltmeter in parallel with the input to the precision coaxial attenuator. The diode voltmeter reads the input voltage directly at the input to the attenuator, and the calibration of the RF Voltage Standard is not affected by standing waves on the cable ahead of this point.

The 60 ohm attenuator input impedance is shunted by approximately 300 ohms diode impedance, which together form approximately a 50 ohm termination for the 50 ohm input cable. The voltage seen by the diode voltmeter at the input to the attenuator will be nearly the same as that applied at the input BNC connector, subject to voltage standing waves on the cable. Variations in the characteristic impedance of the cable and the diode impedance introduce a moderate standing wave of voltage on the cable which increases with frequency.

The ratio for each coaxial attenuator is individually determined, and the correct input voltage for the 1 microvolt level meter setting is given on the voltmeter calibration data plate on top of the instrument. This information can be used for checking the instrument at low frequencies (below 500 kc) and for measuring rf input voltages. With the range switch in the 1 microvolt position, the input voltage is increased until the meter indicates 1 microvolt. The input voltage is then equal to the value stamped on the data plate. Input voltages of ½ and 2 times this value can be determined by adjusting the input for meter indications of 0.5 and 2 with the range switch in the corresponding position.

Accuracy

The method of setting up the calibration of the RF Voltage Standard at the factory is such that the initial accuracy is determined by the care with which the 60 ohm and the 0.0024 ohm resistors are measured, the accuracy of the voltage source used to set up the rf voltmeter circuit, and the Voltage Standing Wave Ratios (VSWR) of the input to the coaxial attenuator and the impedance match of the output cable termination. Of these, the VSWR is the least accurate measurement and also the greatest contributor to the overall tolerance.

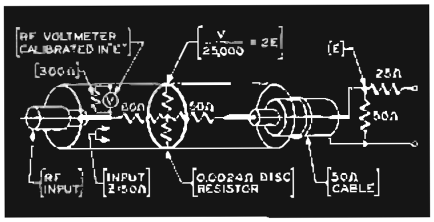

Figure 8. RF Attenuator and Voltmeter.

The 60 ohm film resistor is the center conductor of a terminated transmission line[2] and together with the 0.0024 ohm disc resistor it provides an accurate attenuator useful over a very wide range of frequencies[6]. The actual ratio is taken into account in setting up the voltage into the attenuator and adjusting the meter to read the desired output voltage. The uniformity with frequency of the attenuation ratio is determined by comparing each unit against a carefully measured standard unit at several points over a wide frequency range.

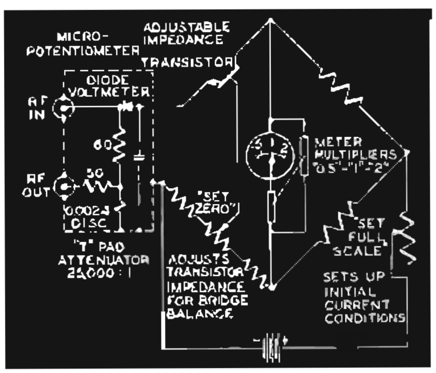

Figure 9. RF Voltage Standard-Basic Circuit.

The long term accuracy, which is of considerably more importance and upon which the specifications are based, includes the stability of several components not involved in the initial calibration. A circuit has been chosen in which these variations are minimized by the procedure used to place the instrument in operation.

The simplified circuit of Figure 9 shows the basic dc metering system associated with the rf voltmeter. The rectification efficiency, or ratio of rectified dc current to applied ac voltage, of a semi-conductor diode at a controlled value of bias current is a very stable characteristic.

The transistor is used in conjunction with the diode to raise the impedance level presented to the meter for proper damping. The diode current passes through the junction transistor with a constant efficiency of about 98% regardless of resistance changes. This current transfer factor, known as "alpha," is very stable and therefore does not contribute any significant variation in accuracy. The action is somewhat analagous to the unity voltage gain characteristic of a cathode follower circuit which also presents a large impedance ratio between input and output circuits.

As seen in Figure 9, the transistor impedance is located in one arm of a bridge. Hence the bridge can be brought to balance by varying the transistor impedance by means of its base voltage. This is done during the initial adjustment procedure with the SET ZERO control. This does not affect the 98% efficiency of current flow through the transistor.

Precautions

Several points of technique in handling low-level radio frequencies become of particular importance when checking the calibration of a signal generator. RF voltage leakage out of the signal generator, sometimes along the power cord, will cause trouble if the receiver is not well shielded. Likewise, interfering signals from adjacent equipment or broadcast transmitters will affect poorly shielded receivers and prevent accurate measurements.

The conditions of impedance match and corrections for standing waves on output cables must be accounted for before the performance of a signal generator can be evaluated. The connections between the output cable from the RF Voltage Standard and the signal generator to the receiver input should be as short as possible. The insertion loss of any matching pads must be included in the comparison.

Sharp receiver response will cause critical tuning and stability problems. and will pass only the low frequency components of the noise which make the meter bounce. A wider pass-band will produce a higher, but much steadier, noise level to which the desired signal is added.

Tune only the signal generator when searching for maximum receiver response to avoid changes in the receiver input conditions.

Always check the signal generator tuning when going from the condition of high level into the RF Voltage Standard to the low level into the receiver. It is sometimes advisable to re-tune the signal generator frequency each time the low level output is re-adjusted in order to get significant results.

When first placing the RF Voltage Standard in operation, it is advisable to re-check the SET FULL SCALE and SET ZERO positions. Initial drift can be caused by changes in battery voltage when the instrument is first turned on and by changes in the resistance of the transistor due to a sudden change in temperature, such as bringing the instrument from storage into a warm laboratory. There is no significant heat developed inside of .the instrument. Re-adjusting the SET FULL SCALE and the SET ZERO controls restores the calibration accuracy of the instrument even though the transistor and diode dc resistances may have changed.

Summary

By judicious use of the RF Voltage Standard Type 245-A it is possible to check the low and high level calibration of signal generators over a wide range of frequencies, and to establish signals for testing receivers at the microvolt level with a confidence not formerly possible.

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Calibration of the Internal Resonating Capacitor of the Q Meter

SAMUEL WALTERS, *Editor, The Notebook*

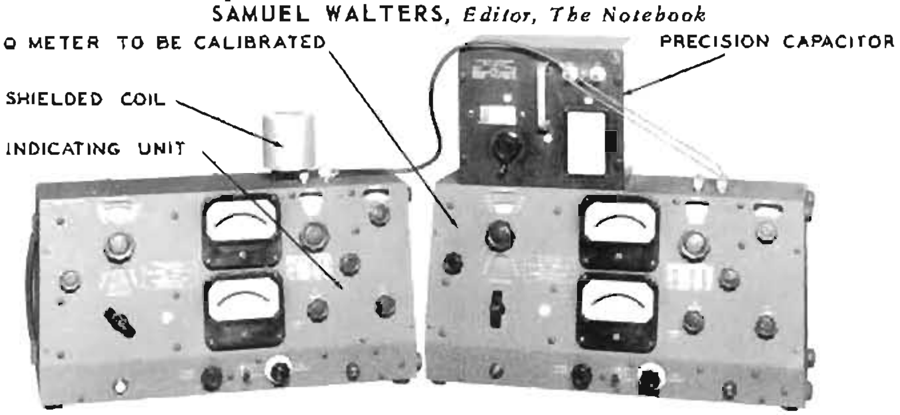

Figure 1. Interconnections of equipment that can be used in the calibration of the Internal resonating capacitor of a Q Meter. Here shown are Q Meters Type 260-A and a GR precision capacitor Type 722-D.

Recently we have received a number of inquiries on this subject. They are numerous enough to indicate a wide-spread interest in the technique of calibrating the Internal Resonating Capacitor of the Q Meter. This interest is understandable since the versatility of the Q Meter in performing a host of functions besides measuring Q* depends, in some special cases, on the additional accuracy obtainable from an error curve for the Internal Resonating Capacitor.

The Q Meter contains (1) an RF oscillator, (2) a measuring circuit including the main and vernier tuning capacitors (Internal Resonating Capacitor), (3) a vacuum tube voltmeter and (4) a system for injecting a known amount of the oscillator voltage in series in the measuring circuit.

The Internal Resonating Capacitor is used to adjust the value of capacitance so that the circuit under test can be resonated at the measurement frequency. Calibration of this capacitor should be done at a relatively low frequency with respect to the instrument's operating range in order to prevent stray inductance effects.

The calibration method described here is based on substitution of a known amount of capacitance from a precision capacitor for an indicated amount of capacitance in the Q Meter, using a resonant circuit on a second Q Meter for the comparison.

Equipment Required

1 Q Meter to be calibrated—BRC 160-A or 260-A (referred to as No. 1).

1 Q Meter (BRC 160-A or 260-A) used as an Indicating unit (referred to as No. 2).

1 Precision Capacitor with a range covering at least 600$\mu\mu f$ (G.R. 722 or equivalent).

1 Shielded Coil that will resonate between 200-500 KC.

Preliminary Check

Before beginning calibration it is advisable to inspect the Internal Resonating Capacitor to be calibrated. A quick check of the following points may save a needless repetition of calibration and avoid waste of time since the instrument can not be calibrated properly if any of these mechanical conditions prevail:

(1) Examine capacitor for foreign matter, specks of dirt, etc. Such matter tends to lower the Q of the capacitor by introducing a spurious resistance across it.

(2) Check main bearing of rotor sections of both main and vernier capacitors. Shafts should be firm to prevent mechanical backlash or electrical instability.

(3) Check spring gear take-up of both capacitors. Improper loading of gears will also cause backlash.

(4) Make certain rotors are centered with stators. Check plate spacing visually. Run out rotor plates to notice any wobble.

Procedure of Calibration

A. Calibration of main Q Capacitor.

(1) Set the Q Meter, No. 2, which is to be used as the resonance indicator, to 450$\mu\mu f$ and turn on the power. Mount on the instrument a suitably shielded coil that will resonate between 200 kc and 500 kc such as the 103-A32.

(2) Connect the precision capacitor to the Hi and Gnd terminals of the indicating Q Meter through a short piece of coaxial cable. Now set the capacitor in the indicating Q Meter, No. 2, to the minimum value of 30$\mu\mu f$.

(3) Connect the grounded terminal of the precision capacitor to the Gnd terminal on the Q Meter being calibrated (No 1) with a No. 18 stranded copper wire. Arrange another lead from the insulated terminal on the precision capacitor to a point in air 3/8" to 1/2" above the Hi capacitor terminal post of the Q Meter being calibrated (No. 1) using a no. 20 AWG bare tinned signal conductor copper wire. The tip of this self-suspended lead must be straight, without hooks or loops, and must point down to the Q Meter terminal. Isolate this lead from surrounding objects.

(4) Set the precision capacitor to 600-$\mu\mu f$ or more. Now adjust the oscillator frequency control of the indicating Q Meter, No. 2, for a maximum indication of Q. Resonance will occur at a lower frequency than in step 1. Note the setting of the precision capacitor, calling the reading C_1.

(5) Set the main capacitor dial of the Q Meter being calibrated (No. 1) to 30 and the vernier dial to zero. Do not energize this Q Meter.

(6) Touch the suspended lead, moving it as little as possible, to the Hi terminal post on the Q Meter being calibrated and re-resonate with the precision capacitor. Note this reading as C_2.

The difference between the two recorded readings, C_1-C_2, plus 0.15$\mu\mu f$ is the true capacitance corresponding to a dial reading of 30†.

(7) Using the reading noted in step 4 as C_1, other values of capacitance on the

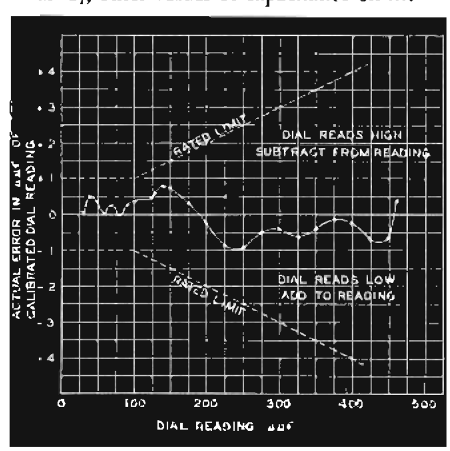

Figure 2. Correction Chart.

Q Meter main capacitor can be checked by successive settings of the unknown capacitor and the precision capacitor as above to obtain new values of C_2.

B. Calibration of Vernier Tuning Capacitor

The same procedure is followed as above except that the range of the precision capacitor must be expanded to obtain greater accuracy of calibration. The main tuning capacitor is left at $30\mu\mu f$, and the vernier capacitor is moved successively from 0 to +1, +2, +3, and −1, −2, −3. The amount of change on the precision capacitor dial necessitated in each case for resonance on the indicating Q Meter represents the corresponding value on the vernier capacitor.

By substracting the calibrated values from the dial readings and plotting the errors against the dial readings as shown in Figure 2, a calibration chart for the main capacitor can be drawn up. It is possible through this method of calibration to obtain an error curve which permits use at an accuracy somewhat better than our specified tolerance‡, depending of course on the skill of the operator and the accuracy of calibration of the precision capacitor used. A similar but expanded chart (since the actual error will be in tenths) can be drawn for the vernier capacitor.

* *See lead article in Winter, 1955 issue of Notebook on "A Versatile Instrument—The The Q Meter" by L. O. Cook.*

† *The VTVM adds about 0.15μμf when the meter is energized for normal Q Meter operation.*

‡ *Specified accuracy is plus or minus 1μμf from 30 to 100 μμf and plus or minus 1% above 100 μμf.*

A NOTE FROM THE EDITOR

We have noticed, as the publication date for each issue of THE NOTEBOOK draws near, that members of our engineering department pause when passing the editorial sanctum on the way to the water cooler and gape over our shoulder at the three-inch layer of chaos spread over the desk. This we charitably attribute to the engineer's curiosity concerning the mysterious journey of The Notebook to the printed page, (rather than wonderment as to why anyone would get paid for doing that sort of thing), and we feel that our readers, members of the same genus, might also be curious, if not in the mechanical process of preparation of THE NOTEBOOK, then certainly in the mystery of why another group of engineers should be so interested.

Dispensing with a description of the blood, sweat and tears generated by the authors in the course of their creative labors (many of our readers are painfully familiar with the picture), we will begin the journey at the point where the copy is ready for typesetting. The type for THE NOTEBOOK is "set" by a monstrous machine called a Linotype, which spews castings, or slugs, each of which corresponds to a line of type. This machine also has the ability to make an even right hand margin by regulaitng the spacing between individual letters and words.

When the copy has been linotyped and edited as carefully as time and the human factor permit, it is cut up and pasted in page form on large sheets of paper. The larger type used for headings is set by hand, using commercially available pads of paper letters. Glossy photostats of the line drawings are also pasted in position.

When our eight "repro" pages are ready, we take them to the offset printer, who proceeds to photograph them with a camera which is roughly half the size of a master bedroom. In this photographic process he reduces the size of our repro pages, which have been arbitrarily set up 10% larger than the final page size. The developed negatives are then placed over a light box and all extraneous lines, marks and paste-up details picked up by the camera are removed by means of opaquing fluid. The negatives are then carefully laid out in two rows of four each on a large sheet of paper, each page having a special position with respect to the others. This operation is called "stripping." A large plate of thin, sensitized metal is then exposed to light through this bank of negatives. When the plate is "developed" the exposed areas retain a greasy substance which attracts and holds printing ink. The plate is now mounted on the cylinder of the offset press and rotated, first against ink rollers which deposit ink only on the greasy areas, then against a large rubber roller which, in turn, transfers the ink to the paper passing through the press.

Each sheet of paper, when printed on both sides in what the printer terms a "work and turn" sequence, contains two complete copies of THE NOTEBOOK. These sheets are then fed into a folding machine which simultaneously folds and creates the glued binding. There remain only to cut, trim and punch the loose-leaf holes and our NOTEBOOK is ready for shipment.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Circuit Effects On Q

CHI LUNG KANG, *Development Engineer*

The Q of a practical simple resonant circuit is always lower than that of the component coil or capacitor because of additional losses in the circuit which often appear quite unexpectedly. In a measuring circuit, as used in a Q Meter, small internal losses are always present, whose significance is often not fully realized. Under the general heading of circuit effects on Q, this article points out how the loading accumulates in some practical circuits and examines the appreciable effect of residual parameters in Q Meter circuits on Q-readings obtained. Due to the effect of differences in residual parameters, Q-readings of the same coil but from different Q Meters may differ. Correlation of results between the low frequency Q Meter Type 260-A and the high frequency Q Meter Type 190-A in overlapping ranges is presented.

About Simple Resonant Circuits

A few assertions will be made about the simple resonant circuit to serve as a starting point for later discussion.

For a reactive component, either capacitive or inductive, if the Q is greater than 10, the following transformation is valid as shown in Figure 1.

$$X_s = X_p = X$$

$$Q = \frac{X_s}{R_s} = \frac{R_p}{X_p} \quad \text{i.e.} \quad R_s R_p = X^2$$

It is customary to talk about shunt loss or series loss in either a coil or a capacitor. But as long as either of them is considered as a two terminal component, it is simply an impedance and can be expressed either in the series or the shunt form as shown above. The following transformation illustrates this, $Q > 10$ being assumed:

Figure 1. Series and parallel forms of impedance.

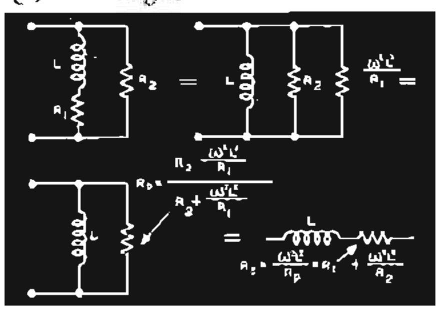

Figure 2. Different circuit representations for a coil.

The extent to which any change of loss affects the Q depends upon the loss already present. A resonant circuit of reactance X and quality factor Q has

$$R_s = \text{series resistance} = \frac{X}{Q}$$

$$R_p = \text{shunt resistance} = QX$$

Consider a 250μh coil which resonates with about 100$\mu\mu$f at 1 mc. Assume Q of the whole circuit is 320.

$$X = \omega L = \frac{1}{\omega C} = 1600\,\Omega$$

$$\therefore \quad R_s = \frac{1600}{320} = 5\,\Omega$$

$$R_p = 320 \times 1600 = 512{,}000\,\Omega$$

From these figures, it can be reasoned, for example, that any change of 0.02Ω in series resistance would be of little consequence but any additional shunt load of 5 megohms would have appreciable effect.

If a certain Q value is implicitly assumed, then the magnitude of X is itself an indication of impedance level, which is

$$R_s = \frac{X}{Q}$$

in the series case and

$$R_p = QX$$

in the parallel case. Thus at a fixed frequency, low resonating capacitance means a high impedance level. Consequently, at a low C, a shunt loss will have a great effect on Q while the effect of an additional series loss will be negligible.

Considerations In Practical Circuits

Taking the view point of a simple resonant circuit, the following circuit aspects will be examined to see how the circuit loss accrues and how circuit Q is affected:

1. ***Single tuned interstage coupling circuit:***

For a narrow band or a single frequency amplifier, a special form of impedance coupling is a parallel tuned circuit as shown in Figure 3. When the coupling capacitor C_c is large enough so that its reactance is negligible, (this is the usual case), then the interstage circuit has only two terminals and is in fact a simple resonant circuit in parallel form. The resulting Q of the circuit is of interest because it concerns not only the stage gain but also the passband or frequency selective characteristics. Usually, the circuit Q is much lower than the combined Q of the coil and tuning capacitor C because there are several other losses involved.

a. *r_p of the first stage:* Expressed in equivalent circuit form, the first stage becomes a current source of $g_m e_g$ with r_p, the plate resistance of the tube, as a shunt load across the resonant circuit. Therefore, a pentode is almost always

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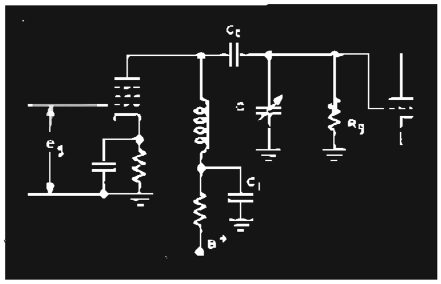

Figure 3. Single tuned interstage coupling circuit.

used as the first tube since its higher r_p means less loading.

b. *Input loading of the following stage:* Vacuum tube input impedance is generally considered high, but relative to the impedance of a parallel resonant circuit, the contrary is more often the case especially at higher frequencies.

c. *Stray capacitances with associated losses:* This could be important if the stray capacitance is an appreciable part of the total resonating capacitance. To keep Q high, the metal parts with which the stray capacitance is associated should be well grounded and dielectrics involved should have low loss.

d. *Grid resistor R_g loading on the resonant circuit:*

e. *Loss due to B+ feeding circuit:* As shown in Figure 3, the decoupling capacitor C_1 may introduce some series loss into the coil. And if parallel feed through a choke is used, a shunt load is added.

If R_p is the final equivalent shunt losses of the whole circuit, including losses of coil and capacitor, then

$$Q = \frac{R_p}{\omega L}$$

and the input voltage to the next stage will be $g_m e_g R_p$.

2. Feeding a parallel resonant circuit by a signal generator:

When a signal generator like BRC Sweep Signal Generator Type 240-A is used to feed a parallel resonant circuit, care must be taken so that the output impedance of the generator does not unduly affect the Q of the circuit. The output impedance of a signal generator is generally 50 ohms, which is much too low to be connected directly across the resonant circuit. (Also too high to be used for series feeding the resonant circuit). The usual practice is to increase the output impedance by inserting a high series resistor, R, in series with the signal generator. This resistor, R, should be high relative to the tuned impedance ($R_p = Q\omega L$), because $R+50$ is indeed loading the circuit. Any detector connected across the resonant circuit, of course, is an additional load.

Similar considerations hold when the cathode output of a tube is used to feed a resonant circuit.

3. Physical aspects of components in a circuit:

What is under consideration here is the change that is involved when a component is physically connected into a circuit. When, for example, a coil is shunted across a capacitor, in an idealized circuit analysis, this means nothing more than putting two symbols together. But actually, changes are involved in two general aspects: (1) due to proximity of two components, change of both inductance and capacitance is possible; (2) the physical connecting link, perhaps a copper strap, may have an effect on circuit performance which cannot be ignored. This kind of critical consideration primarily arises in problems of measurement, but in practical circuits stray capacitance and lead inductance mean practically the same thing. This situation becomes more important as the use of lumped constant circuit elements is extended to higher frequencies where coils become small and series impedances very low. A $0.1\mu h$ coil at 50 mc has a reactance of about 32Ω. A Q of 320 means a series resistance of $32/320 = 0.1\Omega$. If at such a low impedance level, the contact resistance of a plug-in connection is of the order of a milliohm, it will show an appreciable effect on the Q of the coil. When this same coil is measured on a Q Meter, a poor connection will lead to a jitter in the Q reading or wide variations of results.

4. Circuit Q and Effective Q:

Why does a coil of $Q = 300$ measure only 280, for example, on the Q Meter? Why do different types of Q Meters sometimes give different readings for the same coil? These are the questions to be clarified here.

The coil:

Consider a coil expressed in a series form with inductance L and resistance R_s; i.e., a two-terminal element of impedance $Z = R_s + j\omega L$. It has a quality factor of

$$Q = \frac{\omega L}{R_s}.$$

The Q and Z are primarily characteristics of the coil alone. But when the coil is connected into a circuit, the proximity effect can change the distributed capacitance and create mutual inductance. Hence, it should be realized that strictly speaking an impedance, hence its Q, is not completely defined until the way it is connected into a circuit is specified.

The measuring circuit:

The Q Meter measuring circuit consists of a signal source, a variable tuning capacitor, and a voltmeter. Ideally with the coil connected, the circuit should be as shown in Figure 4a, but actually the circuit for the Q Meter Type 260-A should be represented as in Figure 4b. (For Q Meter Type 190-A, see its Manual) These spurious elements that are unavoidably introduced are known as residuals — residual inductance and residual shunt or series losses.

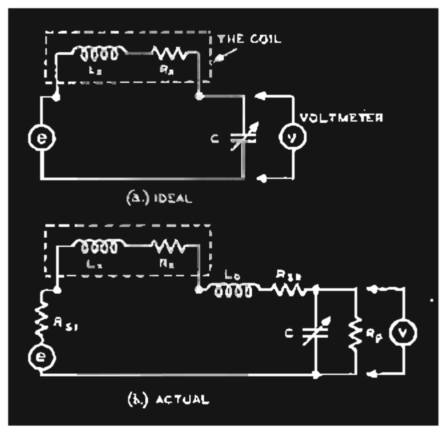

Figure 4. Q Meter circuit.

R_{s1} is due to the oscillator injection circuit.

R_{s2} includes series resistance of binding posts, connecting straps, etc.

L_o is the residual inductance of binding posts and connecting straps.

R_p includes the voltmeter input resistance, the 100 megohm grid resistor and other dielectric losses across the capacitor.

The Q Meter is designed to read the Q of the whole circuit. The reactance is now $L_o + L_x$ instead of L_x and internal losses are added. The resulting Q is of course not the same as the Q of the coil. Recognizing the effect of the internal losses, two kinds of Q are defined: (a) *Circuit Q:* Q of the whole Q measuring circuit including losses of the coil, all the internal residual losses and the effect of the residual inductance. (b) *Effective Q:* Q of the coil itself as mounted on the Q Meter used (proximity effect included). In many but not in all cases, circuit Q is essentially equal to effective Q.

The term *indicated* Q is often used; this refers to the circuit Q as indicated by the Q Meter which is designed to indicate circuit Q. This means that the difference between *indicated* and *circuit* Q is strictly a matter of the accuracy of the Q Meter. If accuracy of the Q Meter is not in question, *indicated* and *circuit* Q mean the same thing.

Now, circuit Q depends on coil loss as well as internal losses. Therefore, as internal losses may differ among different Q Meters, either of the same type or of different types, the circuit Q measured on different Q Meters will differ from each other even if the coil measured is the same one. This difference is usually small, especially if the Q Meters used are of the same type. But in the overlapping ranges (20—50 mc) of the 260-A and 190-A Q Meters, it can be as much as 50% in some unusual cases. This may seem startling but as explained in the next section, this difference can be completely accounted for by the difference in residuals, which are much lower in the 190-A Q Meter since it is designed to cover a higher frequency range.

Correlation of 190-A and 260-A Q Meters in Overlapping Frequency Ranges

When the residual parameters in a Q Meter are all known, correction can be made on the circuit Q to allow for the effects of the residuals and thus obtain the effective Q by computation. The effective Q readings for the same coil as computed from the circuit Q in different Q Meters should be the same, except for a possible difference due to the difference in proximity effects and contact resistance. When proper care is taken, this difference should be very small.

In the overlapping ranges (20—50 mc) both Type 190-A and 260-A Q Meters can be represented by the circuit shown in Figure 5a, which can be transformed into Figure 5b and 5c. (A more refined circuit for the 190-A Q Meter is given in its Manual).

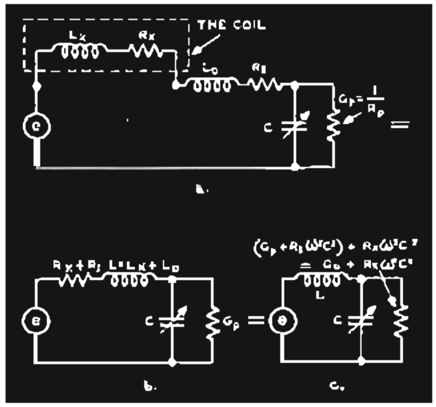

Figure 5. Q Meter equivalent circuits.

where

L_x, R_x = inductance and resistance of coil

L_0 = total internal residual inductance

R_s = total internal series resistance

$G_p = \dfrac{1}{R_p}$ = total internal shunt conductance.

$G_0 = G_p + R_s\omega^2C^2$ = total internal loss expressed as shunt conductance.

The correlation of the two Q readings of the same coil in Type 260-A and 190-A Q Meters will be demonstrated by comparing the results of effective Q computed in each case from the indicated Q readings. The indicated Q will be taken the same as circuit Q; i.e., each Q Meter is assumed to have perfect accuracy. To get effective Q, the correction has two parts; (A) for residual inductance, (B) for residual losses.

A. Correction for residual inductance: Here the reactance of the coil itself will be computed.

$L_x = L - L_0$

$\omega L_x = \omega L - \omega L_0$

$$\text{since } \omega L = \frac{1}{\omega C} \text{ at resonance}$$

$$\omega L_x = \frac{1}{\omega C} - \omega L_0$$

$$= \frac{1}{\omega C}(1 - \omega^2 L_0 C)$$

$$= \frac{1}{\omega C}\left[1 - \omega^2(L_0+L_x)C\frac{L_0}{L_0+L_x}\right]$$

$$= \frac{1}{\omega C}\left(1 - \frac{L_0}{L_0+L_x}\right)$$

$$\text{since } \omega^2(L_0+L_x)C = \omega^2LC = 1$$

The equivalent capacitance that will resonate with L_x is

$$C_x = \frac{1}{\omega^2 L_x} = \frac{C}{1-\omega^2L_0C}$$

$$= \frac{C}{1 - \dfrac{L_0}{L_0+L_x}}$$

$$L_0 = \begin{cases} 0.015\ \mu h \text{ in 260-A Q-Meter} \\ 0.0026\ \mu h \text{ in 190-A Q-Meter} \end{cases}$$

At 50 mc, 100 $\mu\mu f$ is in resonance with 0.1μh. Evidently, L_0 in 260-A Q Meter becomes appreciable then.

B. Correction for residual losses.

Q_1 = circuit Q = indicated Q (Perfect Q Meter accuracy assumed)

Q_e = effective Q

$$Q_1 = \frac{\omega C}{G_0 + R_x\omega^2C^2}$$

$$Q_e = \frac{1}{R_x\omega C_x} = \frac{1-\omega^2L_0C}{R_x\omega C}$$

Q_1, G_0, ω, C and L_0 are known quantities. Q_e can be computed by eliminating R_x between equations for Q_1 and Q_e above. So the computation for Q_e is straightforward in principle and does not require discussion. The method outlined is only to minimize the work of computation and also give an indication as to the relative weight of different parameters and their interrelation.

Steps to compute Q_e will be given below with a derivation outlined later. Due to difference in expressions for shunt loss, the 260-A and 190-A Q Meters have to be treated separately:

1. For the 260-A Q Meter:

a.) Compute C_x from $C_x = \dfrac{C}{1-\omega^2L_0C}$

b.) Find α from α vs C^2 graph (Figure 6)

c.) Correct α to get $\alpha' = \alpha + \dfrac{1}{3}\left(\dfrac{250}{Q_1} - 1\right)$, if this appears significant.

d.) Compute γ by $\gamma = 0.00114 \dfrac{Q_1 f_{mc}}{C\mu\mu f}\alpha'$

e.) Obtain Q_e by $Q_e = \dfrac{C}{C_x} \cdot \dfrac{1}{1-\gamma} Q_1$

where α

$$= \frac{\text{equivalent shunt loss due to residual series resistance}}{\text{actual internal shunt loss}} + 1$$

γ = Total internal loss as a fraction of total circuit loss

So α shows the relative importance of shunt and series residual losses.

And if $\gamma = 0.10$, it means 10% of the total circuit loss is not due to the coil measured but due to the internal loss of the Q Meter. The effective Q should therefore be higher than the circuit Q by the factor

$$\frac{1}{1-\gamma} = \frac{1}{1-0.1} = 1.11$$

The $\frac{C}{C_x}$ factor in $Q_e = \frac{C}{C_x} \cdot \frac{1}{1-\gamma} Q_l$ takes care of the effect due to residual inductance. When frequency and capacitance are changed, the resulting effect on γ, i.e., on the difference between effective Q and circuit Q can be easily estimated from the expression

$$\gamma = 0.00114 \frac{Q_l f_{mc}}{C\mu\mu f} \alpha'$$

together with the α vs C^2 graph.

The α to α' correction refers to level effect not discussed so far. In the 260-A Q Meter, at a frequency above 20 mc, the voltmeter loading increases as the signal level (i.e., the Q reading) decreases. The given correction for α is good for $Q_l > 100$.

2. *For the 190-A Q Meter.*

a.) Compute C_x from $C_x = \frac{C}{1-\omega^2 L_o C}$

b.) Find β from β vs C^2 graph (Figure 7)

c.) Compute η by $\eta = \frac{0.00573\, Q_l}{C\mu\mu f} \beta$

d.) Obtain Q_e by $Q_e = \frac{C}{C_x} \cdot \frac{1}{1-\eta} Q_l$

where the physical meaning of β and η correspond respectively to those of α and γ.

Outline of derivation of above relations (for 260-A Q Meter):

$$\gamma = \frac{G_a}{G_t} = \frac{G_p + R_x\omega^2C^2}{\frac{\omega C}{Q_l}}$$

$$= \frac{Q_l G_p}{\omega C}\left(1 + \frac{R_x\omega^2C^2}{G_p}\right)$$

$$\alpha = \frac{R_x\omega^2C^2}{G_p} + 1, \quad G_p = kf^2$$

$$\therefore \gamma = \frac{Q_l fk}{2\pi C}\alpha = 0.00114 \frac{Q_l f_{mc}}{C_{\mu\mu f}}\alpha$$

$$Q_e = \frac{1}{R_x\omega C_x}$$

$$G_t = G_o + R_x\omega^2C^2$$

Eliminate R_x between Q_e and G_t.

$$Q_e = \frac{\omega^2C^2}{G_t - G_o} \cdot \frac{1}{\omega C_x}$$

$$= \frac{C}{C_x} \cdot \frac{1}{1 - \frac{G_o}{G_t}} \cdot \frac{\omega C}{G_t}$$

$$= \frac{C}{C_x} \cdot \frac{1}{1-\gamma} Q_l$$

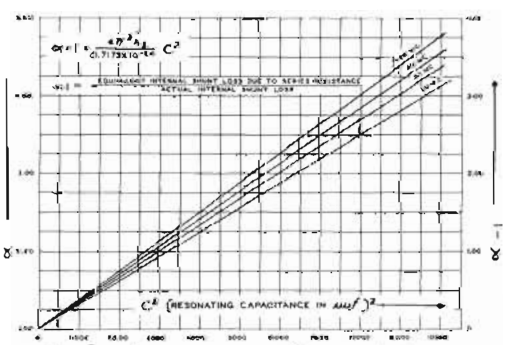

Figure 6. α versus C^2 graph for Q Meter Type 260-A.

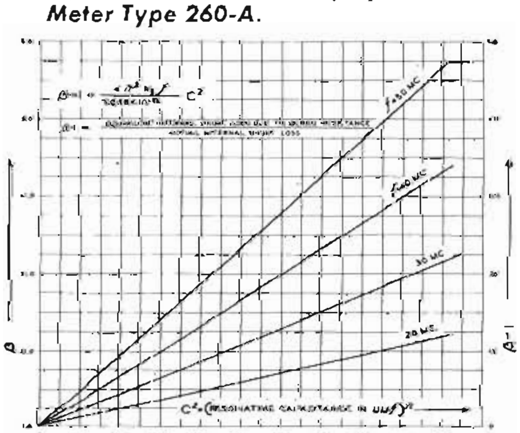

Figure 7. β versus C^2 graph for Q Meter Type 190-A.

Example: Computation of Q_e from measured data of a 0.1 μh Coil at 50 mc.

260-A Q Meter		190-A Q Meter	
Measured data:			
C	91.5 $\mu\mu f$	C	100.5 $\mu\mu f$
Q_l	187	Q_l	301
To get C_x:			
L_o	0.015 μh	L_o	0.0026 μh
$\omega^2 L_o C$	0.135	$\omega^2 L_o C$	0.0258
$\frac{C_x}{C} = \frac{1}{1-\omega^2 L_o C}$	1.16	$\frac{C_x}{C} = \frac{1}{1-\omega^2 L_o C}$	1.026
C_x	106	C_x	103.1
To get Q_e			
C^2	8380	C^2	10100
α from α vs C^2 graph	4.15	β from β vs C^2 graph	5.70
Correction for α $\frac{1}{3}\left(\frac{250}{Q_l} - 1\right)$	0.113	———	
$\alpha' = \alpha + \frac{1}{3}\left(\frac{250}{Q_l} - 1\right)$	4.263	———	
$\gamma = 0.00114 \frac{Q_l f_{mc}}{C\mu\mu f}\alpha'$	0.4957	$\eta = \frac{0.00573\, Q_l}{C\mu\mu f}\beta$	0.0979
$Q_e = \frac{C}{C_x} \cdot \frac{1}{1-\gamma} Q_l$	320	$Q_e = \frac{C}{C_x} \cdot \frac{1}{1-\eta} Q_l$	326

Measurement of Dielectric Materials and Hi Q Capacitors with the Q Meter

NORMAN L. RIEMENSCHNEIDER, *Sales Engineer*

Dissipation Factor of Insulating Material

A considerable amount of material has been published on this subject by many experts in this field describing the various techniques and the precautions to be observed in making measurements. From our own field work with companies involved in these measurements, we have come to realize the need of methods for use where the expediency required for process control work can be obtained at some possible sacrifice in accuracy by eliminating specially-developed specimen holders, guard rings, etc. It is in this sense that we offer the following suggestions for making measurements of this nature.

To review the overall operation very briefly, let it suffice to say the sample to be measured will be converted into a capacitor by adding suitable electrodes to the two parallel surfaces, and measurements made of its equivalent parallel capacity (C_p) and resistance (R_p). From these two parameters, the Dissipation factor

$$D = \frac{1}{Q} = \frac{1}{\omega C_p R_p}$$

can be determined. The whole operation can be resolved into a sequence of logical steps, with the necessary precautions, described below:

Operating Procedure

The ground plate and clip shown in Fig. 2 have been used with very satisfactory results. Prepare a plate and clip as shown and install on the Q Meter (see Figure 1).

Select the desired frequency and allow the Q Meter to warm up. Use a shielded coil whose inductance is such that it will resonate at the desired frequency with the Q Capacitor set at approximately 50 $\mu\mu$f. It is desirable to use the least possible amount of capacitance to resonate the coil since any dielectric specimen loss added to the circuit later will be more conspicuous when paralleled across a low capacity (high impedance) than a high capacity (low impedance). In any case, the lowest capacity that can be used will equal the sum of the sample capacity plus the minimum capacity (30 $\mu\mu$f) of the Q Meter internal resonating capacitor.

Selection and Preparation of Samples

Inasmuch as the ratio of loss and capacitance vary uniformly, there is quite a latitude in the choice of sample size. Very often either a 2" diameter, 1/8" thick, round disc, or a 4" x 4" square sample is chosen. It is of some advantage to use a configuration whose area can be readily computed if the dielectric constant is to be measured. In any case, increasing the area or decreasing the thickness will tend to increase the measurable "lossiness" which is desirable when measuring materials having very low dissipation factors. The sample should be clean and handling of the *edges* should be avoided to preclude the possibility of any contamination. Apply a very thin layer of Petrolatum and add aluminum or soft lead foils cut to sample size to both sides of the sample. Be sure to "roll out" any air pockets so that intimate contact is made at all points. It is also possible to employ commercially available conductive coatings which can be painted or "vacuum evaporated" on the sample.

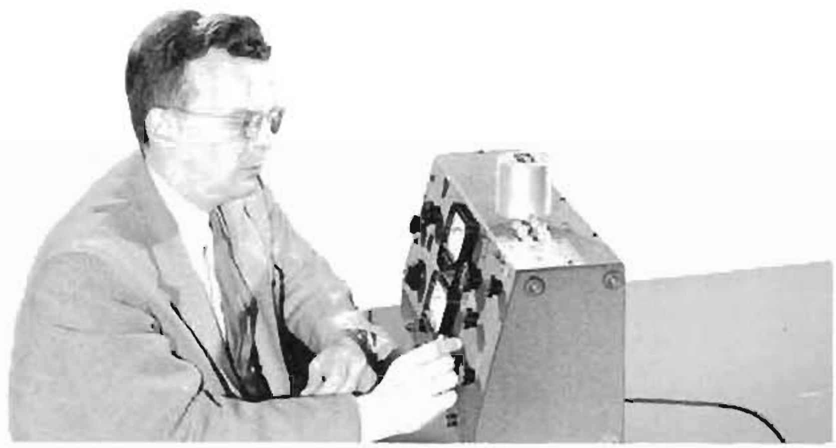

Figure 1. The author measuring the dissipation factor of Teflon.

Measurements with Sample Connected

With the ground plate secured to the case of the Q Meter and connected to "Lo" Capacitor post, mount the specimen on the plate and hold in position with the spring clip connected to the "Hi" Capacitor post. After having adjusted the "Q Zero adjust" knob, increase the oscillator output control until the "Multiply-Q-By" Meter indicates "1". Be sure this needle is at this point during all measurements. Rotate the Q Capacitor to obtain resonance as indicated by a maximum deflection on the "Circuit Q" meter. If the main condenser is set at the nearest calibration on the dial and the vernier condenser adjusted for resonance, a much closer reading can be made of capacitance. Make a record of Q_2 and C_2 at this point. (These readings are designated as Q_2 and C_2 since in Q Meter measurements they are usually recorded as the second reading). The procedure has been reversed here since specimens can be removed faster than they can be mounted and it is desirable to minimize the elapsed time between readings.

Those using a 190-A or 260-A Meter will want to take advantage of the "Delta Q" scale on the instrument. Inasmuch as the Q_1 reading will be higher than the Q_2 reading just made, the "Delta Q" adjustment will not be referenced to zero but to some other convenient point on the scale. It is also very advantageous to hold the "Delta Q" key in its operated position long enough to take advantage of the increased meter sensitivity and refine the circuit tuning as needed. The reference point finally chosen should be recorded.

Measurements with Sample Removed

Remove sample and Spring Clip Connector and resonate the circuit as above. Operate "Delta Q" key when *at resonance* and refine tuning with vernier. Read "Delta Q" value first and then the value of C_1. Q_1 will be equal to Q_2 *plus* the change (neglecting sign) in "Delta Q" reading. The needle, when in "Delta Q" operation, should always move to the right when going from the Q_2 reading (with sample) to the Q_1 reading (without sample). For those using older type Q Meters that do not have the Delta Q scale, it will be necessary to read Q_1 from the meter and compute the change.

Calculation of Dissipation Factor

The dissipation, D, is found to be:

$$D = \frac{C_1 (Q_1 - Q_2)}{(C_1 - C_2)\, Q_1 Q_2} \qquad (1)$$

And of course $Q_1 = Q_2 +$ "Delta Q" when the "Delta Q" scale is used.

Examination of the above formula emphasizes the importance of determining the $(Q_1 - Q_2)$ or "Delta Q" term as accurately as possible, since the measurement accuracy is in the same order as the determination accuracy of this term.

Example of Technique

As a means of illustrating the technique, the following measurements were were made at 1 mc on a piece of Teflon, 2" in diameter and 0.077" thick.

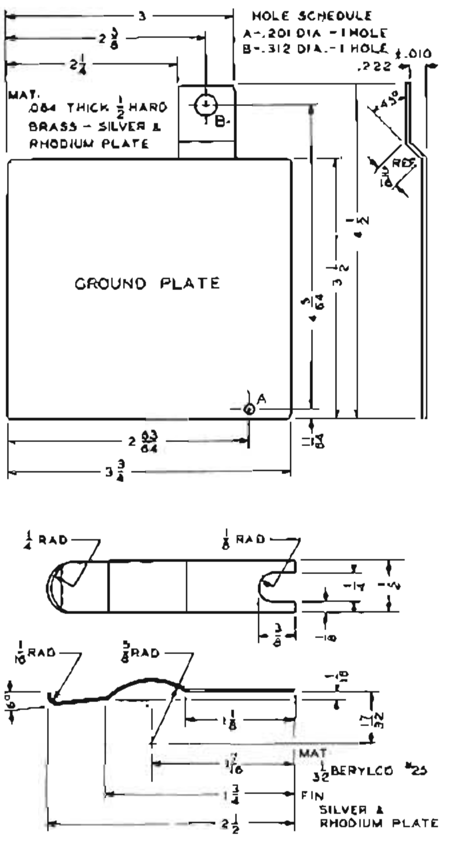

Figure 2. Details of dielectric test fixtures for Type 160-A and 260-A Q Meters.

With Sample Attached:

$C_2 = 73.0$ main condenser, and 0.12 for vernier condenser; total $= 73.12 \mu\mu f$

$Q_2 = 232$

"Delta Q" set at 25

Remove Sample:

"Delta Q" reading $= 21.2$

$\Delta Q = 25 - 21.2 = 3.8$

$C_1 = 93$ main condenser and 0.6 for vernier condenser; total $= 93.6$

$Q_1 = Q_2 +$ "Delta Q" $=$
$232 + 3.8 = 235.8$

Computations:

$$D = \frac{C_1 (Q_1 - Q_2)}{(C_1 - C_2)\, Q_1 Q_2}$$

$$= \frac{C_1 (\text{Delta } Q)}{(C_1 - C_2)\, Q_1 Q_2}$$

$$= \frac{93.6 \times 3.8}{20.48 \times 235.8 \times 232}$$

$$= \frac{3.56 \times 10^2}{1.12 \times 10^6}$$

$= .000318$ dissipation factor.

It might be worth noting that as the insulating properties of the materials tested become better, the quantity $(Q_1 - Q_2)$, called "Delta Q", becomes smaller to the extent that the use of the "Delta Q" scale, incorporated in both the 190-A and 260-A Q Meters, becomes mandatory.

Dielectric Constant

The dielectric constant, K, can be found from

$$K = \frac{4.45\, C_x\, t}{S} \qquad (2)$$

Where:

C_x = Sample capacity $= C_1 - C_2$ (from above)

t = average thickness of material in inches

S = area of active dielectric material between electrodes in square inches

Using the values determined above, the dielectric constant of the sample is:

$$K = \frac{4.45 \times 20.48 \times .077}{\pi}$$

$= 2.28$ for the sample tested.

If S and t are measured in centimeters

$$K = \frac{11.3\, C_x\, t}{S} \qquad (3)$$

Use of the Hartshorne Holder

Specimens of suitable size and thickness can be mounted in a Hartshorne type holder and measured in three ways.

1. "Resonant Circuit, Resonance Rise Method": This is similar to the technique shown above but refined to the extent of mounting the specimen in a Hartshorne Specimen Holder.

2. "Variable Susceptance Method with Air Gap": Here the specimen is mounted in a Hartshorne Holder and an Air Gap of 0.005" to 0.050" is introduced above the surface of the specimen. This method has been found suitable for measurements from 10 kc up to 100 mc.

3. "Variable Susceptance Method without Air Gap": This method, similar to the above except that the specimen is clamped in the holder instead of allowing an air gap, is employed for materials whose losses are too small to be measured by the air gap technique.

Boonton Radio Corporation Drawing Number C-302252 gives the specifications for the adapter plate used to mount the General Radio Type 1690-A Hartshorne Holder to any BRC Q Meter and is available upon request.

Measurements can also be made of insulating liquids with the provision of a suitable cell or container.

SOME NOTES ON INSTRUMENT REPAIR

One of the responsibilities of a manufacturer of precision electronic instruments is to provide facilities for repairing and maintaining his products. In a sense this responsibility begins in the development and design stages of an instrument's history for it is in this stage that proper design will minimize the need for later repair. Also at this stage arrangements which facilitate later repair can be made.

Boonton Radio Corporation operates a factory repair facility and also has authorized repair of its instruments by competent groups operated by its Representatives throughout this country and Canada. When your instrument requires repair, contact the office nearest to you, included in the list on the back page of this issue.

When your instrument is to be returned to the factory for service the following steps will expedite the repair.

1. State as completely as possible both on an instrument tag as well as on your order the nature of the problem which you have experienced. Too much information is far better

Inspector aligning the RF section of Signal Generator Type 211-A to track with the frequency dial.

than too little. If the problem is intermittent in nature be very specific. We sometimes have instruments with this type of trouble which refuse to misbehave for us.

2. State on your order whether we may proceed and bill you in accordance with our standard pricing system or whether you require that we secure your approval of the price before proceeding. The price will be the same in both cases but delay in delivery will be minimized by your permission to proceed in accordance with our standard system. Your acknowledgement copy of the order will always show the price.
3. Return the complete instrument even though you may think that some portion is not at fault. Some of our Signal Generators consist of two units (the power supply and signal generator); send both.
4. If you have made a change in your instrument and want the instrument back in the same form tell us so. Our Inspection Department will always want to make your instrument standard.

Some of our instruments have been in production for several years and we have built up a reasonable amount of repair history. For these instruments we have established standard prices based on our average experience. These prices are based on the age and condition of the instrument. Thus all instruments in a given age bracket in average condition for that age will be priced at one of our standard prices. These prices are studied and modified at the end of each year. This system saves you money and time since it avoids the necessity of a cost analysis on each individual repair.

A good repair facility requires good communication between customer and factory. If you have justified abnormal requirements for quick return of your instrument let us know. We will do our very best to accommodate you. Ask only if you have a real need. If everybody asks we cannot improve our speed for anybody.

Correction of Low Q Reading On Q Meter Type 160-A

SAMUEL WALTERS, *Editor, The Notebook*

Occasionally the mica plate which supports the measuring terminals on top of the Q Meter Type 160-A must be replaced because of surface contamination or cracking of the surface which breaks the moisture-proofing compound. The resultant rf leakage from the HI post measuring terminal to ground effectively adds a shunt resistance across the circuit under test causing a low Indicated Q reading. Sometimes the rupture or contamination is not visually apparent although just as electrically defective as the obvious case. However, the mica terminal insulator should not as a matter of course be changed when abnormally low Indicated Q readings are observed since there are three other conditions that will cause shunt losses and excessive loading of the circuit under test. These other conditions are:

1. Cracked mica internal resonating capacitor stator insulators (in later Q Meters pyrex glass was used which rarely causes trouble).
2. Defective Q voltmeter tube (105-A).
3. Grid leak associated with 105-A (100 megohm resistor) in deteriorated condition.

Before proceeding to isolate the condition causing the shunt loss, one must first determine whether the low Q readings are due to shunt losses or some other cause. This can quickly be determined with the use of the Q Standard Type 513-A*, a shielded reference inductor designed to maintain accurately calibrated and highly stable inductance and Q characteristics. Assuming shunt losses are indicated, we proceed to the next step: the isolation of the condition causing the shunt loss.

Figure 1. Bottom View: Binding post plate assembly — Q Meter Type 160-A.

Determination of Shunt Loss

The procedure is as follows:

a) Position the Q Meter to be tested (Q Meter A) three inches to the rear of another Type 160-A Q Meter (Q Meter B used as an indicating unit). Both Q Meters are to face the operator.

b) Connect a Type 103-A22 Inductor to the COIL terminals of Q Meter B. Interconnect the GND terminals of the two Q Meters by means of a 15 inch length of No. 18 stranded copper wire. Suspend a 17 inch length of No. 20 or No. 18 bare tinned copper wire (single strand) from the HI-COND terminal post of Q Meter B so that the free end of this lead points directly down toward, and is one inch removed from, the HI-COND terminal post of Q Meter A. This lead should be positioned as far as possible from other objects.

c) Apply ac power to both Q Meters.

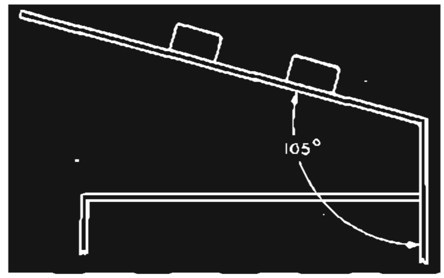

Figure 2. Angle of alignment of wiper fingers.

d) Adjust Q Meter A for oscillator frequency of approximately 800 kc and a Multi-Q-By reading of approximately 1.2. Adjust vtvm zero in usual manner and adjust capacitance dial to 30 $\mu\mu f$.

e) Adjust Q Meter B to read the Q of the Type 103-A22 Inductor at tuning capacitance dial reading of 70 $\mu\mu f$ and an oscillator frequency of approximately 1.13 mc so that the Indicated Q reading should fall between 221-235. Note the exact value of this reading as Q_1[§].

f) Now interconnect the COND-HI terminal posts of the two Q Meters by inserting the tip of the suspended bare lead into the corresponding terminal hole of Q Meter A. Resonate Q Meter B by adjusting its capacitance dial. Note the new Q reading on Q Meter B as Q_2.

g) If $Q_1 - Q_2 = 14$ or less the cause of faulty Q Meter "A" Q reading probably lies elsewhere than in the internal Q measuring circuit.

h) If $Q_1 - Q_2$ exceeds 14 proceed as follows to more specifically locate the cause of excessive errors in Q reading. Turn ac power off Q Meter A. Disconnect grid connector clip from 105-A tube grid cap in Q Meter A, allowing grid lead from Q-unit to hang in space. With the two Q Meters interconnected as for reading Q_2 (see (f) above), but with the ac power still turned off Q Meter A, resonate Q Meter B by adjusting its capacitance dial. Note Q Meter B Q reading as Q_3.

i) If $Q_3 - Q_2$ exceeds 8, voltmeter tube 105-A should be replaced as defective.

j) If $Q_1 - Q_3$ exceeds 6, excessive

Q-unit ΔQ is indicated. Correction usually lies in replacement of binding post mica terminal insulator. In rare instances the trouble may lie occasionally in the Q capacitor stator insulators or in the 100 megohm grid resistor; extension of the above procedure will quickly indict the faulty component. Both components have a ΔQ of 2.

Replacement of Mica Terminal Insulator

The replacement of the mica terminal insulator is a relatively simple procedure. However, this procedure must be followed to avoid damage to the thermocouple and expedite the mica plate's removal and replacement.

1. Unscrew knurled knobs from gold plated binding posts.
2. Remove thermocouple mounting screws.
3. Unsolder thermocouple connecting strip from LO post using hot iron. (Prolonged heat will damage the 0.04 ohm resistor in the thermocouple. This is the main reason for removing knurled knob from binding post first).
4. Remove fiducial, main Q and Vernier dials.
5. Remove 2 top screws and 4 Internal Resonating Capacitor mounting screws located under dials. The Internal Resonating Capacitor can now be removed from the instrument. Facing the front of this capacitor are two screws (on the upper left edge) that secure the top plate. Remove screws. Do not attempt to remove plate until the end flaps of the copper strap are unsoldered from their stator connections. Lift plate *straight* up so that silver contactors will not be disturbed in their alignment.

Place plate on its back with "contacts" sticking up. Use ½" wrench on the one nut securing mica to plate. Remove nut and washer. After unsoldering copper strip, drop mica from plate.

Leave plate in same position and insert new mica terminal insulator. It is important that the untinned binding post be used to secure the mica to the top plate. The remaining binding posts are tinned and should occupy the relative positions shown in Figure 1. Solder copper strap across the two top posts that are pre-tinned for ease in doing this operation. At this point the capacitor is ready for re-assembly by reversing the foregoing procedure.

Special care should be taken so that the six silver fingers are in good contact with the disc on the condenser rotor. An angle of 105 degrees must exist between front of condenser and top plate as illustrated in Figure 2.

There is one precaution in this operation: do not handle mica with fingers if possible. Use cotton gloves or tweezers.

* *See "The Q Standard — A New Reference Inductor for Checking Q Meter Performance" by Chi Lung Kang and James Wachter, Spring 1954, issue No. 1, of The Notebook.*

§ *If the Q reading is above or below this range the limits in described procedure do not apply. However, the general procedure can be used.*

NOTE FROM THE EDITOR

On September 19, 1955, Boonton Radio Corporation became the 56th recipient of the Bureau of Engineering and Safety Award. Sponsored by the Department of Labor and Industry in cooperation with the New Jersey State Industrial Safety Committee, the Award was established some years ago as part of a statewide accident prevention program.

Dr. George A. Downsbrough (right), president and general manager of Boonton Radio Corporation, shown accepting Safety Award from C. George Kruger, deputy director, division of labor of the State of New Jersey.

The goal of this year's effort is a 15% reduction in industrial accidents.

Boonton Radio Corporation was honored on the basis of its "very enviable record" of having worked the period from December, 1952 through April, 1955 with no lost time accidents and for its safety practices which produced this result.

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Determination Of Metal Film Thickness

A non-destructive electronic method applicable to combinations of coating and basis materials, at least one of which must be a conductor.

ANTS PIIP, *Development Engineer*

The measurement of the thickness of thin films, of the order of 10^{-3} inches or less, has always been one of the major problems for the coating industry, whether it be a question of metallic films on a metal basis (e.g., such as are produced by electroplating); or metallic films on an insulator; or an insulating film on a conducting carrier.

There has been a deeply felt need for a reliable, rapid, simple, non-destructive method for obtaining absolute or comparative thickness readings on the majority of film-basis combinations. Hitherto available methods lack one or more of these desirable characteristics.

Thickness measurements have been based on the following methods: mechanical, chemical, electrochemical, optical, X-ray and beta ray scattering; magnetic, electrical conductance, and last but not least, eddy-current properties.

Figure 1. The author shown measuring plating thickness of a capacitor frame.

Description of Available Methods for Measuring Plating Thickness

Mechanical: Mechanical methods involving the use of a micrometer or similar device are useful in a limited number of cases: the specimen is measured at the identical spot before and after plating. Obviously, the shape of the specimen has to be suitable for such a measurement, and the thickness of the deposit has to be appreciable.

The chemical, electrochemical and optical methods are definitely destructive.

Chemical: The majority of the chemical methods can be reduced to a determination of the weight of coating metal per unit area. They involve the following steps: measurement of area, stripping the plating with a reagent that leaves the basis unaffected; direct determination of coating material lost through stripping by means of weighing the specimen before and after stripping, or a determination by appropriate quantitative chemical analysis or colorimetry of the amount of coating material gone into solution. For rough checks, the dropping method can be used in certain cases: a prescribed stripping reagent is made to drip under controlled conditions on the piece to be tested, and the time noted until the basis becomes exposed.

Electrochemical: Electrochemical plating thickness measurement is actually a deplating operation of a known area. The number of coulombs (ampereseconds) required to expose the basis metal is a measure of the amount of material removed, as stated by Faraday's electrochemical law. The sharp change in deplating cell voltage that occurs when the basis metal is introduced into the electrolyte is used as an indicator.

Optical: For optical methods the specimen is mounted in a clamp or protective medium (usually a thermosetting plastic), sliced accurately, the cut polished, etched and measured either under a microscope or projected at known magnification.

X-ray and Beta ray: X-ray and beta ray techniques require quite elaborate instrumentation. Neither of these two methods is necessarily destructive for the specimen to be measured.

The following methods are all non-destructive, and should properly be called comparators, since none of them is capable of yielding an absolute measurement without recourse to precalibrated standards. Likewise, they are sensitive to the geometry of the specimen.

Magnetic: The magnetic methods obviously are limited to combinations where at least one of the components, either the basis or the coating, is ferromagnetic (e.g., steel, iron, nickel). Generally, they are based on measuring the force necessary to pull a small magnet off the specimen. In the case of a non-magnetic coating on steel or iron, this force decreases with increasing coating thickness. With nickel on a non-magnetic base, the force is the higher the thicker the plating. Readings depend on the smoothness and surface

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conditions of the coating, on the geometry of the test piece, on magnetic properties of the material, and local composition of the material; also, the magnetic properties of plated nickel vary widely with the plating process used.

Electrical Conductance: Instruments based on the direct measurement of the electrical conductance of the plated specimen are applicable only when the conductivities of the coating and basis materials differ appreciably. The probes

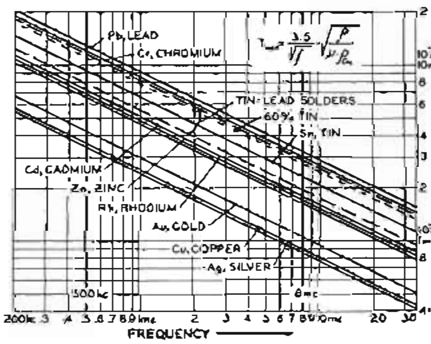

Figure 2. Penetration depths (in inches, T) of the most common plating materials.

associated with this type of instrument require good electrical contact at 4 points (2 for current injection, 2 for voltage pickup.) A hybrid method has also been used, utilizing a combination of thermal conductivity and generated thermoelectric voltage.

Eddy-Current Method

The method described below is the one employed in the instrument shown in Figure 1.

When a conductor is brought into the field of a coil excited by an alternating current, eddy-currents are induced that circulate in closed loops in the conductor. The eddy-currents generate an electromagnetic field, which in turn induces an electromotive force in the exciting coil that tends to oppose the original current in it, thereby changing its impedance. The effect on the coil is equivalent to having introduced an additional "reflected impedance" in the coil circuit.

The amount by which the coil impedance changes depends on a number of factors: the electrical and magnetic properties of the conductor, its configuration, the coil-conductor spacing, the geometry of the coil, and the frequency.

For the sake of simplicity, let us consider the case of a plane, infinite, homogeneous conductor close to the coil and normal to the coil axis. We observe the following phenomena:

Bringing the coil closer to the conducting plane decreases the reactance and increases the effective resistance of the coil. The relative reduction of the reactance depends mainly on the ratio of coil diameter to the spacing from the conductor and is quite independent of the material of the conductor — provided this is non-ferromagnetic. The increase in resistance, however, depends not only on the coil spacing, but also on the conductivity of the conductor.

Keeping the coil spacing and conductor material constant, but varying the thickness of the conducting sheet, we notice that the reflected impedance of the coil follows the increase of the conductor thickness only up to a certain thickness, called the "penetration depth" beyond which eddy-currents do not appreciably penetrate into the metal. The penetration depth, T in inches, can be expressed as

$$T = \frac{3.5}{\sqrt{f}} \sqrt{\frac{\rho}{\mu \cdot \rho_{Cu}}}$$

where: f = frequency in cycles/second

$\frac{\rho}{\rho_{Cu}}$ = ratio of resistivity of conductor to that of copper,

μ = permeability of conductor.

Figure 2 illustrates the dependence of current penetration depths on frequency and conductivity for a variety of metals.

Changes in thickness can be detected by observing the coil parameters only if the conductor is not thicker than T; after that its effect is the same as if it were infinitely thick. T is smaller for materials with high conductivity (low resistivity), ferromagnetic materials have very small penetration depths.

In the case of conductive coatings on conductors, the reflected impedance depends on the conductivities of the basis and coating and the thickness of the coating, with the coil spacing and frequency being kept constant. The relationships between reflected impedance and the properties of the composite conductor become quite complex, but it can be easily established, that

1. Changes in coating thickness are detectable only if the thickness of the coating is not more than the penetration depth in it, approximately, and an appreciable portion of the total eddy-currents circulate in the basis metal. If the coating thickness is larger than the penetration depth, practically all eddy-currents are confined to the coating metal and variations of its thickness have negligible effect on the distribution of the eddy-currents.
2. Combinations of the same basis material with coatings of different metals, but of the same thickness, result in different changes in reflected impedance, referred to that on the bare basis. These changes are larger, if the coating and basis conductivities differ by a larger ratio, since a much larger fraction of the total eddy-currents is confined in the coating than in the case where both the coating and basis are very close in their conductivities.

The foregoing discussion is applicable to magnetic materials as well, if we take into account that the "surface conductivity" is a function of permeability and frequency as well as bulk conductivity.

Having discussed the general properties of eddy-currents, we find that they offer a possibility for measuring film thickness of all three kinds: conducting films on conductors through detecting changes in current distribution; conducting films on non-conductors by variations in the total amount

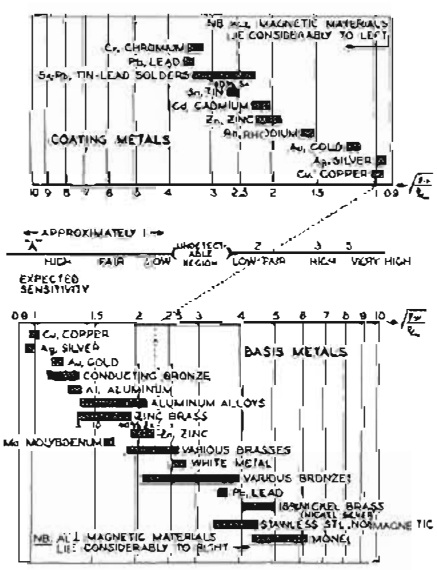

Figure 3. Frequency parameter "A" and expected sensitivity as functions of basis and plating materials.

of eddy-currents in a film thinner than the penetration depth; and non-conducting films on sufficiently thick (more than the penetration depth) conductors by the increased spacing of the coil from the basis. We can also conclude that eddy-currents offer a means for sorting metals according to their conductivities.

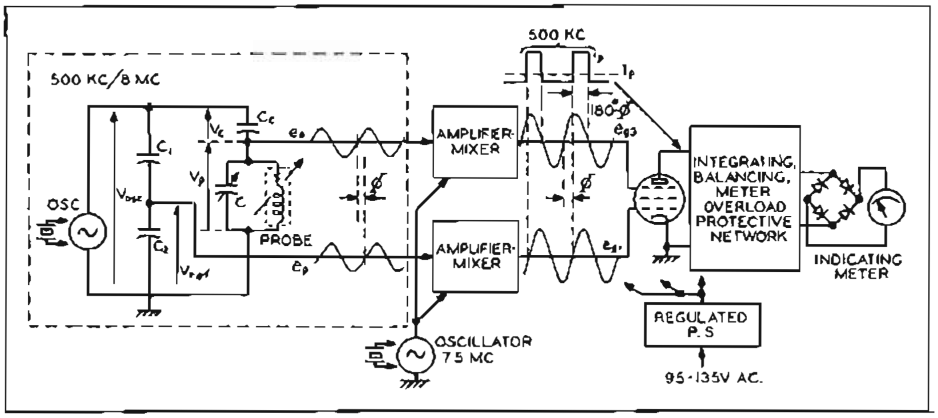

Figure 4. Functional block diagram of the Metal Film Gauge, Type 255-A.

Various techniques using the eddy-current principle have been used in the testing of materials, including measurements of film thickness. F. Forster in Germany, a pioneer in the field; Brenner and his associates at the National Bureau of Standards, and the Atomic Energy Commission are only a few who have made contributions in this field.

Determination of Suitable Test Frequency, and Expected Measurement Sensitivity for Conducting Films on Various Basis Metals

Irrespective of the kind of instrument to be used or the changes of probe impedance it responds to, the test frequency should be chosen accordingly to the thickness of the coating and composition of the film-basis combination to be measured. Quite generally, a thinner coating requires use of a higher frequency. Also, coatings of lower conductivity should be measured at a somewhat higher frequency than those of higher conductivity for comparable plating thickness on the same basis metal.

These relationships can be visualized more easily if we study Figure 3, a nomograph prepared in conjunction with Figure 2 for the choice of test frequency. A factor "A" for each combination of the more common coating and basis metals is obtained by dropping perpendiculars from the boxes pertaining to the coating and basis metals to the appropriate base lines of the nomograph, and joining these points by a straight line. "A" is then found at the intersection of this straight line with the center line. This figure represents the desirable ratio between penetration depth in the coating metal and maximum expected coating thickness.

Let us assume it is desired to measure 1 mil (10^{-3} inches) of silver plating on a yellow brass of average composition. For this combination, we find A is approximately 2. Consequently, we should choose a test frequency such that the penetration depth in silver will be of the order of 2 mils. From Figure 2, this is obtained at a frequency of 1.5 mc. A frequency somewhat lower, e.g., 500 kc, will increase the maximum measurable thickness, but some loss of sensitivity is to be expected on very thin coatings. A higher test frequency, e.g., 8 mc, will prevent us from measuring up to the full 1 mil thickness although we get better sensitivity on very thin platings.

Considerations for the Design of an Eddy-Current Film Thickness Gauge

The eddy-currents induced by the coil in the conductor circulate in the latter and diminish in amplitude as one goes deeper into the material, and also with distance from the coil along the surface. This decrease of the amplitude of the eddy-currents with distance from the coil determines the diameter of the coil which must not be more than about ½ the smallest linear dimension of the surface to be measured

Since variations of reflected impedance depend on the ratio of coil diameter to mean spacing from the conductor, the coil should be flat and very close to the specimen if sufficient sensitivity is to be obtained.

In general, variations of the resistance component of the reflected impedance are characteristic of the conductors in the coil field, and we want this variation to be as large as possible in relation to the resistance of the coil proper for good sensitivity. Therefore, a high-Q coil is indicated. The sensitivity can be increased furthermore if the probe is made the inductance in a resonant circuit operating at or near resonance.

The spacing of the probe coil to the conductor (or their relative angular orientation, which is equivalent to a change of mean spacing) exerts a major influence on the reflected impedance. Therefore, means should be provided to guarantee that the spacing and orientation of the probe coil are always kept constant in use.

Phase changes are less affected by slight variations in probe spacing and orientation and reasonable surface imperfections (such as scratches, dirt, scale), than functions which also involve the magnitude of probe impedance. Phase changes are also relatively independent of voltage levels.

A careful study of the coating-basis combinations commonly encountered in the plating art, and the coating thicknesses thereof, shows that 500 kc and 8 mc are suitable test frequencies, which together provide a continuous range of thickness measurements of about 20:1 for any given combination of plating and basis material.

Metal Film Gauge, Type 255-A

Figure 1 shows the Metal Film Gauge Type 255-A which was designed to meet the requirements outlined above.

Fundamentally, the instrument consists of an oscillator driving probe and reference phase circuits. The probe circuit is made resonant at the oscillator frequency with the probe placed on the basis material. Both the probe signal and the reference signal are impressed, after suitable amplification and amplitude limiting, on the grids of a gated-beam phase detector tube. The plate current of the phase detector varies in accordance with the phase difference between the probe and reference signals, and actuates the indicating meter. The meter has easily interchangeable scales which can give a direct reading on any plating-basis combination and obviate the need for separate calibration curves. Two complete probe assemblies go with the instrument, one for 500 kc and one for 8 mc. The probe coils are approximately ¼ inch in diameter, making possible measurements on flat samples over ½″ across.

A more thorough understanding of the operation and design of the instrument can be obtained by referring to a functional block diagram, Figure 4, and

V_{osc} = const = $V_p + V_c$, oscillator tank voltage

V_{ref} = const, in phase with V_{osc}; reference voltage

I — current through C, coupling capacitor and probe circuit.

V_p — voltage across probe circuit.

V_c — voltage across coupling capacitor C; lagging 90° behind I

φ — phase angle of probe circuit.

ϕ — phase angle between V_{ref} and V_p

$\theta = \phi_1 - \phi_0$ = change of phase angle between V_{ref} and V_p due to a change of the phase angle of the probe circuit. (θ is the angle by which conduction angle of phase detector changes)

Subscript $_0$ — at probe resonance. $\varphi_0 = 0$

Subscript 1 — with detuned probe.

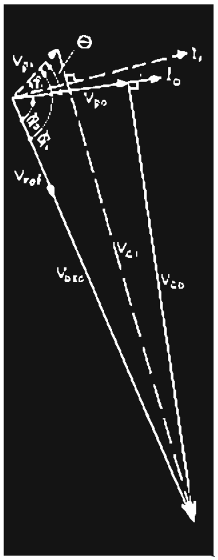

Figure 5. Vector diagram of the 255-A probe circuit.

the vector diagram of the probe circuits, Figure 5.

The test frequency, either 500 kc or 8 mc, is generated in a crystal-controlled oscillator which can be switched to either of these frequencies. Since the limiting and phase detection circuits operate at 500 kc, a 7.5 mc heterodyne oscillator frequency is provided for mixing with the 8 mc signals from the high frequency probe and reference phase circuits when in use.

The output of the oscillator is applied across the series combination of a coupling capacitor and the probe circuit which together effectively form a phase shifting network (Fig. 5). The probe is resonated by a variable air capacitor. A fraction of the total oscillator voltage, in phase with it, is used as the reference signal. With the probe resonated, there exists a certain phase difference ϕ between the probe and reference signals which are fed to the grids of two identical limiter amplifier-mixer heptodes, whose plate circuits are tuned to 500 kc. On the 500 kc range, the heptodes function as straight amplifiers. However, on the higher frequency (8 mc) range the heptodes are made to act as mixers and to yield 500 kc signals in their outputs, transposing the phase difference ϕ from 8 mc to 500 kc.

The outputs of the heptode stages are impressed on the first and third grids of a gated beam limiter-phase detector tube, type 6BN6. The amplitudes of these grid signals have been made sufficiently high so as to cause saturation, thus making the phase detector quite insensitive to any changes in signal amplitudes. The average plate current is a linear function of the phase difference.

The plate circuit of the 6BN6 is made part of a dc bridge containing facilities for bucking out the quiescent plate current (at probe resonance), an overload meter protection circuit, a sensitivity control providing constant meter damping, and a full wave bridge rectificer, to make the meter reading unidirectional no matter how the basis and coating conductivities are related to each other. Full scale meter deflections, at full sensitivity of the instrument, correspond to phase angle changes of the order of 4°.

The sample cards carry a piece of the basis material and specimens of known plating thickness on this basis material together with the appropriately calibrated meter scale and are inserted in the holder in such a way that the calibration comes against the edge of the transparent meter case, thus making the instrument direct reading in thickness. The encapsulated probe coil is carried on a spring loaded plunger.

Properties and Advantages of the Metal Film Gauge

The 255-A is an instrument capable of detecting small differences in the surface conductivities of metals, or of small variations in the metal-to-probe spacing. It becomes a direct reading instrument when standards of known thickness and composition are used for calibration.

Since both the conductivity and permeability can show considerable variations, depending on the local composition of the material, its heat treatment and previous history (work hardening, etc.) care must be exercised that the calibrated standards really are representative of the material encountered in the subsequent measuring process.

The absolute accuracies obtainable with the instrument depend on the accuracy with which the thicknesses of the standards are known. The instrument by itself is capable of distinguishing between film thicknesses differing only by a few percent of the thickness corresponding to full scale, when used at the correct frequency.

To use the instrument for absolute thickness measurements at least two identical sets of standards of at least three thicknesses should be made up. One of the standard sets is measured by some other means for the actual plating thickness, and the other mounted on a sample card together with a piece of bare basis metal. All the standard samples are measured in succession at the appropriate frequency and the corresponding readings on the 0-100° scale plotted on graph paper as a function of film thickness. After drawing a smooth curve through the four points (three readings and zero), a calibration curve is obtained. This curve can be transferred to the sample card and is used to measure the thickness of work pieces having the same materials and falling within the thickness range of the calibration. The calibrating and measuring techniques described above are in principle applicable to all three kinds of film basis combinations.

Moderate amounts of curvature and slight differences in basis conductivities of the specimen can be compensated by a modified operating technique.

In addition to film thickness determinations, the instrument can be used for sorting materials according to their conductivities. The main balancing ("Set Basis") control can be calibrated directly in conductivities, and the instrument used as a null device.

In measuring film combinations involving at least one ferromagnetic component, the instrument readings can under certain conditions become ambiguous, but this ambiguity can be eliminated by a proper choice of frequency.

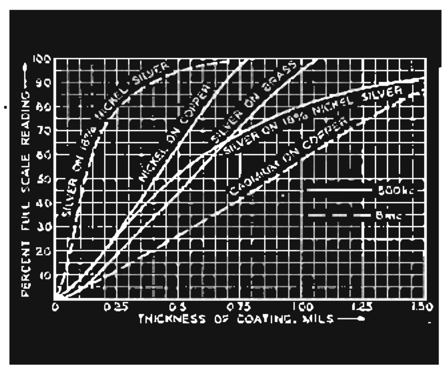

Figure 6. Typical samples of calibration curves.

THE AUTHOR

Ants Piip joined the engineering staff of Boonton Radio in the fall of 1954. His background covers work in naval electronics and ultrasonics in Tallinn, Estonia; development and research in the electronics laboratory of "AGA", Stockholm (ultrasonics, radiobeacons, special instruments); semi-conductor test equipment for Sylvania, and a miscellaneous experience in allied engineering fields. He received his EE from the Royal Institute of Technology in Stockholm (1950) and his MEE from Notre Dame (1953) where he subsequently taught and did additional post graduate work in physics.

A Method Of Measuring Frequency Deviation

JAMES E. WACHTER, *Project Engineer*

There are in use today several methods of measuring the frequency deviation of a frequency modulated signal. Most of these methods require the use of specialized equipment such as linear FM detectors and panoramic frequency analyzers. However, there is one method which, although not the least time consuming, is both straight forward and accurate, requiring the use of commonly available laboratory equipment. This method is known by various names, probably the most used of which is "The Bessel Zero Method". As this name implies, it is related to the Bessel functions, the zero-order Bessel function J_0 in particular. The fact that the zero-order function passes through zero amplitude at certain points, which correspond to discrete modulation indices, forms the useful basis for the method.

The equation of a sinusoidal signal can be expresed as

$$e = A_c \cos(\omega_c t + \theta) \qquad (1)$$

where A_c is the maximum amplitude of the carrier and ω_c is the angular frequency of the carrier. With a sinusoidal modulating signal and assuming $\theta = 0$, a frequency modulated signal can then be expressed as

$$e = A_c \cos\left(\omega_c t + \frac{\Delta\omega_c}{\omega_m} \sin \omega_m t\right) \qquad (2)$$

where ω_m is the angular frequency of the modulating signal and $\Delta\omega_c$ is the peak angular frequency deviation of the carrier.

$$\frac{\Delta\omega_c}{\omega_m} = \frac{\Delta f_c}{f_m} = B = \text{modulation index.} \qquad (3)$$

Expanding equation 2 results in

$$e = A_c [\cos \omega_c t \cos(B \sin \omega_m t) - \sin \omega_c t \sin(B \sin \omega_m t)]. \qquad (4)$$

It can be shown that [1]

$$\cos(B \sin \omega_m t) = J_0(B) + 2J_2(B)\cos 2\omega_m t + 2J_4(B)\cos 4\omega_m t + \ldots \qquad (5)$$

and

$$\sin(B \sin \omega_m t) = 2J_1(B)\sin \omega_m t + 2J_3(B)\sin 3\omega_m t + \ldots \qquad (6)$$

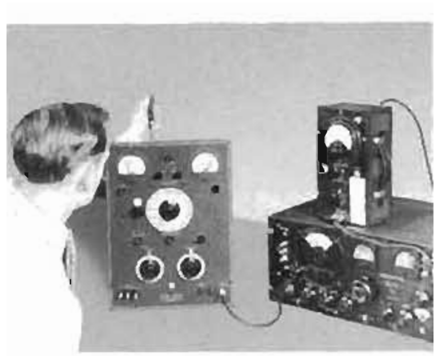

Figure 1. The author checking the frequency deviation of the Signal Generator, Type 202-B.

where the coefficients $J_n(B)$ are Bessel functions of B. On substituting equations 5 and 6 in equation 4 there results

$$\begin{aligned} e = A_c [J_0(B)\cos\omega_c t &+ J_1(B)\cos(\omega_c + \omega_m)t \qquad (7)\\ &- J_1(B)\cos(\omega_c - \omega_m)t\\ &+ J_2(B)\cos(\omega_c + 2\omega_m)t\\ &+ J_2(B)\cos(\omega_c - 2\omega_m)t + \ldots] \end{aligned}$$

which is the expression for sideband frequencies of a frequency modulated signal. It will be noted that there are possible an infinite number of sideband frequencies and that each sideband is spaced from the carrier by integral multiples of the modulating frequency. Also, it may be seen that the amplitude of the carrier decreases from a value of unity in the unmodulated condition to $J_0(B)$, which is zero for some values of B, during modulation. A graphical representation of equation 7 for $B = 25$ is given in Figure 2.

In employing the Bessel zero method, a heterodyne type receiver is tuned to the unmodulated carrier frequency of the source to be tested so that a beat frequency of some several hundred cycles is obtained, which can be monitored with earphones or a voltmeter. If some loss in measurement sensitivity can be tolerated, a crystal frequency calibrator may be substituted for the receiver, providing that the carrier frequency of the source is a harmonic multiple of the calibrator frequency. As a carrier is frequency modulated by a single frequency, the beat frequency will be observed to disappear at several points as the amplitude of the modulating signal is increased. As previously stated, these null points correspond to specific modulation indices, the first five of which are:

2.4048
5.5201
8.6537
11.7915
14.9309

These five points are graphically illustrated by the curve of the zero-order Bessel function in Figure 3. The modulation index being the ratio of frequency deviation to modulating frequency (equation 3), it becomes apparent that knowing the indices at which the carrier is zero and knowing the modulating frequency, the frequency deviation at each carrier zero is readily determined.

As an illustration, let us take a value of 10,000 cps for the modulating frequency. As the amplitude of the modulating signal is increased, nulls will be obtained at the deviation frequencies of 24.1, 55.2, 86.5, 117.9 and 149.3 kc.

It is well to bear in mind that in order to perform this test with any degree of success, it is necessary that the modulating frequency be considerably greater than the beat frequency being monitored. This is so because the sidebands of a frequency modulated signal are spaced at intervals equal to the modulating frequency (see figure 1). If such is not the case and the ratio of modulating frequency to beat frequency is, for an extreme case, only of the order of 2 to 1, the possibility exists of beating with the first side-band frequency rather than the carrier. If unknown to the operator this will produce erroneous results, since the side-bands, like the carrier, pass through points of zero amplitude at particular modulation indices (see Figure 3).

The accuracy to be expected from this method is dependent upon the accuracy of the modulating frequency and how well the nulls can be defined. For example, let us assign a value of $\pm 0.5\%$ as the accuracy of the modulating frequency. Holding the modulation index constant, this $\pm 0.5\%$ is applied directly to the frequency deviation. Further, let us assume that the sensitivity of our system is such that the smallest amplitude beat-frequency we can detect is 40 db below the beat-frequency signal due to the unmodulated carrier. Because the receiver signal amplitude remains constant, the amplitude of the beat frequency varies directly as the amplitude of the carrier frequency of the source under test. If we arbitrarily assign a value of 1 to the amplitude of the beat-frequency due to the unmodulated carrier, we may interpolate directly from a table of zero-order Bessel functions [3] to determine the definition of the nulls. Now, 40 db below 1 is 0.01, which we

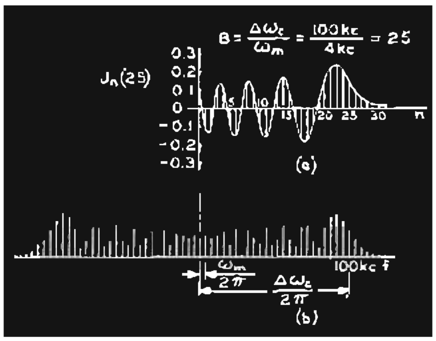

Figure 2. Bessel functions and frequency spectrum for $f_m = 4$ kc and $\Delta f_c = 100$ kc. [2]

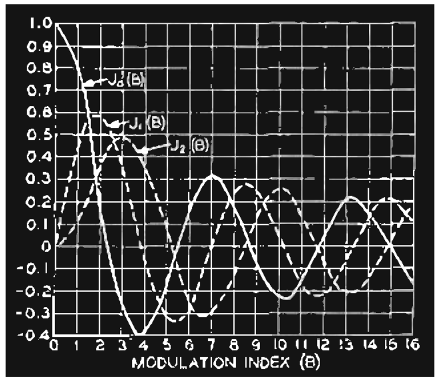

Figure 3. Bessel functions of the first kind.

can locate in the table in the vicinity of the first zero point. If what we believe to be 0 is actually 0.01, then the modulation index we obtain is approximately 2.38 instead of 2.4048, or an error of about 1.0%, which if the Bessel curve is assumed linear in the vicinity of zero, is possible on either side of zero, or ±1.0%. For this case, if we hold the modulating frequency constant, this error too is applied directly to the deviation frequency. Taking both errors into account, for this example the maximum possible error in determining the deviation frequency at a modulation index of 2.4048 is ±1.5%. Because the slope of the zero-order Bessel curve decreases as it passes through zero at higher modulation indices, the maximum possible error in determining frequency deviation will increase slightly with higher modulation indices.

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RF Calibration of the Sweep Signal Generator Type 240-A

SAMUEL WALTERS, *Editor, The Notebook*

It sometimes becomes necessary to re-calibrate the rf circuit of this generator because of changes in frequency determining components, tube aging, etc. These changes will normally cause a frequency variation of no more than 2 or 3%. Changes larger than this usually indicate serious circuit problems which re-calibration will not overcome.

The parameter most likely to affect the frequency stability is the dc reactor biasing current which is an integral part of the saturable reactor method used in this instrument for generating a sweep frequency.* This method has many advantages such as a wide sweep range, good stability and accuracy, inherently non-microphonic operation and a linear sweep. However, it introduces another variable into the oscillator circuit besides the conventional L & C: a specially shaped saw-toothed current that drives each of the five saturable reactors (one for each range). The saw-toothed current is super-imposed in the sweep condition on the dc reactor current, which provides the proper inductance at the center frequency. This dc current may change in value should some frequency determining component in the power supply through aging or some other reason change its value, thus affecting the frequency.

Discussion Of Marker System Used In Calibration

The 240-A has a self-contained means of calibration through the use of a zero beat type marker system. As shown in Figure 1, a harmonics generator produces a set of crystal-controlled reference frequencies. A front panel control permits the choice of harmonically related reference frequencies at the fundamental frequencies of 2.5 mc, 0.5 mc or 0.1 mc. The rf sample output in cw or sweep condition heterodynes with the related frequencies in a mixer stage to produce audio frequency beat notes or "birdies", as they are sometimes called.

The table below designates the variable resistors which control the reactor bias current and thus the cw and center sweep frequencies for each range. All of the variable resistors are located in a circle around the switch at the front right hand corner of the sweep chassis (see Figure 2). In all cases turning the resistor element in a *clockwise* direction causes the frequency to *increase*.

Freq. Range (MC)	Res. (CW Operation)	Res. (Sweep Operation)
4.5 to 9.0	R 540	R 533
9.0 to 18.0	R 538	R 531
18.0 to 35.0	R 537	R 527
35.0 to 75.0	R 535	R 526
75.0 to 120.0	R 534	R 525

Recalibration Procedure

A. Sweep Operation Adjustment

1. Connect the *Sweep Out* terminal posts to the horizontal input of an oscilloscope.
2. Connect shielded cable from the *Composite Signal Out* BNC jack to the vertical input of the oscilloscope. No rf detector is required.
3. Turn the *Crystal Marker* switch to the 2.5 mcs position.
4. Turn the *CW-Sweep* switch to *Sweep*.
5. To avoid any error caused by scope non-linearity, turn the scope horizontal gain all the way down. Center the resulting vertical line on the scope face, this point will mark the position on the scope of the center

of the sweep. Restore the scope horizontal gain to normal. Tune the signal generator so that one of the markers or "birdies" is at the center of the sweep. Decrease the signal generator sweep width until only

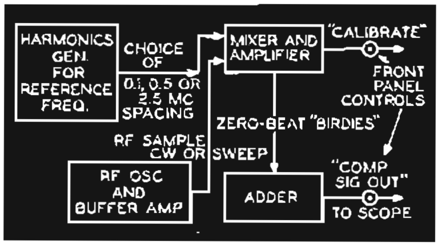

Figure 1. Block diagram of circuits used in RF calibration of the 240-A.

one marker appears and retune the frequency as required to center the marker.

6. The frequency on the dial should be a multiple of 2.5 mcs within the ±1% tolerance of the instrument. If it is beyond the tolerance, set the dial on the frequency and center the marker on the sweep display by adjusting the proper resistor (see table) while observing the precautions listed below.
7. Identify the frequency read on the dial with some frequency determining instrument such as a receiver or grid dip meter to insure that calibration was made to the correct multiple of 2.5 mcs. This should be done whether or not a recalibration adjustment was necessary.

B. CW Operation Adjustment

1. Plug a pair of headphones into the front panel jack marked *calibrate* and adjust *Center Frequency* control knob for a zero beat. If the use of an oscilloscope is preferred, connect a shielded cable from the *Composite Signal Out* BNC jack to the vertical input. Use the internal horizontal sweep on the oscilloscope to display the zero beat.
2. Turn the *CW-Sweep* switch to *CW*.
3. Turn on the 2.5 mcs crystal marker and adjust the frequency dial to obtain a zero beat.
4. The frequency indicated should be a multiple of 2.5 mcs within the ±1% of the instrument. If it is beyond the tolerance, set the dial on the frequency and adjust the proper resistor for zero beat.
5. Identify the frequency.

Precautions

The following precautions should be observed:

1. Leave the dust cover intact but remove the bottom plate.
2. Operate the instrument in its normal vertical position. Set the instrument on the bench so that it hangs over the front sufficiently to allow access to the resistors with a screw driver from the bottom.
3. Use a screwdriver with an insulated handle and a protective insulated sleeve on the shaft. This is recommended to prevent accidental shorting of the dc voltage on the resistor control to chassis ground.
4. Allow 1 hour warmup before attempting any recalibration.
5. Wait five minutes when changing frequency ranges to allow proper "settling".
6. Before making any adjustments in the *Sweep* position of the *CW-Sweep* switch, turn sweep width control fully clockwise and wait five minutes.

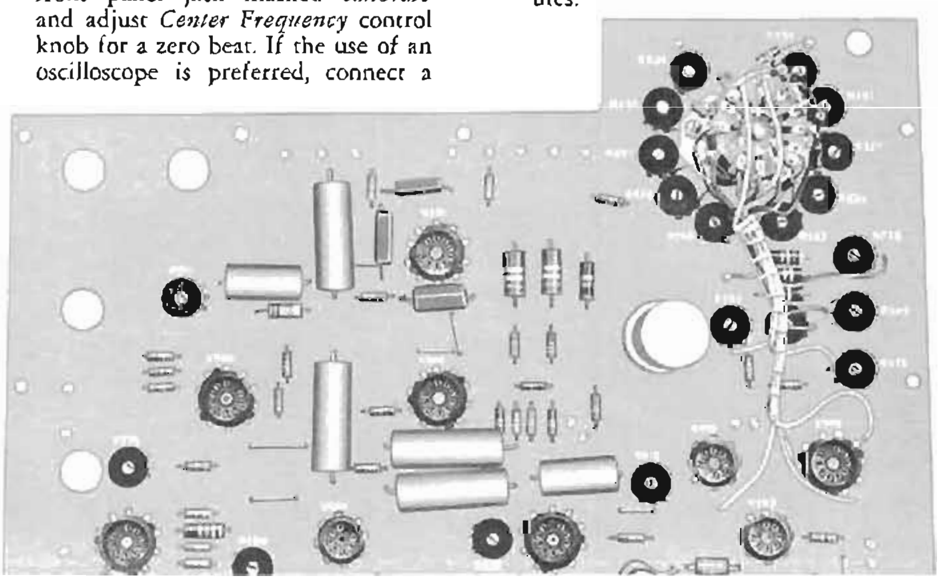

Figure 2. Sweep generator chassis, showing location of the variable resistors which control the CW and center sweep frequencies.

NOTE FROM THE EDITOR

The annual trek East each Spring to see a picture in the round of electronic progress and exchange views presents the opportunity to talk of other things, "of shoes — and ships — and sealing wax — of cabbages — and kings," things that may appear irrelevant but are actually germane.

Industrial growth, for example, is rooted in local history. Such disparate articles as cannonballs and Q meters, although produced almost two centuries apart and separated even further in function, are linked by threads of politics, geography, even geology.

In our town of Boonton we can follow the threads back to the American Revolution when the iron in its soil provided the main source of cannon balls for Washington's Army. More than once, while posted at Morristown, about ten miles distant, Washington visited the works to inspect the processes on army contracts. One of the roads the Revolutionary Army traveled between Morristown and Pompton Plains was through old "Boonetown" and legend records that the town was included among the many places where our indefatigable first president spent a night (at the home of a Colonel Ogden, who reputedly gave "Boonetown" its name in honor of Thomas Boone, Governor of New Jersey, in 1760)

Colonel Ogden at this time was the owner of the great metal slitting mills at Boonetown (now lying 60 feet under the Jersey City Reservoir at nearby Parsippany — see photo) and one of the Revolutionary Army's prime sources of military supplies. He ran considerable risk before the war in the operation of his iron works since it was unlawful to manufacture iron in the Colonies. He tried to conceal it by building a harmless grist mill in front and over it. However, word reached the authorities and Governor William Franklin (a son of Benjamin) came to inquire into a report that Colonel Ogden was flouting His Majesty's authority. A stout partisan of the belief that "a bumper of good liquor will end a contest quicker than justice, judge, or vicar," Colonel Ogden wined and dined the governor 'til the inner man glowed and the discerning eye turned myopic. Governor Franklin's subsequent report bristled at the "slander" that Colonel Ogden was making bootleg iron, and hailed him as one of "your Majestie's loyal subjects".

Prior to the Revolution, Boonton was a quiet spot of great natural beauty, nestling in an area of lakes, streams and forests. Lenni-Lenapes or Delaware Indians were the first known settlers in this area which they called Parsippanong ("where the brooks leap down the hills").

The Indians were called by the Whites after the Indian name of the river by which they dwelt; hence the Whippanongs, the Pomptons, the Rockawacks, the Parsippanongs, the Minnisinks and the Musconetcongs — the suffix "ong" meaning water and the remainder of the word describing the exact kind of water.

At or about 1700 the Indians sold their land which took in the entire northwestern territory of New Jersey and migrated to Pennsylvania and Ohio. The purchasers were a group of White proprietors of West New Jersey, most prominent among whom was William Penn who acquired approximately 4,000 acres in this area. They in turn broke the property into smaller areas and sold them to resolute settlers.

During most of the 19th century Boonton's progress depended almost solely on its Iron Industry whose fortunes waxed and waned with the times until the 1870's. In this period a double calamity broke the back of the Industry and prostrated the economy of the town — a national depression that coincided with the discovery of cheap surface ores. Many efforts were made to re-industrialize the area: silk works, hat company, soft goods industry, varnish factory, even a doll and toy company were established. None took solid root until a chemist, Edwin Scribner, started a business in 1891 that was destined to bring prosperity to the town, lure technicians, engineers and factory workers here and make Boonton's name synonymous with progress in the field of molded plastics, electronics and precision instruments the world over.

The Boonton Rubber Company, as it was known, made the first commercially

Cradle of the Iron Industry, on the site of the old Forge, where cannon balls were made for the Revolutionary Army. This site is now 60 feet under the waters of the Jersey City reservoir.

molded parts of Bakelite, which were sold to the Weston Electrical Instrument Company; thus Boonton became the birthplace of molded plastics. Today there are three firms in Boonton engaged in the molded plastics industry.

The burgeoning radio industry in the early 1920's created a great demand for molded parts, presenting technical problems whose solution gave birth to an important new industry. It was found that the available molded material which could be used at dc and audio frequencies became too lossy at radio frequencies; so engineers were brought in to make and test new materials. These pioneers of the radio art devised new electronic devices to assist them in their work and, in so doing, they developed circuits and instruments which were an innovation to the art. And thus was born in the Boonton area a new electronic instrument industry beginning with the founding in 1922 of the Radio Frequency Laboratories. There followed in relatively rapid succession and keeping pace with the development of the radio industry, Aircraft Radio Corporation, 1928; Ballantine Laboratories, 1932 (established by Stuart Ballantine, a foremost authority in his field); Ferris Instrument Company, 1932; and in 1934 William D. Loughlin, one of the original RFL staff, formed the Boonton Radio Corporation and concentrated on the development of measuring equipment which was in great demand by the radio industry. Measurements Corporation, the latest addition to the growing electronic family of this area, followed in 1939.

Early contributions of this group to the electronic art included amplifier circuits, single-control broadcast receivers, automatic volume control circuits, high sensitivity airborne receivers, throat microphones and broadcast station antenna systems. Later developments produced amplitude modulation signal generators, vacuum tube voltmeters, "Q" meters, field strength meters, pulse generators, sweep frequency signal generators and many other instruments invaluable to the electronic industry and the armed services.

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Useful Concepts Of Frequency Modulation

A non-mathematical discussion of the various ways in which a frequency modulated signal manifests itself and how its characteristics dictate the design of circuits.

W. CULLEN MOORE, *Engineering Manager*

Concept: A mental image of a thing formed by generalization from particulars; also, an idea of what a thing in general should be. (Webster)

A concept is a very personal affair, involving mental images which are subject to infinite variety. The usefulness of a concept is also quite personal; it may even fail to conform with known facts and still be useful. Indeed, a carefully selected assortment of concepts is a powerful addition to the engineer's kit of tools.

The apparent dual personality of frequency modulation is easily observed with common measuring instruments and therefore attracts attention. For example, a sweeping signal which produces a smooth response in a frequency detector circuit may, under certain conditions, fail to excite a response at some sideband frequency. Also, the frequency deviation is not an inherent characteristic of the signal, but sometimes depends on the passive circuits through which the signal is passed.

Modulation

Information is in general transmitted by changing, or modulating, some medium. In communications two aspects of the information are usually transmitted, the amplitude of the modulation signal and the frequency of the modulating signal. In frequency modulation this information is conveyed by changing the frequency of an alternating voltage whose frequency lies well above the highest frequency we wish to transmit as information. Amplitude data is carried as the extent of the carrier frequency change and frequency data as the rate at which the carrier varies above and below its average frequency.

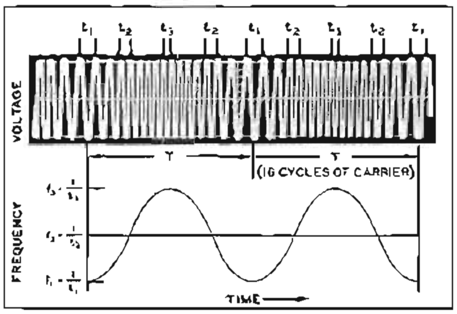

Figure 1. The variation of voltage with time as a signal is deviated ±50 kc about an average frequency of 160 kc by a 10 kc modulating signal.

In amplitude modulation it is conventional to refer to a *carrier* and *sidebands* with the carrier having a single assigned frequency. In the case of frequency modulation, however, the term *carrier* may be thought of literally as meaning the total energy used to carry the information, and as consisting of the vector sum of the center frequency and all of the sidebands. We can of course speak of the average or center carrier frequency. It is also useful to consider the instantaneous carrier frequency at any moment. The frequency swing from the average carrier frequency is called the *deviation*.

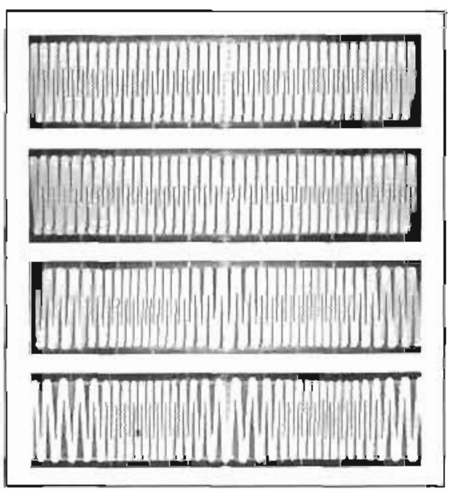

Figure 2. Increasing intensity of cycle-bunching as a 160 kc carrier is frequency modulated by a 10 kc modulating signal to deviations of 25 kc, 50 kc and 75 kc.

Figure 1 is an oscilloscope display of the manner in which the instantaneous voltage of a frequency modulated carrier, having a center frequency of 160 kc, varies with time when deviated ±50 kc by a 10 kc sine wave modulating signal. The frequency and time relationships are shown to illustrate the above terms. Figure 2 is an oscilloscope display of the manner in which the peak deviation conveys *relative* information about several different peak values of modulating voltage.

Sidebands

One might introduce the side frequencies, or sidebands as they are more commonly known, with the rather trite comment that if the peak amplitude or instantaneous frequency of the carrier change, something must have changed them. In fact, the change is proportional to the instantaneous voltage of the modulating signal.

These characteristics are very much like a familiar concept in our daily experience: inertia and force. To suggest that a carrier has inertia may seem a bit far feched. However to change its amplitude or frequency, energy must be added in much the same way as we can change the course of a rolling ball only by adding an *external* force. This added energy, the right amount in the right places at the right time, we call the sideband energy. In the case of frequency modulation, it is a matter of *re-distributing* the original energy, rather than *adding*

YOU WILL ALSO FIND . . .

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energy as is the case in amplitude modulation. Figure 3 shows how the complexity of the sidebands increases with an increase in the deviation of the carrier frequency from its unmodulated value.

Consider a uniformly rotating flywheel, representing the average frequency of the carrier, on which one spoke has been painted red and carries a direct current from the hub to the rim of the wheel. As the wheel rotates in a counter-clockwise direction, the current will induce a sinusoidal variation of voltage in a fixed conductor located along a diameter. A complete cycle of the alternating frequency will occur for each rotation of the wheel which will have in the process rotated through 2π radians; hence, our concept of *angular* frequency.

If we wish to change or modulate this frequency, we must do something about the speed of the wheel. Specifically, we must apply external accelerating or retarding forces, analogous to sidebands, which combine by vector addition with the original energy of the rotating wheel to raise or lower the output frequency. The amount and distribution of sideband energy will vary with both the rate at which we alter its speed (the modulation frequency) and the amount by which we alter its speed (modulation amplitude) as shown in Figure 3.

About Rotating Sideband Vectors

Our rotating wheel concept showed us that the frequency was directly related to the speed of rotation of the wheel which can be described as rotating so many degrees a second. Therefore, we obtain the concept of frequency modulation as being a form of *angular*, or phase, modulation in which *frequency* is defined as being the *rate of change of phase*.

In order to more readily observe changes in phase, or changes in speed of rotation (frequency), imagine that we climb onto a second wheel rotating at the *average* speed. As the carrier frequency is increased its wheel will appear to speed up counter-clockwise, and vice versa. This concept gives rise to a useful set of vector addition diagrams, shown in Figure 4, which are snapshots taken at the peak of the carrier voltage, distributed throughout the modulation cycle.

The vertical line of unit length is the vector representing the voltage at the average frequency. A pair of sideband voltages is represented by two short vectors of equal magnitude, one rotating faster than the carrier center frequency and the other an equal amount slower than the center frequency. We can determine their net effect by adding up the individual contributing voltages vectorially as shown by the dotted lines, taking first the resultant of the two sidebands and adding it to the carrier center frequency vector to obtain the final resultant voltage magnitude and relative phase angle.

The upper portion of Figure 4a shows the phase of the rotating sideband vectors adjusted with respect to the carrier in such a way that the phase (or frequency) of the carrier remains constant with respect to the wheel on which we are riding but the peak amplitude of the individual cycles of the carrier varies in accordance with the instantaneous voltage of modulating frequency. This effect we call *amplitude* modulation.

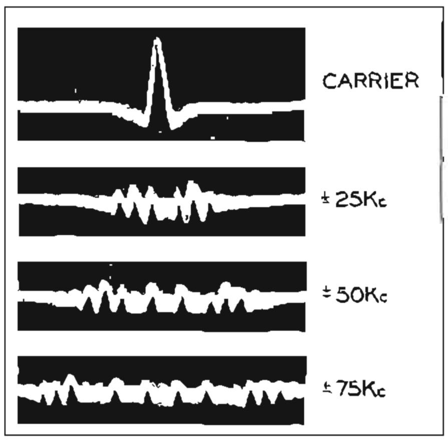


Figure 3. Relative sideband spectrum generated by frequency modulating a 60 mc carrier with a 10 kc modulating frequency to maximum deviations of ±25 kc, ±50 kc and ±75 kc (width of trace about ±80 kc).

If we now pluck off the small cluster of rotating sidebands, rotate this cluster in phase by 90° and tack it back on to the original carrier center frequency, we get a dramatically different result, shown in Figure 4b. Instead of a resultant vector whose relative phase relationship remains fixed and whose amplitude varies, the relative phase of the resultant varies and its magnitude remains reasonably constant. If we now sketch out the peak magnitude of individual cycles in the resulting wave, we see that the end result is the same kind of alternate bunching of individual cycles as was displayed on the oscilloscope for an actual frequency-modulated wave having known characteristics as shown in Figure 1.

It is interesting to note that frequency-modulated transmitters have been based on both of the concepts discussed so far. A reactance tube shifts the resonant frequency of a tuned circuit in accordance with the modulating voltage thus producing directly the waveform demonstrated in Figure 1. Likewise a type of frequency-modulated transmitter actually operates by stripping the sidebands from an amplitude-modulated wave in a balanced modulator, rotating them 90° and adding them back on to the carrier center frequency in the manner shown in Figure 4b. This process is known as "indirect" frequency modulation.

The two portions of Figure 4 indicate an interesting aspect of modulation. Amplitude variations are carried by sidebands in pairs whose vector resultant is in phase addition or cancellation with the average carrier vector. Small phase changes are carried by pairs of sidebands whose resultant is at right angles, or in quadrature, with the average carrier vector. The sine and cosine lend themselves well to the mathematical description of this perpendicular relationship.

The Constant Amplitude Problem

Pure frequency modulation imposes an additional condition on those mentioned above: the rms amplitude of the vector resultant voltage shall remain constant. This requirement demands a complicated assortment of sidebands of proper amplitude, frequency and phase. In Figure 4b only the first pair of sidebands was shown, and the vector resultant was seen to increase in magnitude with increasing change in phase. This unwanted increase can be corrected by the addition of a second pair of sidebands rotated an additional 90° to cancel our part of the amplitude change. The next pair will be rotated still another 90° to act as correction on the first phase shift pair.

Figure 5 shows how successive pairs of sidebands come into play to produce a constant-amplitude resultant peak carrier voltage swinging back and forth about an average value of phase throughout a cycle of the modulating frequency.

If through the use of too narrow a frequency bandpass, some of the outlying sideband components have been attenuated (or to look at it another way, the vector resultant voltage has slid down on the skirts of the amplifier response curve at the extremes of its excursion) the resultant will not have constant amplitude and some of the original sideband energy (or *information*) will have been lost. The resultant carrier voltage not only varies in amplitude but does not deviate in frequency in accordance with the original signal. This phenomenon occasionally is overlooked Deviation is not an inherently built-in characteristic of a frequency-modulated signal.

While a limiter cannot restore lost peak deviation information, it can restore almost constant amplitude by holding down the peak voltage at the center of the frequency excursion to a level equal to that at wide excursions. The limiting action introduces, in the form of distortion, the missing sidebands in the proper phases and magnitudes required to restore the vector resultant voltage to a constant value; and the following circuits must have sufficient bandwidth to handle these sidebands up to the point of detection. In the case of an interfering signal, the original phase information has not been lost, but only the amplitude needs correction.

The two most common methods for recovering the original information are the resonant coil discriminator in its various configurations which usually operates on a constant amplitude voltage and the so-called linear detector, which can be made reasonably independent of signal amplitude and operates by counting the rate of cycles without the use of resonant circuits. Both respond only to the vector resultant voltage which is the root mean square value of all of the individual voltages present at the input to the detector system at each instant.

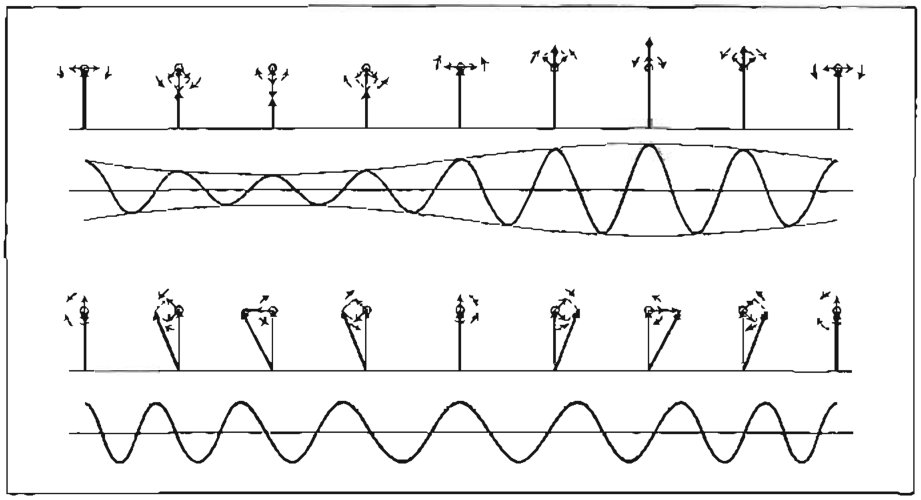


Figure 4. Amplitude Modulation of a Carrier by the First order Sidebands Compared with Phase (or frequency) Modulation of a Carrier Obtained by Rotating the Pair of Sidebands 90° with Respect to the Initial Carrier Vector, with $\beta = .5$.

Disappearing Frequencies

If one looks at an *amplitude*-modulated wave with a panoramic frequency analyzer and observes energy occurring only at the carrier frequency and at sideband frequencies spaced by the modulating frequency, one is perhaps not too surprised. However, it can be somewhat surprising to find the same kind of answer for a frequency-modulated carrier which all ones intuition and many measurements demonstrate quite clearly sweeps continuously back and forth throughout the entire frequency range under observation. Since the following exercise is both an interesting experience and a useful tool, may we suggest that the output of a sweep signal generator and the output of a frequency-modulated generator be simultaneously added in a crystal diode circuit whose output contains a RC filter and the result displayed on an oscilloscope whose horizontal trace is synchronized with the sweep signal generator?

Figure 6 is typical of the results obtained under various conditions of modulation frequency and deviation. Two things immediately strike us: first, that the energy occurs only at discreet positions in our frequency spectrum which is clustered rather symmetrically about the center frequency; and secondly, that under some conditions of modulation even these signals, including the one which we are accustomed to associating with the average carrier frequency, disappear. We must hasten to point out that we are here dealing with what the mathematician chooses to call the *frequency domain* rather than the *time domain* previously used in our oscilloscope display. We are in effect taking a cross section of frequency and displaying the energy values averaged over a period of time.

We have previously observed the physical existence of a vector resultant voltage whose instantaneous magnitude varies in approximately a sinusoidal fashion and whose separation between points of corresponding phase on adjacent cycles (a measure of frequency) varies smoothly in accordance with the applied modulating voltage. Figure 5 shows graphically how, by the proper magnitude and phase configuration of these individual voltages or "sidebands", the vector resultant may be caused to have any desired phase relationship at any given instant of time; and therefore how it is possible to cause the vector resultant voltage to sweep over any desired frequency range at any desired rate.

Those Bessel Functions

The "time average" aspect is the essence of either a physical or mathematical approach to the variation of sideband amplitudes. Specifically, the results which we observe by a physical measurement or obtain by a mathematical manipulation do not say that the energy at any given frequency is missing at all times; but only that its time average is zero.

This effect can be analyzed by mathematics using as its point of departure either of the concepts which we have discussed: a continuously-sweeping vector resultant voltage, or an assortment of sidebands. Using the idea presented

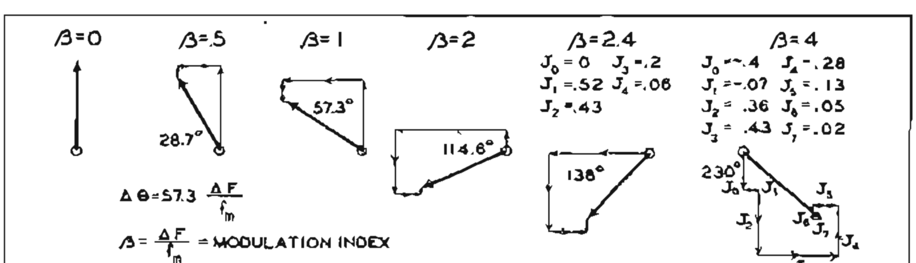
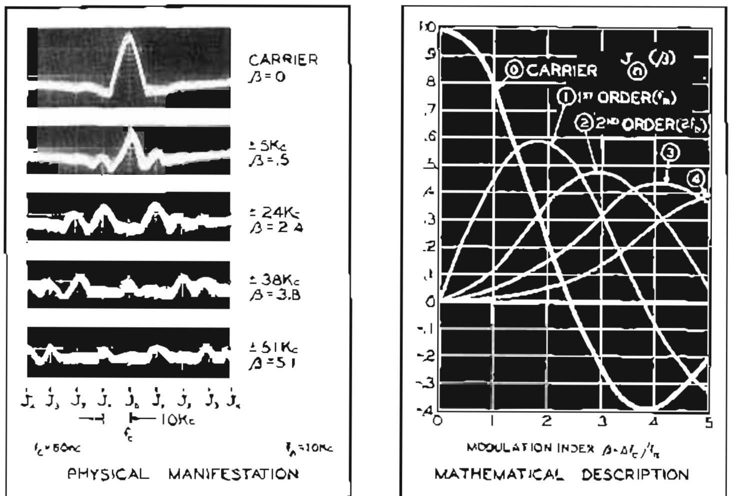


Figure 5. Generation of a Constant-Amplitude Rotating Vector by Addition of Successive Orthogonal Sideband Pairs.



FREQUENCY DEVIATION	MODULATION INDEX β	J_0	J_1	J_2	J_3	J_4
0	0	1.0	0	0	0	0
5Kc	.5	94	.24	.03	0	0
24Kc	2.4	0	.52	.43	2	.06
38Kc	3.8	-.4	0	.41	42	.25
51Kc	5 1	-.15	-.32	0	32	.4

Figure 6. Comparison of the Observed Variations in Strength of the Discrete Signals in a Frequency Modulated Wave with Their Mathematical Description as Provided by Bessel Functions.

by Figure 1, the usual textbook approach sets up an equation involving sines, cosines and phase angles which describes in mathematical symbols the way in which the instantaneous carrier voltage varies as a function of the carrier frequency, the modulating frequency, and the deviation.

The time integration of this equation to get average values leads to a never-ending series of alternating sine and cosine terms whose coefficients are the magnitudes of the successive orders of sidebands, each advanced 90° from its predecessor. The relative numerical values of these coefficients are most conveniently obtained by resort to a mathematical table known as "Bessel Functions", generated at considerable effort from a similar equation. The graph of Figure 6 shows the values of some of the coefficients as a function of the *modulation index*, which is the ratio of the deviation to the modulating frequency. We find that the mathematical formulation has indeed faithfully described the physically observed disappearance of various sideband components, designated as J_0, J_1, J_2, J_3, at critical values of the modulation index.

Having found a satisfactory mathematical description for the wave-form and concept displayed in Figure 1, it should be equally possible to derive a mathematical description of the concept involved in Figures 4 and 5. This has been done by Harvey, Leifer & Marchand[2], in which the authors use a mathematical technique which is the practical equivalent of adding up all of the vector component contributions to the final voltage over a cycle of the modulating frequency, such as might be achieved by a sufficiently large number of graphical solutions. We should not be surprised to find the solution to their equation leading once again to Bessel Functions.

Filters and Sweeping Frequencies

So far all of our physical measurements and concepts have been independent of resonant circuits. However, one of the very interesting characteristics of frequency modulation is associated with the response of a resonant circuit to a swept frequency. Let us assume a reasonably high Q (narrow passband) resonant circuit lying to the side of the average carrier frequency in a region through which the carrier is sweeping. Once again there is more than one way to look at the problem.

The first approach applicable to low repetition rates satisfies our intuitive feeling that it takes time for energy to build up in a resonant circuit, and if the circuit is sharp or the signal sweeping by quickly, only partial response will result. In fact, the peak of the response will not even coincide in resonance with the applied signal.[1] This is a serious difficulty in frequency marking circuits or in the use of fast sweeps on narrow-band amplifiers. Alternately, we might suggest that the highly selective circuit will not accept the high frequency components required to reproduce the sharp leading edge of an impulse of energy and therefore the current cannot rise quickly in the resonant circuit.

At high modulating frequencies a more subtle interpretation of the response of a filter to a swept frequency involves the stored energy.[3] A filter having a passband lying within the deviation range of a frequency-modulated signal has energy applied to it only in short bursts occurring at intervals determined by the modulating frequency and the maximum deviation. Unless the phase of the newly applied voltage has a component lying in phase with the current, no energy will be absorbed by the filter. Once again these relationships are *time* averages and apply at *all* frequencies, as well as at the sideband frequencies and are the basis for part of the derivation in the Harvey-Leifer-Marchand paper previously cited[2].

If the resonant frequency of the filter corresponds to a frequency which we call a sideband frequency, then for most combinations of deviation and modulation frequency there will be an in-phase component of the applied voltage and power will be transferred. The phase of the succeeding pulse of radio frequency energy with respect to the energy stored in the filter depends on how wide the carrier is being deviated and the rate at which it is deviating. For certain critical values the phase of the newly applied voltage will be in quadrature with the current circulating in the resonant circuit and no power will be transferred. The fact that the energy accepted by a narrow band filter does occasionally go to zero at the frequency of the carrier and the various orders of sidebands is a very useful tool.[4] It enables us, for example, to determine when the modulation index has reached a certain value as shown in Figure 6; and knowing the modulation frequency, we can determine the deviation.

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Use Of Smith Charts For Converting RX Meter Readings To VSWR And Reflection Coefficient

ROBERT POIRIER, *Development Engineer*

In a previous Notebook article, Fall 1954, issue number 3, page 7, it was mentioned that impedance measurements could be made at remote distances from the RX Meter along known lengths of 50 ohm co-ax. In this case the results obtained at the RX Meter must be transformed either by means of the Smith Chart or transmission line equations—for the ideal case

$$Z_i = Z_o \frac{Z_l \cos \beta l + jZ_o \sin \beta l}{Z_o \cos \beta l + jZ_l \sin \beta l} \quad (1)$$

where Z_i = Impedance at βl distance in radians from
Z_l = Load impedance
and Z_o = Characteristic impedance of the interconnecting transmission line.

and for the general case,

$$Z_i = Z_o \frac{Z_o \cosh \gamma l + Z_l \sinh \gamma l}{Z_l \cosh \gamma l + Z_o \sinh \gamma l} \quad (2)$$

where γ =
the complex propagation constant:

$$\sqrt{(R + j\omega L)\ (G + j\omega C)}$$

For distributed resistance, R; inductance, L; conductance, G and capacity C; per unit length for the general case.

It is the purpose of the present article to illustrate the use of the Smith Chart solutions of the transmission line equations with a view to obtaining the values of reflection coefficient ρ and VSWR from RX meter readings.

Preparation of Data

The Smith Chart is usually a plot of impedance or admittance with the rectangular coordinates curved into circles and contained within a unit circle of which the polar coordinates are reflection coefficient, $\rho = \dfrac{\text{V refl.}}{\text{V inc'd}}$ and phase angle, βl in the ideal case. The RX meter reads directly in terms of parallel resistance, R_p, and either + capacity, C, equal to the capacity in the test circuit or — capacity equal to the capacity required to resonate the inductance in the external circuit at the test frequency; neither value will plot directly onto the Smith Charts. To plot the RX Meter readings it is necessary to—

1. Evaluate the parallel reactance

$$X_p = \frac{1}{\omega C_p}$$ (where the + sign of C_p shall denote equivalent parallel capacity and the — sign of Cp shall denote equivalent parallel inductance.)

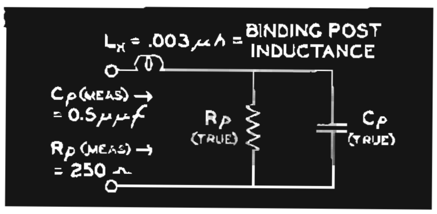


Figure 1. Equivalent Circuit of Sample Measurement.

2. Transform R_p and X_p to either rectangular admittance coordinates;

$$G = \frac{1000}{R_p} \text{ millimhos and}$$

$$B = \frac{1000}{X_p} = 1000\omega C_p \text{ millimhos,}$$

or rectangular impedance coordinates;

$$R_s = \frac{R_p X_p^2}{R_p^2 + X_p^2} \quad (3)$$

$$\text{and } X_s = -\frac{R_p^2 X_p}{R_p^2 + X_p^2} \quad (4)$$

where Q and the — sign of Xp results from the convention that inductive reactance is considered positive and capacitive reactance is considered negative.

3. Plot G and B or R_x and X_x directly onto a Smith Chart having appropriate coordinates.

For plotting on Smith Charts with normalized coordinates the following additional conversions are indicated:

$$\text{Resistance component} = \frac{R_s}{Z_o}$$

$$\text{Reactance component} = \frac{jX_s}{Z_o}$$

$$\text{or, Conductance component} = \frac{G}{Y_o}$$

$$\text{Susceptance component} = \frac{jB}{Y_o}$$

where Zo and Yo may be any source impedance or admittance respectively. These conversions are likely to be found printed on the normalized Smith Charts.

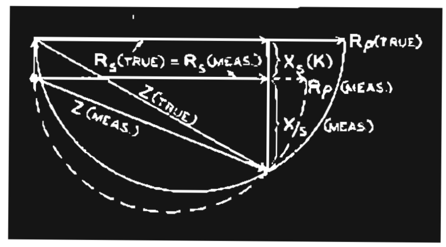


Figure 2. Impedance Circle Diagram of Figure 1.

EXAMPLE: Consider a 270 ohm ½W carbon resistor measured on the RX Meter at a frequency of 225 megacycles per second. The RX Meter readings in a typical case could be R_p = 250Ω, C_p = +0.5 μμf. The only source of error to be considered in this case is the series inductance of the binding posts, and for a first approximation this may be neglected. The binding post inductance may be best accounted for as follows: Consider the equivalent circuit of the measurement and the impedance circle diagram in Figures 1 and 2 below. From the circle diagram it is recognized that R_x (True) = R_x (Meas) so that R_x may be computed directly from (3)

$$R_x = \frac{R_p X_p^2}{R_p^2 + X_p^2}$$

$$= \frac{250 \times 2.0 \times 10^4}{6.25 \times 10^4 + 2.0 \times 10^4} = 243\Omega$$

From (4)

$$X_s\ (\text{Meas}) = -\frac{R_p^2 X_p}{R_p^2 + X_p^2}$$

$$= \frac{-\,6.25 \times 10' \times 1.41 \times 10''}{6.25 \times 10' + 2.0 \times 10''} = -42.6\Omega$$

Also from the circle diagram (Fig. 2) it is seen that

$$Xs\ (\text{True}) = Xs\ (\text{Meas}) - Xs\ (K)$$
$$= -42.6 - \omega Lk = -46.9\Omega.$$

Now let us suppose that this considered resistor is to be connected as a load for a 300Ω signal source and it is desired to predict the reflection coefficient and/or VSWR on a 300Ω transmission line connecting the source to the load. R_s (True) and X_s (True) may be normalized by $\frac{Rs}{300\Omega}$ and $\frac{Xs}{300\Omega}$ respectively. The normalized resistance component, 0.81 and reactance component −j0.156 are plotted directly on a normalized Smith Chart (expanded scale) as shown in Fig. 3. The reflection coefficient, ρ, that is the ratio of the voltage reflected to the voltage transmitted is equal to the length of the radius vector from the center of the Smith Chart to the plotted impedance. In this example $\rho = 0.13$. The reflection coefficient in terms of power,

$$\frac{P\ \text{refl.}}{P\ \text{inc'd}} = \left(\frac{V\ \text{refl.}}{V\ \text{inc'd}}\right)^2 = 0.017$$

Additionally, a return loss may be expressed as:

$$10 \log \frac{P\ \text{inc'd}}{P\ \text{refl.}} = 10 \log \frac{1}{.017} = 17.7\text{db}$$

and a transmission loss, expressing in db the incident power which is not absorbed by the load, may be written as:

$$10 \log \frac{P\ \text{inc'd}}{P\ \text{abs'd}}$$

$$= 10 \log \frac{1}{1 - .017} = 0.075\ \text{db}$$

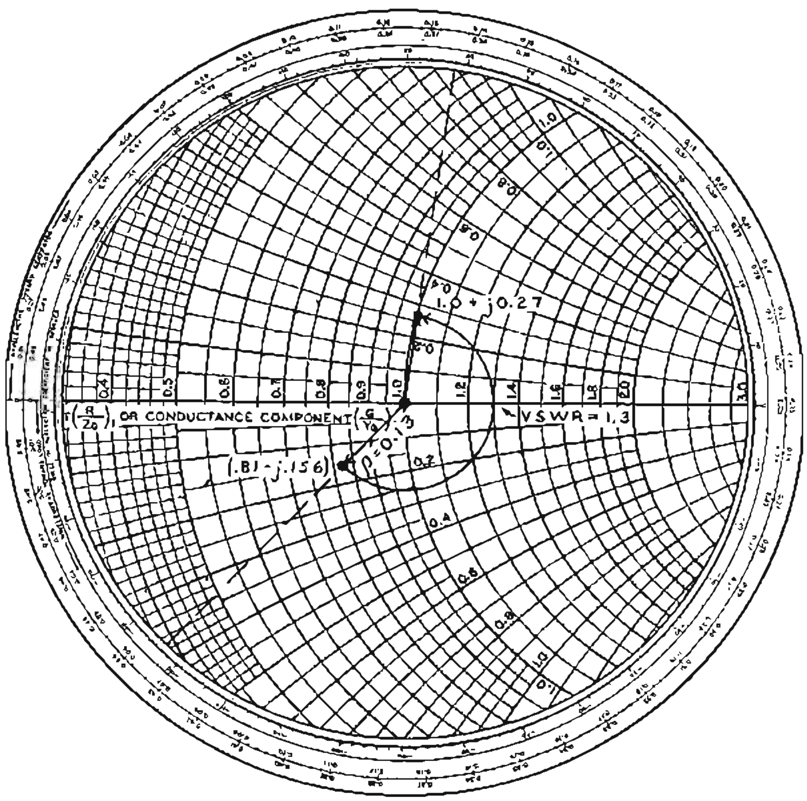


Figure 3. Smith Chart (Expanded 2X).

Modified Procedure For Transmission Line

The foregoing definitions of reflection coefficient are applicable whether a transmission line is involved or not. Now let us further suppose the Z (true) (normalized) = .81 − j.156 was measured at one end of a transmission line and that the other end of the transmission line was terminated in an unknown load to be determined. Let the transmission line be measured and found to be 0.3 wavelength long. VSWR on a transmission line is defined as the ratio of the maximum voltage to the minimum voltage; viz,

$$\frac{V\ \text{inc'd} + \rho\ V\ \text{inc'd}}{V\ \text{inc'd} - \rho\ V\ \text{inc'd}} = \frac{1 + \rho}{1 - \rho}$$

On a Smith Chart the reflection coefficient for a given load is a radially scaled constant so even though the now unknown load is not represented by .81 − j156 we can use $\rho = 0.13$ previously obtained and find,

$$\text{VSWR} = \frac{1 + 0.13}{1 - 0.13} = 1.3$$

which in db is written,

$$20 \log \text{VSWR} = 2.28\ \text{db}$$

Except in the ideal case, however, this is an approximation which is very good for short (in terms of wavelength) low loss transmission lines.

As previously stated the Smith Chart[1] provides a ready solution to the transmission line equations (1) and (2). In the ideal approximation the impedance at any point along a transmission line may be found by rotating the constant radius vector around the center of the Smith Chart the βl distance between the known and the unknown in the direction indicated on the chart. For our example $\beta l = 0.3$ wavelength; the impedance was represented by 0.81 − j.156 measured at what may be considered the input end of the transmission line. To find the impedance at the load end, the radius vector is rotated 0.3 wavelength toward the load (counter clockwise direction) as shown in Fig 3. The true impedance at the load end of the line is read from the Smith Chart as 1.0 + j0.27 per unit ohms. Since the reference unit in this case was 300Ω, Z = 300 + j81. It is also of interest to note that in crossing the horizontal axis of the Smith Chart to the right of center, the radius vector denotes a pure resistance point of maximum impedance. It is also a point of maximum voltage and minimum current. Since,

$$V\ \text{max.} = V\ \text{inc'd}\ (1 + \rho)$$

$$\text{and, I min.} = \frac{V\ \text{incd}}{Zo}\ (1 - \rho)$$

$$Z\ \text{max.} = \frac{V\ \text{max.}}{I\ \text{min.}} = Zo \frac{1 + \rho}{1 - \rho}$$

$$\frac{Z\ \text{max.}}{Zo} = \text{VSWR}$$

That is to say the per unit impedance denoted on the horizontal axis of the Smith Chart to the right of center is equal to VSWR, and we read from Fig. 3, VSWR = 1.3 as obtained previously.

Reference: P. H. Smith "Transmission Line Calculator" Electronics Jan. 1939.

Frequency Calibration Of Q Meter Type 260-A

SAMUEL WALTERS, *Editor, The Notebook*

Two of the principal reasons for checking the frequency calibration of the 260-A are (1) replacement of the oscillator tube and (2) a desire to obtain more accurate inductance readings. The former involves an adjustment on only one frequency band since all bands are affected in the same direction and approximately in the same degree. On the other hand, an inductance reading of greater accuracy than factory tolerances may require a correction curve for more accurate use of the F dial*.

Each of the eight frequency ranges has two calibration adjustments, a threaded magnetic core for the inductance of the tuned circuit and an adjustable piston type trimmer condenser for varying the capacitance of the tuned circuit. The former is used to establish frequency calibration at the low frequency end of the range and the latter to establish calibration at the high frequency end. The threaded magnetic core is adjusted at the factory and is then sealed in its coil form with a high Q lacquer to prevent movement. This adjustment should not be disturbed.

In addition to the adjustments for each range there is a variable plate trimmer condenser (C-129) for adjustment of the circuit minimum capacitance when the oscillator tube is replaced. Continuous tuning of each range is handled by the two-section variable capacitor whose sections have been precalibrated to follow a standard capacitance vs. rotation curve. No further adjustments of the plates should be necessary or attempted without a clear knowledge of the interdependence of all eight ranges. If adjustments are absolutely necessary, however, the outer rotor plates are slotted to provide minor corrections. For ranges 10-23 mc and 23-50 mc, adjust the 13 plate section; for ranges 300-700 kc, 700-1700 kc, 1.7-4.2 mc and 4.2-10 mc, adjust the 25 plate section; for ranges 50-120 kc and 120-300 kc, adjust both sections. In making these adjustments, care should be taken that rotor to stator plate spacings are not less than 0.015 inches.

Oscillator Re-Calibration Following Tube Change

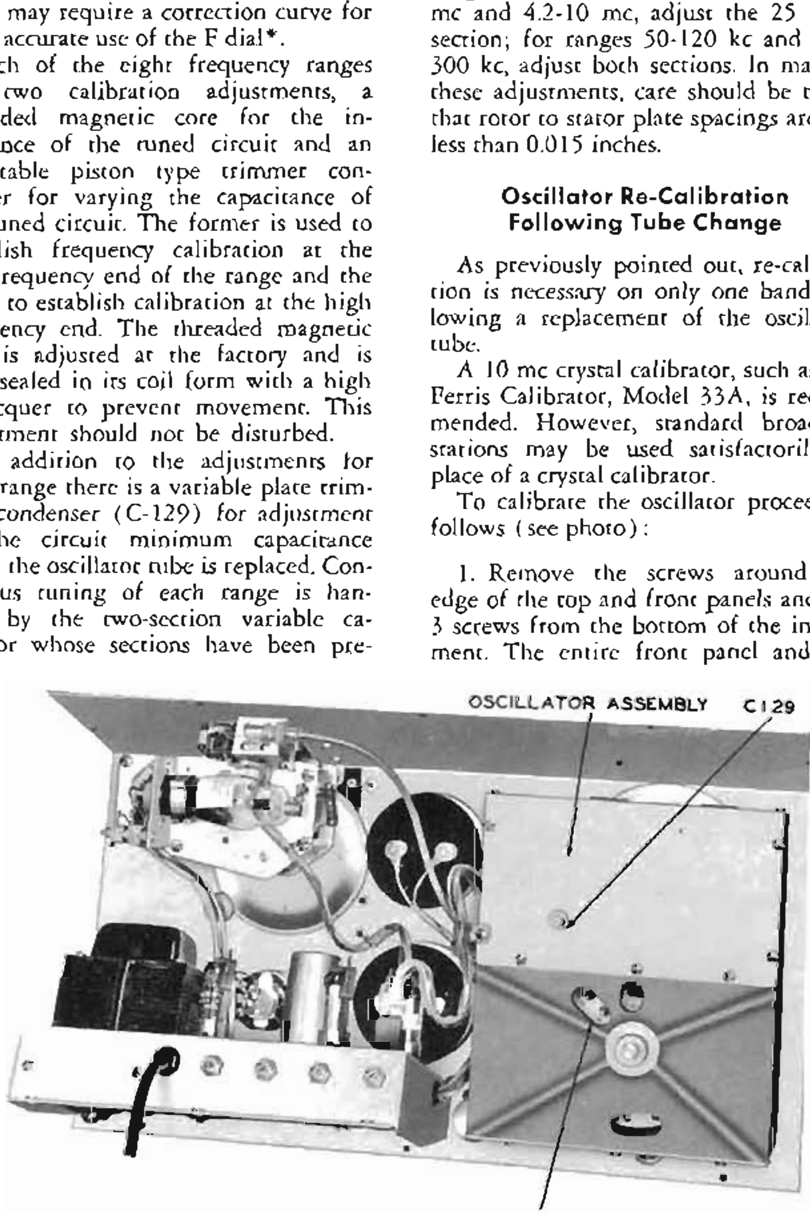
As previously pointed out, re-calibration is necessary on only one band following a replacement of the oscillator tube.

A 10 mc crystal calibrator, such as the Ferris Calibrator, Model 33A, is recommended. However, standard broadcast stations may be used satisfactorily in place of a crystal calibrator.

To calibrate the oscillator proceed as follows (see photo):

1. Remove the screws around the edge of the top and front panels and the 3 screws from the bottom of the instrument. The entire front panel and top can now be gently lifted out of the cabinet. As noted in the Figure, the shaft of the plate tuning condenser, C-129, extends beyond the top oscillator shield wall. The piston type trimmers can be easily reached through the access holes in the oscillator casing.
2. Turn on the Q Meter and allow instrument to warm up for 30 minutes.
2. Connect the rf input terminals of the crystal calibrator to the LO and GND terminals of the Q Meter.
3. Adjust the calibrator to 10 mc.
4. Switch the frequency range to the 4.2 - 10 mc range. Set the *Megacycle* dial to exactly 10 mc.
5. Adjust the XQ controls for a reading of 1.0 on the *Multiply Q By* Meter.
6. Carefully adjust C-129 until a zero beat is heard in the calibrator headset.

Standard broadcast stations in the neighborhood of 700 kc or 1500 kc can also be used in conjunction with a radio receiver to calibrate the oscillator. The upper ends of either the 300 - 700 kc or 700 - 1700 kc ranges may be used to zero beat the Q Meter oscillator with the station carrier.



Rear of Q Meter Type 260-A Showing Access Holes of Frequency Adjustments.

Correction Curve

Should it become necessary to read the dial with an accuracy greater than factory tolerance, a correction curve plotting dial reading error against dial reading can easily be made. Using a crystal calibrator similar to the Ferris Model 33-A one can re-calibrate any range through the use of the built in multivibrator circuit, which in the case of the Ferris calibrator is locked by the 100 kc oscillator. Fundamentals of 50, 25, 20, 10 kc are thus made available to check the 50-120 kc range.

The general procedure outlined above should be followed. A pair of phones may be plugged into the output jack of the Calibrator so that beats between the Q Meter oscillator and the standard frequencies of the Calibrator can be heard.

It must be remembered that the accuracy of the correction curve is a function of the number of points checked, type of calibrator used and, of course, the skill of the operator.

**Although the C dial calibration is also important, the F dial has a greater influence on the over-all tolerance except at the low settings of the C dial. See Fall, 1955, issue No. 7 of The Notebook on "Calibration of the Internal Resonating Capacitor of the Q Meter."*

EDITOR'S NOTE

The public sees the engineer as a gnome-like creature, a sort of electronic sorcerer conjuring up electronic devices so complex that even he is startled when they work. Oblivious to the surrounding world, he plods daily from his laboratory cubicle to his home submerged in abstruse engineering problems 'til he reaches his front door where he makes

an outward show of normalcy. With an effort that produces mental fatigue he greets his wife with a perfunctory peck and his kids with a dutiful pat on the head (mixing up their names if he happens to have more than one), and enters the bosom of his family in a trance-like state from which he does not emerge 'til safely ensconced once again in his familiar world of electronics. His only "social activity," it is believed, is periodically attending meetings where technical papers full of solemn nonsense are intoned by other engineers of mien as grave as his own.

This of course is gross exaggeration. Worse. It is a myth and difficult to lay. Part of the difficulty can be attributed, we fear, to the engineers themselves. Too often the only view the public obtains of engineers or their work is in the photographs that appear in the newspapers from time to time showing an engineer at an instrument replete with dials, meters, switches, etc. The caption might say: "John Smith, development engineer for the Whynot Company, demonstrating the operation of an electric pretzel bending machine," but this belies the attitude of the engineer who, far from appearing to demonstrate anything, sits frozen in solemn disbelief of the entire proceeding. This, of course, engenders in the viewer a feeling only of pity for the engineer and a morbid curiosity for the instrument which has apparently placed its inventor in a hypnotic state.

The fact is however, despite such superficial appearances, the average engineer is a human being in the accepted (non-anthropological) sense of the word, e.g., a social being. He is a joiner. It is true he is usually a member of technical societies where "shop talk" is the rule. But that is also true of a horse breeding organization or any other homogeneous group with intellectual, economic or other interests in common.

For the benefit of that segment of the general public that may see this publication as well as for that group of engineers whose self esteem may be under a cloud because of the "queer bird" looks of the uninformed, we shall scan the professional and avocational interests of the group here at BRC with the sure knowledge it is typical of the engineering field as a whole.

Our president, Dr. Downsbrough is a member of the Board of Trustees of the Riverside Hospital here in Boonton, N. J. in addition to his membership in the Scientific Apparatus Makers Association (SAMA) where he serves as a member of the Electronics Committee. He and his family pursue the interesting hobby of Gliding, his wife holding the U. S. distance record for women. Frank G. Marble, our vice-president, is active in the Little League, a baseball organization for the pre-teen agers. He also serves as Chairman of the Exhibitors Advisory Committee of the IRE. W. Cullen Moore, our Engineering Manager, is the Scout Master in his home town as well as the Secretary of the New Jersey Chapter of the IRE. Our engineers, exclusive of their professional activities, have as motley a collection of hobbies and activities as one could find anywhere — amateur cartoonists, photographers, cabinet-makers, science fiction devotees, astronomers, Civil Air Patrol activities, and a host of others.

These multifarious interests are, we firmly believe, representative and characterize the engineering profession in general as well rounded and imaginative, ingredients essential to all creative vocations and healthy citizenship.

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FALL, 1956

NUMBER 11

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Applications of the Metal Film Gauge, Type 255-A

DOUGLAS K. STEVENS, *Sales Engineer*

In a previous issue of the Notebook,* an article was devoted to the theoretical aspects considered in the design and development of an instrument for plating and film thickness measurement. This article is intended to show how the Metal Film Gauge, Type 255-A, is being employed in the field, in research laboratories, on production lines, and as a tool for the quality control engineer.

The instrument is useful in several different types of measurements, namely: (1) the measurement of the thickness of a non-magnetic plating such as silver, gold, cadmium, chromium, rhodium, etc., on a non-magnetic basis metal such as copper, brass, aluminum, etc., (2) the measurement of the thickness of a conductive, non-magnetic material such as copper on an insulating basis material such as glass, phenolic sheet material, ceramic, etc., (3) the measurement of the thickness of any insulating film such as an organic paint on a conductive, non-magnetic basis material such as aluminum, magnesium, etc., (4) the sorting of materials by means of their electrical conductivities, (5) the sorting or matching of materials according to their magnetic properties, (6) the determination of the degree, or effectiveness of the annealing process in metals, and finally, (7) the measurement of plating thickness of magnetic materials under certain conditions. Each of these types will be discussed separately in the paragraphs which follow along with a general discussion of suitable "standards" for use with the instrument. For simplicity, each type will be considered as a flat, continuous surface at least 3/8″ on each side. Treatment of geometrical configurations other than flat surfaces will be covered later in this article.

*A-Piip, *"Determination of Metal Film Thickness,"* BRC Notebook, No. 9, Spring, 1956.

Figure 1. Author Demonstrating Sample of Organic Film on Brass Basis.

Measurements Of Non-Magnetic Combinations

As an example of this type of plating measurement, let us consider a plating of cadmium on a basis material of copper which is to cover the thickness range of 0.00025″ to 0.002″. First of all, since the Metal Film Gauge, Type 255-A, is inherently a comparator, the use of a reference standard is always required in order to make absolute thickness measurements. Such a standard is shown in Figure 2.

This sample standard card is designed to slip into a holding device on the front panel of the instrument and provides the scale for the meter. In order to measure the thickness of cadmium plating for this particular combination, we proceed as follows: Place the gauge head on the sample of basis metal (in this particular case, copper) and adjust the instrument to zero reading on the meter by means of the "Set Basis" control. When approaching zero it is well to have the "Set Standard" control turned to its extreme clockwise position in order to establish a firm zero reading. Next, place the gauge head on the sample of cadmium plating which is known to be 0.0014″ and turn the "Set Standard"

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control counterclockwise until the meter needle reads 1.4 mils. The instrument has now been calibrated and is ready for use simply by placing the gauge head on any piece of cadmium plated copper of the same basis material and taking the plating thickness reading directly from the scale. Calibrated Standards, similar to that shown in Figure 2 are available as accessories for most non-magnetic metal combinations.

The manner in which a Calibrated Standard is established may be of interest. Let us assume that a Metal Film Gauge, Type 255-A is available and that we wish to measure silver plating on brass, which is a combination frequently found in such fields as waveguide, RF fittings, etc. Let us further assume that we are interested in absolute plating thickness measurements which range from 0.0001″ or 0.1 mil to 0.001″ or 1 mil. First, we select a sample of the basis material (in this case, brass), place the gauge head on this sample and adjust the instrument to zero by the method previously described. This establishes a point (the zero point) on a calibration curve. Next, we require three (3) different thicknesses of the plated material (let us assume 0.3 mil, 0.5 mil, and 0.75 mils) which have been accurately measured by another method; chemical, optical, or X-ray etc. Using a linear scale from 0-100 for the meter, we place the gauge head on the 0.75 mil sample and set the meter needle to an arbitrary reading approximately two-thirds full scale by means of the "Set Standard" control. Without any further adjustment of controls, we place the gauge head on each of the two remaining samples and record the meter reading. Transfer the meter readings to a piece of graph paper having plating thickness from zero to 1 mil marked off as the abscissa and the meter scale marked off from zero to 100 as the ordinate. We have now described a calibration curve having four (4) known points and a smooth curve may now be drawn through the four points. All that remains is to mount the sample of basis material and the three plated samples on a standard blank card and inscribe a meter scale on the card from data taken from the calibration curve. We can now measure silver plating on brass and obtain absolute readings from the meter scale. This is the method employed by BRC in establishing reference standards that are available as accessories.

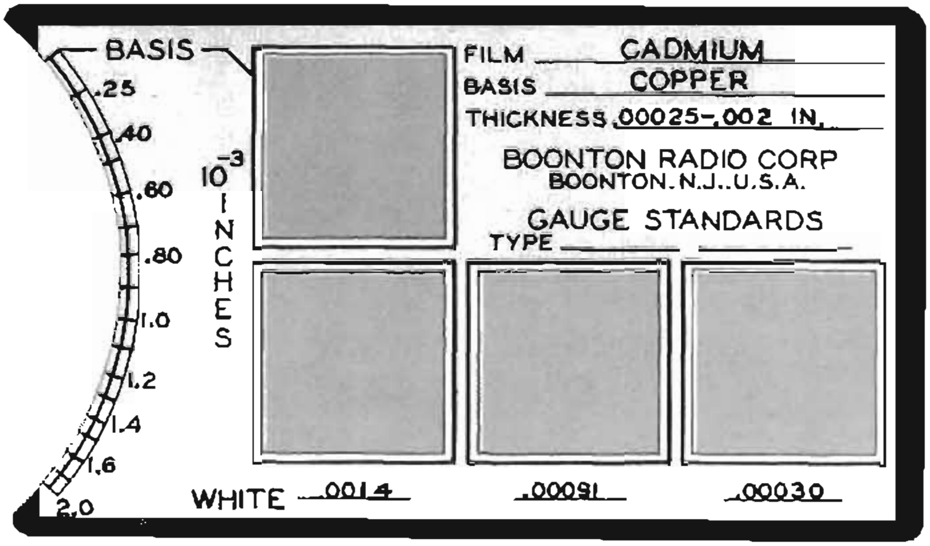


Figure 2. Cadmium Plate on Copper Basis.

Measurement Of A Non-Conductive Film Or Coating On A Non-Magnetic Basis Material

In this general classification are such measurements as organic and other non-conductive paint coatings on basis materials of aluminum, brass, magnesium, titanium, etc. In fact, any insulating coating or film on a non-magnetic basis material can be measured easily and with good sensitivity with this instrument. Again we must provide a reference standard for absolute measurements. It turns out that a reference standard may be rather easily established by using known thicknesses of mica sheets capable of resisting deformation under the gauge head. Since the instrument cannot differentiate between various kinds of insulating coatings, the mica serves as a general standard. Our field experience indicates that the range of thickness measurements between approximately 0.1 mil to several mils covers most cases. Figure 3 shows a reference standard for non-conductive coatings on a basis material of brass and covers a range from zero to three mils. This standard was established in the same manner as that ex-

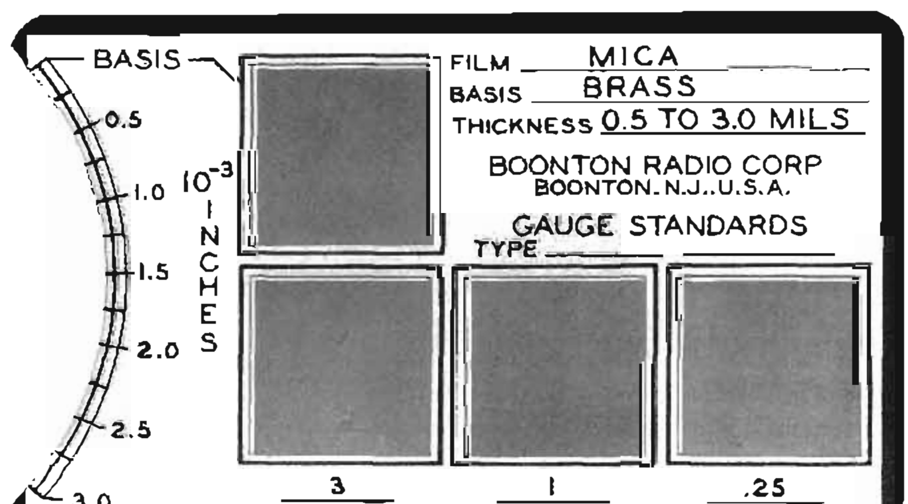


Figure 3. Non-Conductive Film on Brass Basis.

plained for silver on brass except that mica was used in place of known thicknesses of non-conductive coatings. It turns out that only one reference standard is required for many non-conductive coatings. In Figure 3 the basis material is brass. However, this same standard with its inscribed meter scale is used for non-conductive coating on basis materials of aluminum, magnesium, titanium, etc. It is only necessary to place the gauge head on the basis material and adjust the instrument to zero when changing from one basis material to another.

Conductive Layer On An Insulating Basis Material

In this category fall such combinations as copper flashing on a phenolic basis material, gold on quartz or glass, chrome on steatite, etc. Here the gauge head is placed on the insulating material and the meter zeroed as before. A standard is established with three known thicknesses of the conductive coating in the same manner as previously described. Limiting factors are very thin insulators coated on both sides and an extremely thin film of conductive coating which is under 1000 Angstrom units in thickness.

Magnetic Materials

Plating thickness on magnetic basis materials may be checked provided that the magnetic characteristics are homogeneous. This is a very broad term and is contingent upon many factors. The test for this condition is very easily made with the Metal Film Gauge, however. Simply place the gauge head on a sample of the magnetic material and adjust the meter to read zero. Now move the gauge head around on the sample (and other samples) to see if the zero point remains reasonably constant. If it does, plating thickness can be measured. If it does not remain constant, plating cannot be measured. Many of the better grades of steel, for example, exhibit this constant zero characteristic. Since the instrument is sensitive to changes in magnetic characteristics and changes in electrical conductivity, this feature leads naturally to the next application.

Sorting And Matching Of Materials

In some applications it is important that materials exhibit the same electrical and magnetic properties. The Metal Film Gauge, Type 255-A is a valuable tool in checking electrical conductivity and matching materials according to their magnetic properties. No standard other than a sample of material known

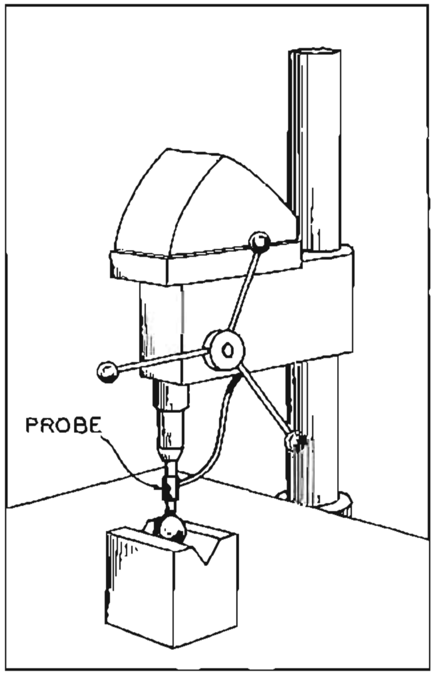


Figure 4. Jig and Fixture Combination.

to exhibit the desired characteristics is required here since we are interested only in relative readings. For sorting according to electrical conductivity, the "Set Base" control could be calibrated in conductivities in place of a linear scale from 0-100, and materials sorted accordingly. For matching magnetic properties it is only necessary to place the gauge head on a sample of the material, establish an arbitrary reading on the meter and match materials according to the deviation from the arbitrary setting on the meter scale. It is also possible to check the degree of annealing in such metals as steel, berylium, etc. Since the annealing process electrically "softens" the material, an arbitrary meter reading can be established with the gauge head on a sample of untreated material, and the deviation noted when the gauge head is placed on the annealed samples.

Measuring Irregular Surfaces

The gauge heads which are furnished with the instrument were designed for measurements on either flat or cylindrical surfaces with relatively large diameters. The instrument is not limited to this kind of surface area however, and other geometrical configurations may be handled provided there is at least 3/8" of continuous surface area available. The problem is best handled by providing a jig or fixture for holding the probe itself (which can be removed from the gauge head) and, if necessary, a holding fixture for the part to be measured. Figure 4 is a sketch of one possible way in which the jigging problem may be handled. It is, of course, necessary to establish reference standards which have the same geometrical pattern.

Standards For Production Line Use

When the Metal Film Gauge, Type 255-A is used on a production line, the reference gauge standards usually take a form similar to that shown in Figure 5. Here we are interested in a "Go-No-Go" type of test rather than absolute measurements. In this case let us assume that we are concerned with measurements of a plating or coating thickness the acceptable limits of which have been specified as 0.6 to 1.2 mils. Two samples of the coated or plated material — one representing a maximum thickness and the other a minimum

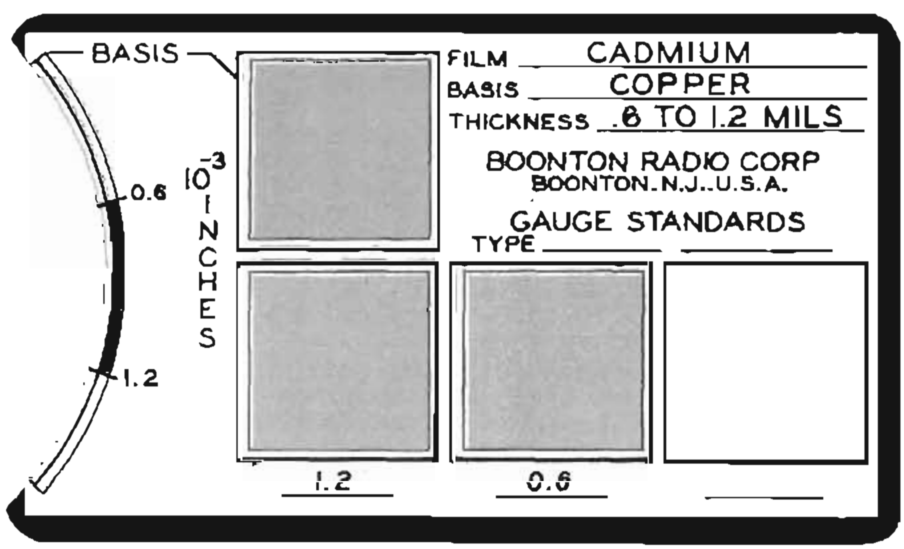


Figure 5. Production Card for "Go-No Go" Test.

thickness — are mounted on a sample card along with a sample of the basis material. The gauge head is placed on the basis material and the instrument adjusted to read "zero" on the meter scale by means of the "Set Base" control. The gauge head is now placed on the maximum thickness sample and an arbitrary meter reading of approximately ¾ full scale is established by means of the "Set Standard" control. This spot on the meter scale should be marked on the scale since it represents the upper limit — in this case 1.2 mils. Without any further adjustment of controls, place the gauge head on the minimum thickness sample and mark the spot where the needle falls on the meter scale. We have now established the two limit points, and this portion of the meter scale can be marked off in a suitable contrasting color. This type of measurement greatly simplifies making a gauge standard and provides the production line with a tool it can use rapidly with a minimum of indoctrination.

Increasing The Sensitivity

In the case of extremely thin films the apparent, or usable, sensitivity of the instrument is sometimes greatly reduced. A technique has been developed which somewhat improves the condition. As an example, consider the measurement of a thin layer of gold, silver, etc., on a basis material such as glass or ceramic. Under certain conditions it will be found that when the gauge head is adjusted to read zero on the sample of glass or ceramic and is placed on the gold or silver coating, the meter needle will move only a short distance away from its zero point even with the "Set Standard" control tuned to its maximum clockwise position. This loss in apparent sensitivity limits our range of thickness measurements and the readable accuracy because of the small spread in meter movement. This condition can be improved by adjusting the gauge head to read zero on the coated sample instead of on the basis material. The instrument will now read any deviation from this thickness as the gauge head is placed on other samples of the coated material. However, care must be exercised in this technique. The instrument includes a rectifier which is designed to always read deviations in the same direction on the meter. Suppose we zero the meter on a coated sample which is known to be 0.02 mils in thickness. Further suppose that we next measure a coated sample which is 0.015 mils in thickness. We will read a certain deviation on the meter scale. Now,

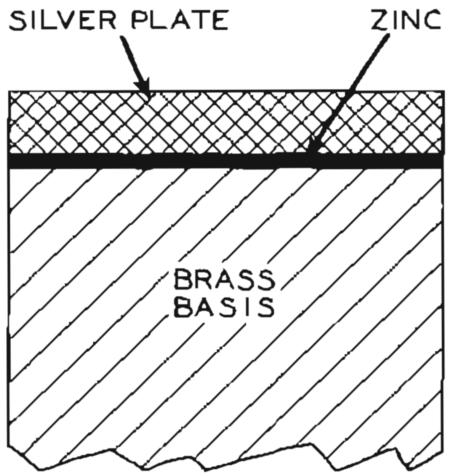


Figure 6. Cross-Section View of Multi-Layer Sample.

measure a sample which is 0.025 mils in thickness. The deviation read on the meter scale will be identical with that read on the 0.015 mil sample. If one is aware of this source of error the technique is very useful since it has the effect of "spreading out" the readings on the scale. Therefore, when the zero point is established on a plated sample, care must be taken that all subsequent measurements are made on samples which are all thinner in plating or all thicker in plating in order that an amgibuity is not introduced.

Multi-Layer Measurements

In instances where there is more than one layer of plating material, there are several ways in which the desired thickness measurements can be made. Sometimes a "flashing" of rhodium or silver is plated on a basis material before the final coating is applied. Figure 6 represents such a case. Here the basis metal is brass with a "flashing" of zinc and a plating of silver. If it is desired to measure the thickness of the flashing this must be accomplished in the usual manner before the silver plating is applied. If the zinc coating is reasonably constant and it is desired to measure only the silver plating the gauge head may be placed on the brass and the meter adjusted to zero. The gauge head may then be placed on the silver plated sample and the instrument will then measure the combined thickness of both the zinc and the silver. Subtracting the constant thickness of zinc yields the thickness of silver plating. Any error introduced by this method will ordinarily be small since the ratio of silver plating to the zinc "flashing" is ordinarily large. However, this error may be eliminated completely by zeroing the gauge head on the zinc instead of on the brass. This has the effect of "washing out" any discrepancies in the zinc "flashing". All multi-layer platings may be handled by this general approach.

Selection Of Proper Gauge Head

The foregoing discussion applies to either of the two gauge heads which are furnished with the Metal Film Gauge, Type 255-A. One of the gauge heads operates at a frequency of 500 kc while the other operates at 8 mc. These are coded red and white respectively. The gauge head to be used should be chosen according to the thickness of the coating and composition of the film-basis combination to be measured. In general, a thin coating requires use of the white, or 8 mc, gauge head. Coatings of lower conductivity also require the white gauge head while those of higher conductivity with comparable thickness and having the same basis material require the use of the red, or 500 kc, gauge head.

Summary

The Metal Film Gauge, Type 255-A provides a fast, accurate, and non-destructive method for the measurement of plating thickness of non-magnetic material combinations. Of particular interest is its ability to measure the thickness of non-conductive films in the protective coating field. Other applications include the sorting of materials according to their electrical conductivities, the matching of magnetic properties, and the effect of the annealing process on metals.

THE AUTHOR

Douglas K. Stevens is a graduate of Catholic University, Washington, D. C. where he received a B.E.E. degree. He has had a wide range of engineering experience: Electronic Engineer with the National Bureau of Standards, Washington, D. C.; Project Engineer with the Research Division of the United Shoe Machinery Co., Beverly, Mass.; Test Equipment Engineer with the Western Electric Co., Lawrence, Mass. and a miscellaneous experience with other engineering firms. During the war, he did radar work as a member of the U. S. Air Force. Since joining the staff of Boonton Radio, Mr. Stevens has been active in field work on customer problems and special applications of Boonton Radio Corporation equipment. Mr. Stevens is married, has two young children and lives in Boonton, N. J.

A VHF FM-AM Signal Generator System

JAMES E. WACHTER, *Project Engineer*

The 202-E FM-AM Signal Generator, covering the frequency range of 54 to 216 mc, incorporates many of the features most desired in a VHF signal generator; some of the more prominent of these being the high degree of stability of the r-f carrier frequency, the constant frequency deviation sensitivity with respect to the carrier frequency, the ability to deviate the carrier either plus or minus by known increments, and the continuously variable, calibrated output attenuator system.

How these features are realized is most easily explained by a discussion, in short, of how the instrument is constructed and of how it functions.

General Features

Outwardly the instrument is composed of two units, the signal generator proper and its power supply, shown (with accessory 207-E Univerter) in figure 1. The design of the cabinets is such that the instrument may be used as conventional bench equipment or by simply removing the cabinet end-bells, it may be mounted in a standard 19 inch rack as shown. All operating controls and indicators are functionally located on the front panels and those which are calibrated are direct reading.

The heart of the 202-E Signal Generator is, of course, the RF Assembly, shown less shielding in figure 2. The rugged construction of the supporting frame and the close mechanical tolerances applied to the four section variable capacitor are in part responsible for the high degree of stability of the r-f frequency.

Functional Description

Referring to the simplified schematic diagram of the RF assembly, figure 3, the functions of the various circuits is more readily understood. For FM, an audio voltage is applied directly to the grid of the reactance tube. The reactance modulator operates as a controllable inductance in parallel with the tuned circuit of the r-f oscillator and provides from 0 to ±240 kc deviation of the generator output frequency. The deviation is monitored by the front panel modulation meter. To produce constant frequency deviation sensitivity

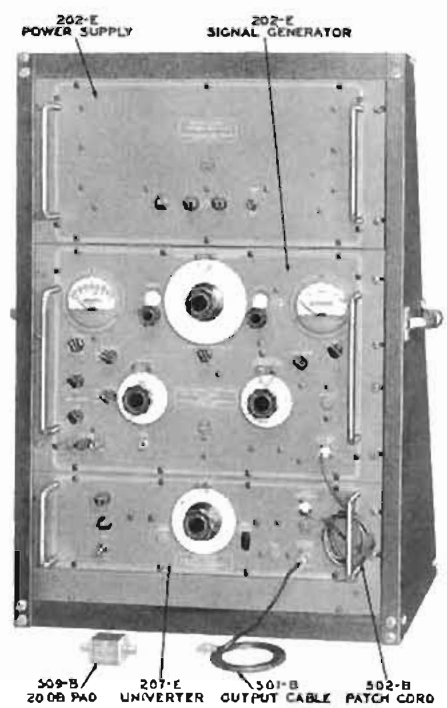


Figure 1. The RF Signal Generator, Type 202-E with Accessory Univerter, Type 207-E.

over the tuning range of the oscillator, the amount of inductance injected by the reactance tube is made to vary directly as the oscillator frequency. This is accomplished by the phase-shifter network composed of R1, R2, C1 and the reactance tube grid to cathode capacitance and grid to plate capacitance. The attenuation of this network increases as the oscillator frequency increases causing the degree of modulation

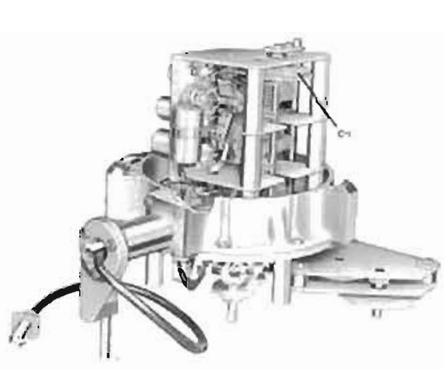


Figure 2. RF Assembly, Showing Supporting Frame and Slotted Rotor of Deviation Capacitor.

effected by the reactance tube to decrease with frequency so as to maintain constant deviation. To account for variations in components and to provide a precise deviation calibration, C1 is ganged to the oscillator tuning capacitor and is adjusted over the entire tuning range for the desired constancy of deviation.

The deviation sensitivity of the FM modulation system, as a function of frequency, is flat to within ±1 db from 30 cps to 200 kc. and the overall FM distortion at 75 kc deviation is less than 2%.

D-C signals from a battery source (for maximum stability) may also be applied to the grid of the reactance tube to shift its operating point and thus shift the oscillator frequency by discrete amounts, either plus or minus. Control is effected by operation of a front panel switch and associated precision resistors. By this method the output carrier frequency may be shifted in steps of ±5, 10, 15, 20, 25, 30, 50 and 60 kc on the 108-216 mc range and half of these values on the 54-108 mc range. The relative accuracy of the steps is ±1.5%. Because of the low current drain, the life of the battery is equivalent to its normal "shelf life". An uncalibrated fine tuning control, operating in conjunction with this incremental tuning circuit permits continuous tuning over a range of about 20 kc on the high frequency range and 10 kc on the low frequency range.

The r-f oscillator is a conventional tuned plate triode oscillator circuit covering the frequency range of 27 to 54 mc. The circuit is properly compensated for minimum drift and is tuned by variable capacitor C2, which is adjusted for an output frequency accuracy of ±1%.

The oscillator output is applied to a self-biased class C frequency doubler, which makes possible the low oscillator frequency range and also provides the required isolation between the oscillator and the output stage necessary for the desired frequency stability of 0.01% per hour.

The output stage operates class C and functions as an amplifier for the low frequency range of 54 to 108 mc, and as a frequency doubler for the high frequency range of 108 to 216 mc. This change in operation is accomplished by operating the front panel range switch, which grounds either of two taps on the output tank coil. In one position the inductance of the coil is such that the tank circuit is resonant at the fundamental frequency of the previous stage and in the other position the inductance is such that the tank is resonant at the second harmonic of the frequency of the previous stage. Damping of the output tank circuit is sufficient to reduce spurious signals to more than 30 db

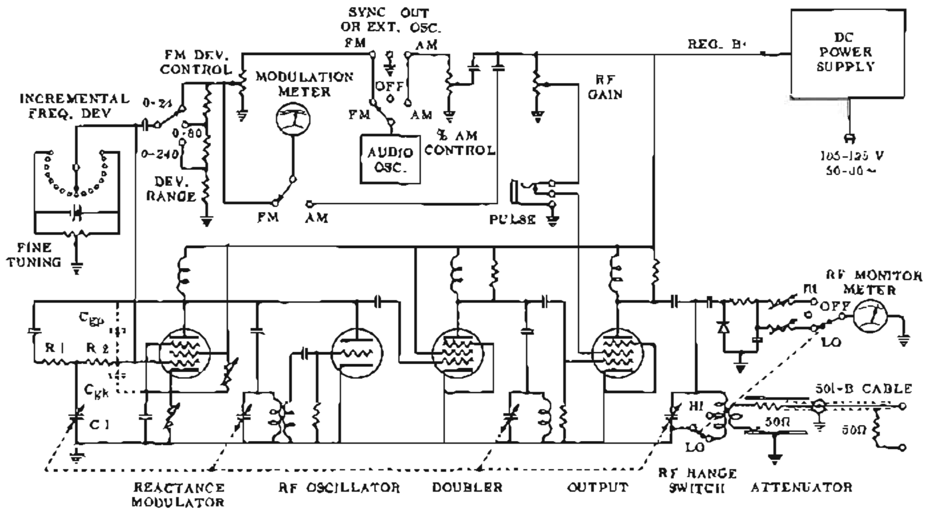


Figure 3. 202-E Simplified Schematic Diagram.

below the desired signal and to restrict, to about 2%, the amplitude modulation present at 75 kc deviation.

Up to 50% amplitude modulation is possible by modulating the screen circuit of this final r-f stage and is monitored by the front panel modulation meter. The modulating power required for this being comparable to that needed for frequency modulation, the same R-C audio oscillator is used for both types of modulation. The distortion of a 50% AM modulated signal is less than 8% and due to the buffer action of the doubler stage, the spurious frequency modulation is held to a minimum. The amplitude modulation system is flat to within ±1 db from 30 cps to 200 kc.

Pulse Modulation

Square wave or pulse modulation of the 202-E is also possible by connecting an external modulation source through the front panel jack to the screen of the final stage. When this connection is made, the modulation meter and screen circuits are disconnected. Under these conditions the rise time of the modulated carrier envelope is less than 0.25 microseconds, and the decay time is less than 8 microseconds.

A piston type, mutual inductance attenuator having an internal impedance of 50 ohms, is coupled to the output tank circuit to provide continuously adjustable attenuation. Because the rate of attenuation is a function of the inner diameter of the attenuator tube, this attenuator can be made to be quite accurate. With the 50 ohm terminated cable, type 501-B, (supplied with the instrument) attached to the 202-E output jack, the attenuator dial indicates, to an accuracy of about ±10%, the voltage at the cable termination. (The generator open circuit output voltage is given by twice the attenuator dial reading). The attenuator dial is calibrated with two equivalent scales, 0.1 to 2000 microvolts and 140 to 14 db below 1 volt.

Simultaneous FM and AM Modulation

Simultaneous FM and AM modulation may be obtained from the 202-E by using an externally connected, low distortion, audio oscillator to provide the FM modulating signal and the internal audio oscillator for the AM modulating signal. The only requirement is that the external oscillator be capable of providing about 5 volts across 1500 ohms, the FM requisite for 240 kc deviation. In use the external oscillator is connected to the front panel FM binding posts and the 202-E modulation selector set to AM; the modulation meter can be switched from AM to FM and the levels of each type of modulation independently set.

It might be noted here that the internal audio oscillator provides the following frequencies to an accuracy of ±5%: 50, 100, 400 cps, 1, 5, 7.5 and 10 kc. In addition, a 60 kc signal accurate to ±2% is available specifically for use in calibrating the d-c incremental frequency circuit.

All electrical connections to the shielded RF Assembly, excepting that to the output attenuator, are made through a low-pass filter, which prevents stray r-f currents from escaping from the RF Assembly.

The voltages supplied from the power supply are all d-c and those for use in the RF assembly are regulated. The power supply operates from 105-125 volts, 50-60 cps.

Frequency Converting Accessory

The 207-E Univerter is a unity gain frequency converter designed to provide frequency coverage below the range of the 202-E Signal Generator. When used in conjunction with the 202-E, the Univerter covers the range of 0.1 to 55 mc; and the two instruments give complete coverage from 0.1 to 216 mc.

The 207-E matches the 202-E in appearance (figure 1) and may also be used either as bench type equipment or rack mounted. All controls and input and output connectors are located on the front panel.

To use the 207-E it is only necessary to connect the 501-B output cable to the Univerter unity gain output and connect the Univerter input to the 202-E Signal Generator output, using the 502-B patch cord supplied with the Univerter. Connected in this manner the 207-E Univerter does not appreciably alter the FM and AM characteristics of the 202-E Signal Generator.

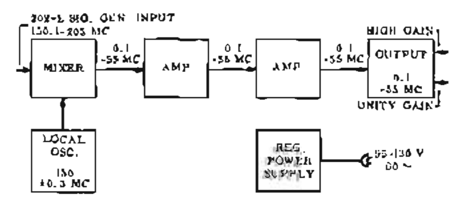


Figure 4. 207-E Block Diagram.

A 20 db attenuation pad, Type 509-B, is supplied with the Univerter for use where the signal level required is low compared to the constant noise level of the Univerter. Used at the output of the 207-E, the pad attenuates both signal level and constant noise level, thus permitting the use of a higher input signal and improving the signal to noise ratio.

Lubrication Of Turret And Switch Contacts

LAWRENCE O. COOK, *Quality Control Engineer*

When frequency ranges must be changed, as in an RF oscillator, a means of switching the circuits is needed. This is usually accomplished by means of movable metallic contacts. Such contacts normally introduce an element of unreliability, i.e., contact resistance uncertainty results in fluctuating oscillator voltage, and sometimes fluctuating frequency, which, in turn, results in unstable meter readings. These effects can usually be minimized by the use of a lubricant on the contact surfaces.

Contact Types and Service Conditions

In this article we wish to discuss the effects of various lubricants on electrical measuring instrument, low-power rf oscillator contacts of the wiping type. A turret containing coils connected to brass pin contacts for the individual frequency ranges is rotated until the contacts for the range to be selected engage stationary clip contacts of beryllium copper. The clip tips wipe or rub opposite sides of the pin during the additional slight rotation of the turret required to reach the indexed position. The contacts then remain stationary during use until another frequency range is selected, i.e., for a period ranging from a fraction of a minute to one week or longer.

The circuit resistance in the instrument stability may be as low as 0.3 ohm, hence the contact resistance stability becomes important.

Contact resistance stability may worsen when the instrument reaches an age of about one year. Improvement of a temporary nature may sometimes be effected by repeated operation of the faulty range switch (this "wipes" the contacts).

Test Method

For test purposes, two turret contact pins were connected together and meshed with two stationary contact clips. The clips were connected by short wire leads to a Wheatstone Bridge to allow measurement of the dc resistance of the resulting "short-circuit" (designated as a "pair" in the Figures 1 and 2). Remeshing and re-measurement were repeated several times, the extremes of the resistance values being plotted. Four con-

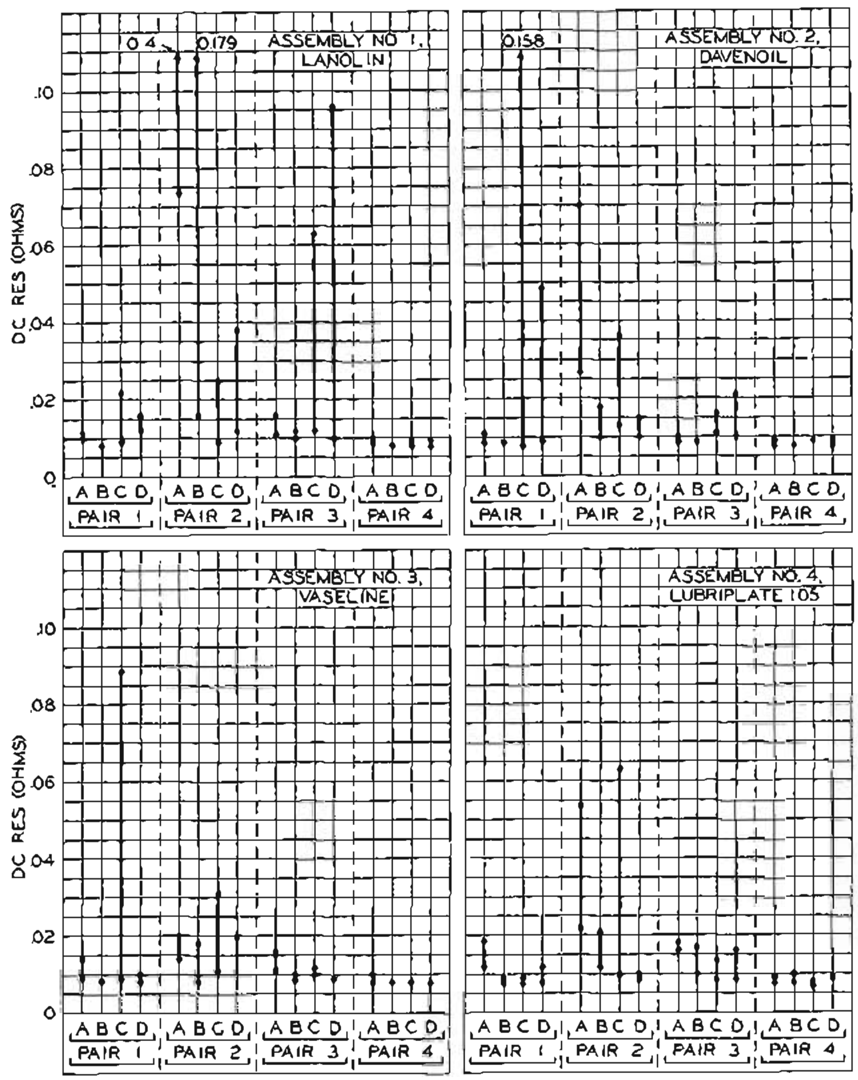


Figure 1. A—Un-Lubricated, B—Following Lubrication, C—Following Three Months Storage, D—Following Eight Months Additional Storage.

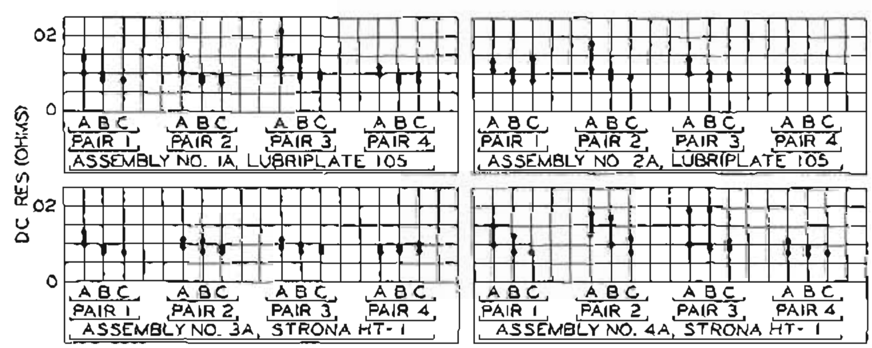


Figure 2. A—Un-Lubricated, B—Following Lubrication, C—Following Six Months Storage.

RF Oscillator Contact Assemblies — Contact Resistance Measurements Versus Lubricants. Plots Show Range of Resistance Measured for Repeated Meshing of Contacts.

tact pairs were tested on each assembly.

To determine the effect of lubricants, the contact resistance measurements were made first on dry contacts. The contacts were then lubricated and re-measured Following a prolonged period of inactive storage, the measurements were repeated.

Test Results

Giving preference to the lubricants showing consistently lower contact resistance we have summarized the test results on the five materials as follows.

CONTACT RESISTANCE PREFERENCE	LUBRICANT
1st	Lubriplate 105
	Strona HT-1
2nd	Vaseline
3rd	Lanolin*
	Davenoil

Lubriplate 105 and Strona HT-1 are considered equal on the basis of contact resistance and both are manufactured for lubricant use. Strona HT-1 showed a slight disadvantage of "stringiness" during application. Vaseline showed somewhat higher contact resistance and is not manufactured for use as a lubricant.

As a result of these tests made on a particular type of contact, our choice favors Lubriplate 105 and Strona HT-1, these materials tieing for first place.

**Anhydrous lanolin dissolved in carbon tetrachloride for ease of application.*

EDITOR'S NOTE

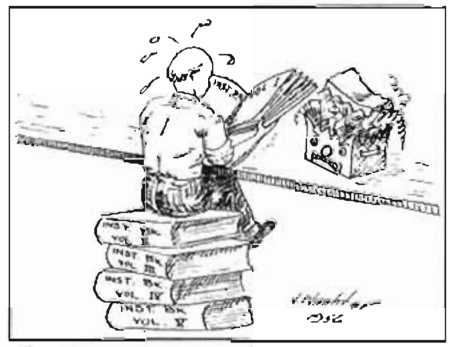
The prevalent attitude towards instruction manuals may be neatly wrapped up in the cynical saw, "If all other methods fail, recourse to an instruction manual may be mandatory." Entirely justified viewpoint in our opinion, for this reason: the role and character of the instrument the manual is supposed to describe is seldom self-consciously considered in the preparation of the manual. In many cases the instrument in question is a piece of test equipment of which the engineer may have a dozen or more types. He simply hasn't the time to digest the information in the form in which it is presented. This is the decisive fact that should determine the kind of material, its quantity and arrangement, and the general format of the manual.

Judging by the tome-like and overblown appearance of many manuals, one can make better use of them as equipment props, (in many cases their only useful function). With such manuals, engineers usually wait for the moment just before the instrument is about to disappear in a cloud of smoke. In other cases the manual is resorted to only after the instrument has been completely disassembled and lays strewn about in impossible confusion.

Organization of the Material

With the specific function of the instrument held firmly in mind, the preparation of the manual can be approached in a more confident manner. In general a manual of instruction for a piece of test equipment (which is simply an electronic tool) should be prepared along these general guide lines:

Instruction Manual should be divided into two main sections. Section one (in our estimation the more important of the two) should be as compact and abbreviated as possible and subdivided into four general categories: (1) *Essential* operating information, (2) Special applications, (3) Tabulation and location



of components *most likely* to need attention, (4) Annotations copiously sprinkled throughout text with *specific* references to material found in the second half of the manual where a full and conventional treatment of the instrument is given.

Such an approach separates the immediate, practical knowledge from the generally useful but otherwise frustrating information. This will be greatly appreciated by an engineer who, for example, wants to find the location of a blown fuse in a hurry but must waste precious time wallowing through pages of extraneous material to find it.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Techniques of Signal Generator Inspection

LAWRENCE O. COOK, *Quality Control Engineer*

A signal generator provides a test signal of controllable frequency, voltage, and modulation and may be judged in terms of various characteristics, such as frequency stability, accuracy of frequency calibration, lack of spurious frequencies, lack of incidental amplitude modulation or incidental frequency modulation, low distortion, etc. The relative importance of the individual characteristics depends greatly, of course, on the specific application in which the generator is to be used.

Table 1 lists various signal generator characteristics of interest. Some items are primarily a fundamental characteristic of the design and need not be tested on an individual basis after the prototype run has been developed. Examples of these are RF output frequency stability, RF output impedance, residual FM, and frequency response of the FM system. Such items need to be tested on only one instrument in many and fall logically within the scope of a Quality Control program.

Other items require inspection on each unit, with corrective adjustments often being needed. Examples are RF output frequency accuracy, RF output voltage accuracy, modulation accuracy and harmonic distortion. These items are included, furthermore, in a Quality Control program calling for "spot checks".

We now propose to discuss some of the testing and calibrating procedures used in making certain that the desired characteristics are obtained in the generators produced. These are procedures generally applicable to the types of generators discussed and, in most instances, are applicable to each unit of a given type. Our aim is to produce units which will meet and maintain closer tolerances than those advertised.

Figure 1. Adjustment of Basic Voltage Standardization Equipment by the Author. Normalized Atmospheric Conditions (Temperature and Relative Humidity) for Improved Stability are Maintained by Air-conditioning Equipment in Background.

TABLE 1

1. *RF Output Frequency Characteristics:* Range • Dial Calibration Accuracy • Accuracy, crystal standardized (internal) • Stability • Crystal Controlled • Vernier Dial • Incremental Frequency Switch • Fine Tuning Control • Spurious Output Frequencies.
2. *RF Output Voltage Characteristics:* Range • Accuracy • Impedance.
3. *Frequency Modulation Characteristics:* Deviation Range • Accuracy • Distortion, harmonic • Frequency Response • Residual • Microphonism • AM on FM, spurious.
4. *Amplitude Modulation Characteristics:* Range • Accuracy • Distortion, harmonic • Frequency Response • Phase Shift • FM on AM, spurious • Pulse • Square Wave.

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5. *Internal Modulating Oscillator Characteristics:* Frequencies available • Frequency accuracy • Voltage available externally.

6. *Modulation Characteristics, miscellaneous:* External Modulation Input, AM • External Modulation Input, FM • Simultaneous AM and FM.

7. *Swept RF Characteristics:* Range of Sweep Widths • Linearity of Swept Frequency • Sweep Repetition Rate • Flatness of Swept RF Output Voltage Level.

RF Range and Accuracy

Frequency indicating dials, tuning coils, and tuning capacitors are usually fabricated to prescribed physical layouts and dimensions. When these uniform components are installed the assembled instruments will lie "within calibration range" of the manufacturing adjustments provided.

For calibration purposes the generator output frequency is usually referred to a commercially available crystal type calibrator, accuracy rating ±0.002%. "Zero beats" are obtained at a sufficient number of generator dial points (usually 6 or 8 per frequency range) to insure accurate calibration. (Individually marked dials must, on the other hand, be calibrated and marked at each division line.)

In instances where the frequency interval between "zero beat" points is very small in comparison to the frequency being checked, a "single response" type of frequency indicator must be used to determine that the correct "zero beat" point has been located. A simple frequency-calibrated tuned circuit and diode voltmeter (or "Megacycle Meter") which may be coupled to the generator output and resonated at any one of the "zero beat" points is usually satisfactory for this purpose.

The crystal-type calibrator, with its accuracy rating of ±0.002%, offers accuracy entirely adequate for checking signal generators having published ratings of ±0.5% or ±0.1%. This type of calibrator may be checked against WWV if desired.

Signal generators having crystal controlled output frequencies are checked for accuracy against specially built crystal calibrators which are adjusted against WWV.

RF Output Voltage Range and Accuracy

In instrument production assurance of the proper voltage range and accuracy usually involves a measurement of output voltage at a high voltage level, and assurance of attenuator linearity or dial tracking accuracy.

Output voltage measuring equipment has an input impedance of 50 ohms, thus providing a normal load and duplicating the input impedance of the accessory 501-B Cable at the signal generator panel connector.

RF Output Calibration — Method 1 . . . A three stage process is involved in our standardization of RF output voltages (see Figure 2).

a. *Basic Standardization* (performed in a temperature-controlled and humidity-controlled room). The current from a DC supply is passed through a thermocouple ammeter and resistor series connected. The DC voltage developed across the resistor is measured on a basic standard consisting of a Weston Model 4 Standard Cell (unsaturated cadmium type) and Leeds and Northrup Type K-2 Potentiometer. Over-all accuracy is 0.1%.

b. *DC to AF Transfer.* Current from a low distortion AF oscillator is substituted for DC and adjusted for the same thermocouple ammeter reading as with DC. The AF signal voltage now developed across the resistor and applied to the electronic voltmeter is equal in magnitude to the previous DC voltage (the electronic voltmeter loading is negligible). The AF voltage accuracy is 0.5%.

c. *RF Standardization.* The specially-built bolometer bridge responds equally to an internally generated low distortion AF voltage, or to RF voltage from the signal generator being calibrated which is substituted for the former. The electronic voltmeter from (b) is used to standardize the AF voltage and, hence, to calibrate the signal generator, usually at 50 KμV RF and at several output frequencies. The bolometer bridge input impedance is equal to the signal generator output impedance, 50 ohms.

Accuracy of the RF voltage measurement, including reading errors, is normally 2% to 3%.

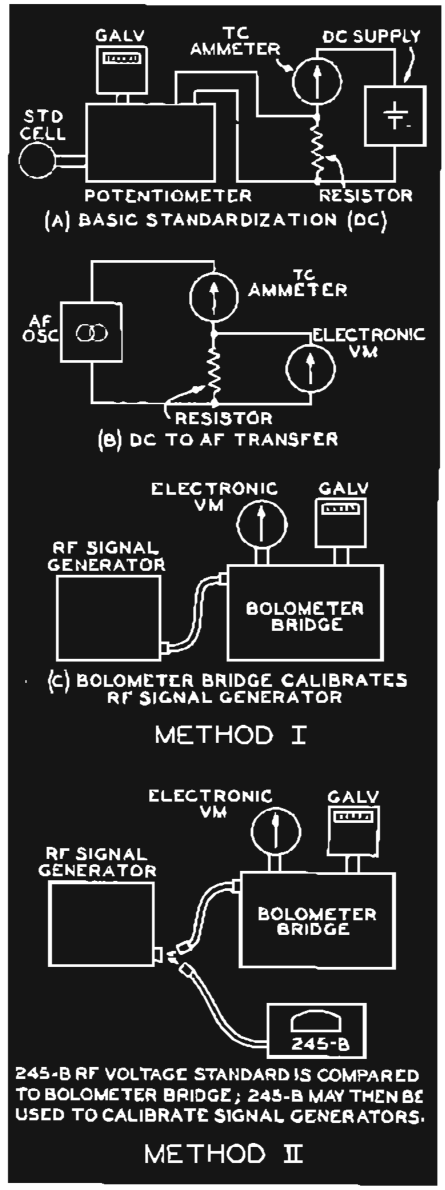

Figure 2. Standardization of RF Output Voltages.

RF Output Calibration — Method 2... RF Voltage Standard Type 245-B[1,2]. This instrument may be used as a calibrated voltmeter of approximately 50 ohms input impedance, and provides a direct check on the accuracy of signal generator output voltages in the 50 KμV region.

Our calibration procedure starts with the Standard Cell but differs somewhat in detail from that used for the bolometer bridge. The 245-B input voltage accuracy rating is ±10% for frequencies of 100 kc to 300 mc; however, the units we use for signal generator calibration are additionally calibrated by direct comparison to the standardized bolometer bridge (RF comparison as in Figure 2), thus attaining accuracy approaching that of the bolometer bridge.

Figure 3. RF Comparison of 245-B RF Voltage Standard to Bolometer Bridge.

Attenuator Linearity . . . The linearity, or dial tracking accuracy, of piston (mutual coupling) attenuators is primarily a matter of mechanical design and control of mechanical tolerances. RF leakage must also be held to low levels.

Low-level attenuator output voltages, i.e., in the one microvolt region, may be electrically measured by means of the RF Voltage Standard Type 245-B[1,2] mentioned above. The 245-B accuracy, when used in accordance with the instructions provided for this application, is:

$\pm 10\%$	100 kc to 100 mc
$\pm 15\%$	to 500 mc
$\pm 20\%$	to 1000 mc

Figure 4. Measurement of FM Deviation and FM Distortion of Type 202-E FM-AM Signal Generator by Use of FM Linear Detector and Associated Apparatus.

Frequency Modulation Range, Accuracy, and Harmonic Distortion

A single "setup" of test apparatus is arranged to measure these three interdependent characteristics. (Figure 4).

The basic test apparatus is a specially built FM Linear Detector having adequate range of FM deviation and AF response. This Detector (including built-in heterodyne oscillator and mixer for signal frequency conversion) connects to the modulated RF output of the signal generator and delivers de-modulated output voltage faithful in amount and harmonic distortion to the amount of FM deviation and harmonic distortion present in the signal generator output.

Associated apparatus consists of an electronic voltmeter and distortion analyzer for reading the Detector output voltage and harmonic distortion.

For calibration of the output voltage readings vs FM deviation the Detector input is connected to the output of a signal generator calibrated by the Bessel Zero method[3]. The calibration thus relies basically upon audio modulating frequencies which can be measured accurately (e.g. by use of a frequency counter). An alternative calibration method employs the Weston Model 4 Standard Cell, Leeds and Northrup Type K-2 Potentiometer, and crystal-standardized signal frequencies. The Detector employs an electronically regulated power supply and has excellent long term stability.

In production testing each generator is checked and adjusted (within the RF unit) at several output frequencies for the correct value of FM deviation. Adjustment for acceptable distortion is also made within the RF unit.

Amplitude Modulation Range, Accuracy, and Harmonic Distortion

For the measurement of the *depth of amplitude modulation* either of two methods may be employed.

Method a. This is the familiar trapezoidal pattern method. A frequency converter, providing conversion of the signal generator output frequency to a frequency suitable for the oscilloscope, and having a linear input voltage vs output voltage relationship, is interposed between the signal generator output and the oscilloscope input.

Method b. This method is particularly adaptable to generators having a modulation meter actuated by the modulation component of the RF carrier, in addition to the usual RF carrier monitor meter.

The basic formula is

$$\% \text{ AM} = \frac{141.4 \quad V_{ac}}{V_{dc}}$$

where V_{ac} and V_{dc} indicate respectively the AC voltage (RMS) and DC voltage components of the demodulated carrier voltage.

For best accuracy a correction, usually of a minor nature, for the rectifier characteristic curvature should be determined.

Harmonic distortion is usually measured by connecting a distortion analyzer to a suitable point in the RF monitor circuit and reading the distortion present in the AC portion of the demodulated carrier voltage.

Microphonism

FM microphonism can be very troublesome when making tests involving small values of FM deviation. Bench vibration may be conducted to a signal generator resting on the bench, or vibrations may be set up within the generator by airborne sounds within the room in which the generator is used.

Within the generator a major item of susceptibility is the oscillator section of the main tuning capacitor. Vibrations transmitted to the plates or blades of this variable capacitor cause undesired FM of the signal generator output frequency, thus interfering with normal usage of the output signal.

Analysis of this problem indicates that maximum cancellation of vibration-caused capacitance variations is obtained when the plate spacing is uniform throughout the capacitor. In practice visual inspection of plate spacing, a very important but not self-sufficient operation, is required to be followed by an electrical check of vibration-induced FM, additional adjustments of the plate spacing often being necessary. It is emphasized that this adjustment is a specialized technique requiring considerable time and experience to acquire the necessary skill.

Spurious AM on FM

In our types 202-B, -C, -D, -E, and -F Signal Generators, which embody a frequency modulated oscillator followed by amplifying or frequency-doubling stages, the generation of undesired amplitude modulation when employing frequency modulation is largely a problem of interstage tracking accuracy. For example, if the RF output stage tuning is "out of track" as referred to the oscillator frequency, the output stage will be operated (or "swung") over a sloping portion of its resonance curve and spurious amplitude modulation will be generated (see Figure 5)[4].

Tuning adjustments of the individual stages are [illegible] made during the inspection of each unit to obtain satisfactory interstage tracking accuracy. The tank circuit resonance curve is observed on an oscilloscope while the generator, operating on FM, is tuned across its range. In this manner spurious AM as a result of FM is kept to a low value, although direct checks of spurious AM are not routinely made.

RF Unit Adjustments

The interlocking nature of various adjustments made within the RF unit (i.e. by potentiometer adjustment or bending of tuning capacitor plates) appears to be worthy of mention.

Examples, particularly applicable to FM Signal Generators, are:

Adjustment For	***Affects***
(a) FM Deviation calibration	Output frequency calibration accuracy and interstage tracking accuracy.
(b) Microphonism	
(c) Interstage tracking accuracy	Spurious AM on FM
(d) Output frequency calibration accuracy	Interstage tracking accuracy (spurious AM on FM and maximum RF output voltage)

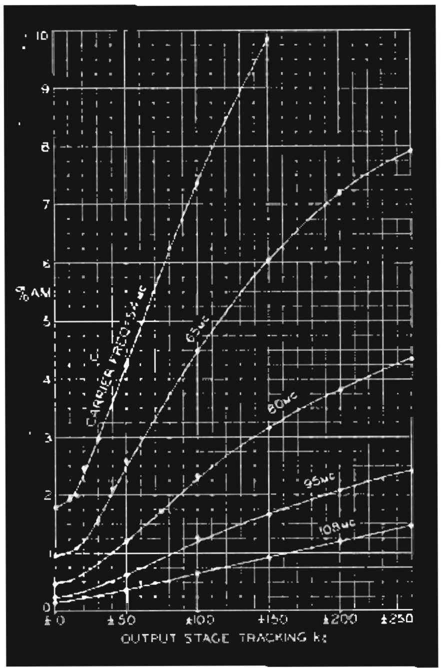

Figure 5. Showing Effect of RF Output Stage Tracking (as Referenced to Oscillator) on Spurious AM on FM. (Type 202-E Sig. Gen.; FM Dev. = 75 kc.)

It is obvious that adjustment and correction of a particular type of RF unit fault may easily introduce a different type of fault. This is not an indication of technical inability on the part of the person performing the adjustments, but is the result of interaction of adjustments.

We wish to hereby discourage user field service of RF units because the end result of such service may, through unfamiliarity with the interactions involved, yield an instrument which is operating below peak performance. This would be to the user's disadvantage as well as ours.

Tube Stability

The stability of vacuum-tube characteristics plays an important part in the long-term stability of signal generator operation. This is particularly true in respect to the maintenance of adequate RF output voltage (RF monitor meter indication).

For tubes operating in class-C RF output amplifier service we have found it necessary to adopt a rack aging procedure employing a pulsed cathode current derived from a pulse generator. Pulse aged tubes are later installed in the instruments in which they are to be used, and operated before and after instrument calibration. This procedure eliminates a varying percentage of tubes from class-C service but has been found to yield greatly improved reliability in the field.

Swept Frequency Range and Linearity

Sweep Signal Generators present problems peculiar to testing of the swept RF output signal.

For measurement of the linearity of frequency deviation of the swept RF output signal vs the low frequency sweep output voltage, a calibrated oscilloscope is needed. The calibration is obtained by connecting an adjustable calibrated voltage source to the oscilloscope X-axis terminals. The X-axis deflection to right and left of the scale window center, using the center as a reference point, is plotted vs the input voltage. For convenience marks, including test limits, may be placed on the scale window for equal voltage increments.

Such a calibrated oscilloscope is used in measuring the linearity of the swept RF output of our Type 240-A Sweep Signal Generator[5]. The type 240-A includes a built-in crystal-controlled frequency identification system. A harmonics generator generates a fence of crystal-controlled reference frequencies with a choice of spacings: 2.5 mc, 0.5 mc, or 0.1 mc. A sample of the swept RF signal beats, in a mixer-amplifier system, with the reference fence, giving "birdie" (zero beat type) markers at the same spacing as the reference fence.

The X-axis terminals of the calibrated oscilloscope are connected to the low frequency sweep output voltage terminals of the 240-A. The Y-axis terminals are connected to the output of the 240-A mixer-amplifier system (COMPOSITE SIGNAL OUT connector). The locations of the birdie markers now displayed on the screen are compared to the calibration marks or plot, thereby measuring the linearity of the swept RF (i.e. the frequency deviation) vs the Signal Generator low frequency sweep output voltage.

The RF sweep width of the 240-A Signal Generator may be measured in the set-up just described, by using the reference fence of "birdie" markers of known frequency spacing.

Amplitude Flatness of Swept RF Output Signal

The amplitude flatness of the swept RF output voltage is measured by interposing a simple broad-band detector between the signal generator RF output connector and the Y-axis input terminals of a DC amplifier type oscilloscope. When the oscilloscope X-axis terminals are connected to the signal generator low frequency sweep output voltage terminals the horizontal trace line now corresponding to the detector output voltage is an indicator of the amplitude flatness of the swept RF voltage. If the rectangular pattern is now expanded vertically by increasing the oscilloscope Y-gain by a known amount, variations in the height of the detector output trace from the horizontal zero reference (blanked RF output) trace will be indicated more accurately. Correction for detector non-linearity may be necessary.

Summary

A signal generator, in order that it may be a useful testing tool, must be calibrated initially to an adequate degree of accuracy, and must maintain adequate accuracy and necessary operating characteristics over a long period of time. These important attributes require sound design, careful fabrication, and careful and thorough testing procedures. It is to this purpose that our testing and calibrating efforts are directed.

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A Standard for Q and L

Figure 2. Varnishing and Baking the Coil Form.

The widespread use of the Q meter in industry and research has brought about the need for some simple and practical means of checking the overall performance and accuracy of this instrument. Up to the present, the 103-A coil series has served this purpose in default of anything better.

Today, after considerable research, BRC engineers have evolved 5 new coils, the Q-Standards Type 518-A. These coils are well shielded and have been designed to maintain highly stable inductance and Q characteristics. Developed for use with Q Meter Type 260-A, the Q-Standards are useful not only as a check on the overall performance of this instrument but can be used as reference inductors for many impedance measurements. Similar in construction and performance to the 513-A, these Standards, in conjunction with the 513-A, provide frequency coverage from 50 kc to 50 mc- the entire range of Q-Meter Type 260-A.

Construction Details

In external appearance the coils are very similar to the inductors Type 103-A which are available for use as accessory coils in a variety of Q meter measurements. This resemblance is only superficial, however, since highly specialized design and manufacturing techniques are required to provide the high degree of electrical stability demanded of such units.

The Q Standard unit consists of a

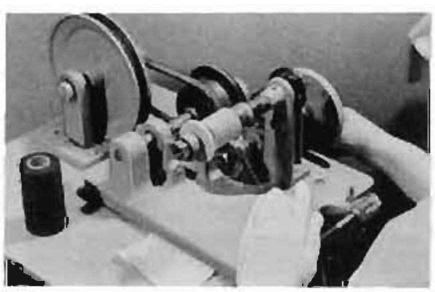

Figure 1. Winding of Coil on Steatite Form.

specially developed high-Q coil wound on a stable low-loss Steatite form and protected by a covering of low loss material (Figure 1). After winding, the coil is heated to remove any moisture present, coated with silicone varnish and baked (Figure 2). Desirable Q versus frequency characteristics are obtained where necessary by a carbon film resistor shunted across the coil. The coil form is mounted on a copper base fitted to a cylindrical shield can. The unit is hermetically sealed, evacuated and filled with an inert gas to a pressure of 1 p.s.i. above atmospheric pressure, (Figure 3). Coil leads are brought through the base to banana plug connectors which can be replaced if necessary without breaking the seal. The high potential connector is slightly longer than the low potential connector and is insulated from the base by a ceramic seal which serves as a stop to insure accurately reproducible positioning of the unit with respect to the Q-Meter cabinet.

Electrical Characteristics

The principal electrical characteristics of each Q-Standard are measured at the factory and stamped on the nameplate of the unit. Three sets of data are thus provided to cover the low, middle and upper frequency ranges of the 260-A Q-Meter. Each set of data contains the following information: a check frequency and its associated values of resonating capacitance and indicated Q.

Indicated Q is defined as the circuit Q of the inductor and Q meter combination as read on an average Q meter. Any instrument whose readings vary from the nameplate values of indicated Q by more than ±8% between 50 kc and 30 mc, increasing to ±13% at 50 mc, is not operating in accordance with the original specifications.

Resonating capacitance is defined as the reading on the internal resonating capacitor dials of an "average" Q meter at resonance for stipulated frequencies which have been checked against a crystal calibrator. The 3 values of resonating capacitance marked on the nameplate are accurate to ±½% or ±½$\mu\mu f$ whichever is greater. (Figure 4).

Applications

Accuracy of Indicated Q: The indicated Q values marked on the Q-Standard provide a convenient check on the accuracy of the Q-indicating meter readings. To make such a check proceed as follows:

Allow the Q-meter to warm up for one hour. Then plug the Q-Standard into the coil posts, making sure that the knurled binding post nuts are tightened and the Q-Standard properly seated. Now zero the meters and adjust the *internal resonating capacitor* to resonance at the three resonating frequen-

Figure 3. Sealing Operation of Q Standard..

cies marked on the nameplate. Adjust the *XQ* controls to obtain a reading of *X1* on the *Multiply-Q-By* meter at each frequency setting of the 260-A. The reading of the *Q-Indicating* meter at each frequency should correspond to the indicated Q value marked on the Q-Standard within the tolerance mentioned above. For best results, the frequency of the Q meter should be set using a crystal frequency calibrator.

Accuracy of Internal Resonating Capacitor Calibration: The coil may be used to check the accuracy of internal resonating capacitor dial readings. This is done in the following manner:

Allow the Q-meter to warm up for one hour before mounting the Q-Standard on the coil posts. Adjust for a *Multiply-Q-By* reading of *X1* while adjusting the oscillator to the desired frequency (which has previously been checked against a crystal calibrator). Now tune the circuit to resonance by

Figure 4. Checking Calibration of Q and C.

means of the *internal resonating capacitor* dials. Reading obtained should be the same value as indicated on the nameplate within the tolerances of the instrument and the coil.

Use as a Work Coil: The Q-Standard is useful as an extremely stable work coil for accurate measurements by the parallel method with Q-Meters Type 260-A and 160-A. A knowledge of the coil's inductance and distributed capacitance will be helpful in this application.*

Distributed Capacitance (C_d): The value of distributed capacitance (the effective distributed capacitance of the coil assembly as measured in position on the Q-meter) for the Q-Standard can be determined by the following formula. Any two of the three sets of data given on the nameplate of the coil may be used.

$$C_d = \frac{C_2 - n^2 C1}{n^2 - 1}$$

Where

C_1 = internal resonating capacitance reading at the first frequency (f_1).

C_2 = internal resonating capacitance reading at the second frequency (f_2), where the first frequency is an integral multiple of second.

n = ration of f_1 to f_2

*For a full discussion of this subject see Boonton Radio Corporation's instruction manuals on Radio Frequency Measurements, the Type 260-A Q-Meter or Type 160-A Q-Meter.

True Inductance (L): The true inductance of the coil may be determined by

$$L = \frac{1}{\omega^2 (C_r + C_d)}$$

and *Effective Inductance (L_e):*

$$L_e = \frac{L}{1 - \omega^2 L C_d}$$

The resonating capacitance for any point may now be determined by—

$$C_r = \frac{1}{\omega^2 L} - C_d$$

Use With Q-Meter Type 160-A

The low frequency and the high frequency performance of the Type 160-A Q-Meter (designed in 1938) is not as good as the current Type 260-A Q-Meter to which the Q data on the coil directly applies. The following table shows the approximate multiplying factor which must be applied to the Q data stamped on the nameplate of the 518-A, making it generally applicable to an average Type 160-A Q-Meter.

CORRECTION TABLE

Coil No.	*Frequency Range*
518-A1	15 mc - 45 mc
518-A2	5 mc - 15 mc
518-A3	1.5 mc - 4.5 mc
518-A4	150 kc - 450 kc
518-A5	50 kc - 150 kc

Approximate Resonating Capacitance

400 μμf	*100 μμf*	*40 μμf*
0.86	0.96	1.15
0.95	0.98	1.00
1.00	1.00	1.00
0.99	0.97	0.95
0.97	0.90	0.83

For additional information on these correction factors see "Q Meter Comparison", BRC Notebook, No. 2, Summer, 1954.

Service Note

Low-frequency adjustment of R X Meter, Type 250-A

Occasionally, a situation develops where it is impossible to achieve balance at the lower frequencies when using the R X Meter, Type 250-A. The following discusses the method by which this condition may be corrected.

All adjustments of the 250-A R X Meter bridge circuit are made at B.R.C. before the R_p drum dial is calibrated. The actual calibration itself, entails the painstaking hand marking of every point along the effective 28 inches of the scale.

However, since the nature of the instrument is such that at frequencies above 100 MCS, the zero balance of the bridge is of necessity, very sensitive to minute variations in internal circuit capacitance; a screw driver trimmer adjustment, described on page 7 of the Summer 1954, Number 2 issue of the Notebook, is provided to compensate for any slight changes that may occur after the dial is engraved.

This is the only adjustment of the bridge circuit that should be made in the field.

About an inch and a half to the right and slightly to the rear of this trimmer is another screw driver adjustment, designated, C 109 on the schematic, which can be seen when the top cover of the instrument is removed. (Figure 1). The setting of this trimmer is extremely critical and upon it depends, to a great extent, both the accuracy of the R_p calibration and the proper balancing of the bridge at all frequencies. Under no circumstances, therefore, should this trimmer be disturbed.

In cases where users of the 250 have, nevertheless, turned this trimmer and ended up with a bridge which cannot be made to balance, especially at low frequencies, the following procedure will in most cases restore the instrument to normal operation.

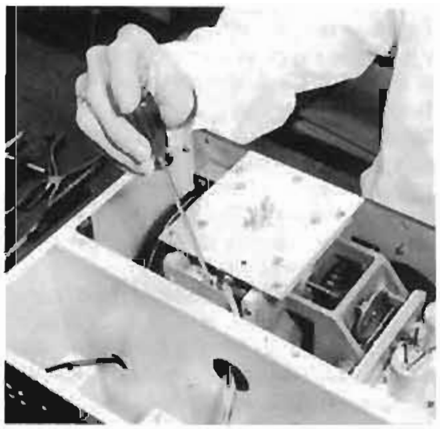

Figure 1. Location of Trimmer, C109 for Adjustment Purposes.

1. Set the frequency to .5 mc and the R_p dial to ∞.
2. Adjust the "detector tuning" knob in the normal manner for maximum deflection of the meter.
3. Position both the "R Coarse" and "R Fine" balance knobs so that the set screws are at "10 o'clock".
4. Turn the "C" balance knob fully clockwise.
5. Using the C_p crank knob and the trimmer C 109 mentioned above, balance the bridge. This should occur at zero or within 10 mmfds in the silver portion of the C_p dial. For the final balance, a slight adjustment of the "R Fine" balance might be necessary.
6. Using precision film resistors, check the accuracy of the R_p dial at 10 k and at 200 ohms (Frequency still at 0.5 mc). If the 200 ohm resistor reads low, turn the "R Coarse" knob a little more counter-clockwise and repeat steps 1, 2, 4, 5 and 6. If the 200 ohm resistor reads high, turn "R Coarse" knob clockwise and repeat as before.
7. Replace cover and check for balance.
8. Check for balance at 200 and 250 mcs as covered in the manual.

The Price of an Instrument

FRANK G. MARBLE, *Vice President — Sales*

A general increase in our prices was put into effect on December 1, 1956; the first such adjustment in many years. This change was necessary because increased material costs and operating expenses could no longer be completely compensated by engineering and production ingenuity and increased volume. These factors had "held the line" for a remarkably long time; one instrument recently replaced by a new model, sold at the same price for almost ten years.

The present necessity for a price change and the long period of stable prices in face of continuing increases in prices in the general economy, gave rise to some consideration of the price of an instrument. Just how does a company establish such a price? Having established the price, how can it maintain that price over sustained periods and still pay its expenses and make a profit?

To produce an instrument, once a market is determined to exist and specifications have been established; engineering, development, and design must be undertaken to prepare models and information from which manufacture can be undertaken. The market determination and specification establishment are overhead expenses applicable to any instrument development while the costs of engineering the instrument are for the most part specific and must be paid out of sales of the instrument. To produce the instrument, materials must be purchased, labor expended and the instrument must be advertised and called to potential customers attention.

The price must be set prior to sale and thus calls for estimates and judgment, since actual specific experience is not available. The costs of engineering can be ascertained since final pricing usually takes place near the completion of the engineering cycle. The length of time over which the engineering costs will be spread and the expected number of sales in that period must be estimated. The engineering costs can then be divided by this number of units which result now forms part of the instrument's cost. Material estimates and labor time estimates, yield additional costs. Finally, overhead, representing the costs of services; purchasing, selling, supervision, building, tool costs, etc. and profit are added and the price becomes available.

The labor time estimates must be projected as an average over a period of time. The early instruments will always take much more time than the average for the life of the instrument. These early instruments will cost more than their selling price. Estimating this "learning curve" is important. As a minimum, the instrument must sell the number of units estimated within the time in which the engineering costs are to be repaid. If it sells additional units over a longer time, capital for additional engineering and a lowered price for subsequent instruments results. General attention to reasonable overhead costs must be continuously maintained.

If all this work is properly done, a reasonably priced instrument will result which attracts customers over a considerable period of time. A properly engineered and priced instrument will return its purchase price many times over to its user. We believe, that electronic instruments are one of the biggest bargains available in today's market.

EDITOR'S NOTE

It is widely feared our nation is falling behind in the technological race. The world's population is increasing at an accelerating tempo but the need for technically trained personnel in this age of automation increases at an even faster rate. While waiting for the educational system to re-adjust itself in the light of present day realities, most employers are searching for ways and means of improving the efficiency of our present crop of engineers. Magazines devoted to this problem have sprung up. Personality and environmental factors are coming in for more attention. Psychological terms — the "problem child", non-conformist, the unconscious, the subconscious, the semi-conscious and the no-account are now found liberally sprinkled throughout our technical literature.

No longer is Management concerned solely with the technical merit of the engineering applicant. In the larger corporations where Research and Development receives a significant share of the budget dollar, the biological or "animate sciences" — psychology, psychiatry, even phrenology are increasingly coming into use in the appraisal of applicants. In some corporations psychological testing is already routine. Sweating palms, twitching of the eyebrows (perhaps due to a sadistic mosquito that prefers applicants over interviewers) may be cause for rejection even though the applicant has his shoes shined, his Sunday suit pressed and an impressive background. A handshake that is too firm may indicate a potential "Stakhanovite," one who is geared up at a high rpm and may therefore cause a serious morale problem among those who work at a more normal rate. On the other hand, an engineer with a weak handshake might be suspect as a latent bohemian, a nonconformist indifferent to the need for "getting along".

If he is unfortunate enough to pass these batteries of tests and possesses the proper phrenological characteristics, he is hired. These phrenological requirements, incidentally, vary from company to company; some companies demand two bumps on the forehead and one aft, symmetrically spaced; others are more modest, requiring only one bump fore and aft. This matter of phrenological standards will be aired by an industry-wide Committee that is being established to standardize criteria on employment as a guide for interviewers. The committee will be known as the Committee on Interview Criteria or the CIC committee (for phonic convenience it is rhymed with sick).

There follows assignment to an R/D team, which must be productive of new ideas to justify itself. Now ideas, creative ones, are difficult things to come by since the technically trained mind, it is believed, is something inhibited from seeing the true relation between things due to its peculiarly one-sided education and isolation from the administrative end of things where company policy is made. Nevertheless a brave stab is being made by utilizing the latest advertising technique of generating ideas for "selling" a product. This is called "brainstorming" or cranial ventilation and is similar to the free association principle used in psychoanalysis to liberate facts from the "unconscious". It goes something like this—

The engineering team is placed around a large circular table in a room devoid of distracting decor such as pictures (especially nostalgic landscapes), curtains, etc. Lights are turned low, and binaural "think" music pours softly out of speakers hidden in the ceiling. A contemplative mood established, the engineers are now to toss any ideas that pop into their heads into a pot. This goulash is gingerly examined from time to time by the Idea Co-ordinator who hopes to find therein a few pearls of thought. But alas, it is being sadly admitted that no sooner do the lights dim than the "think" music is transformed through some perversity of the engineering brain into "dream" music that produces only yawns and snores.

The inevitable reaction is beginning to set in. There is now a real danger the policy of the "hard school" will take effect. This consists of placing the engineer's head in a wine press adapted for the purpose and squeezing. Ideas, if any, are caught in a large beaker of formaldehyde, the heaviest ideas settling to the bottom, the others remaining in suspension.

Any evidence obtained from "modern" methods to the contrary, engineering is still a profession for enthusiastic, hard-working-people well grounded in fundamentals. The horizons have broadened but the basic requirements are not different.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Q Meter Techniques

NORMAN L. RIEMENSCHNEIDER, *Sales Engineer*

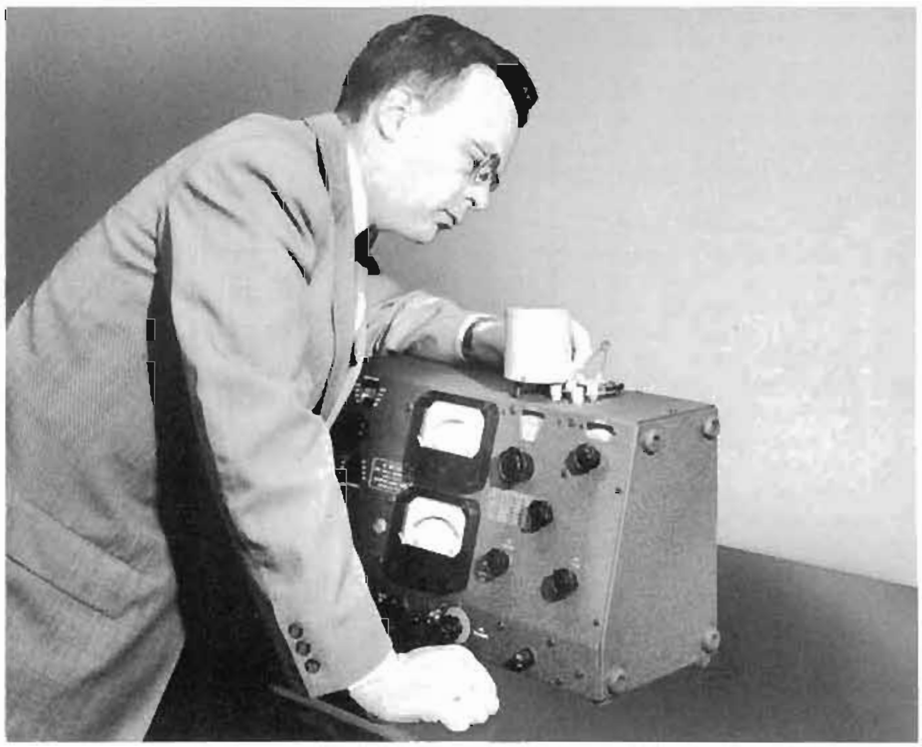

Figure 1. The author connects a fixed capacitor to the terminals of a series jig prior to measuring the capacitance on a Type 260-A Q Meter.

The Q Meter has frequently been described as one of the most flexible instruments available with applications limited largely by the ingenuity of the person using it. It is our desire here to delineate some of those techniques, not normally encountered in everyday work, in the hope that wider dissemination of information gathered through many channels, will prove of some value.

In order to approach our specific problems in a general way, it might be well to review some basic facts relative to the operation of the Q Meter.

The Q Meter is always operated with a coil connected to its coil terminals. If we are interested in measuring the Q of a coil, this coil will be connected to these terminals and it will be measured in one operation. If we are interested in making other measurements; (i.e., the Q of a capacitor, the impedance of a circuit, the parameters of a tuned circuit; etc.), we still need a coil, even though we are interested in that particular coil only as a reference. This so-called "work coil" would probably be a shielded unit to prevent stray coupling, hand-capacitance effects, etc.; and might be selected for its inductance, Q; etc., as needed for the particular application involved.

In making measurements (other than the Q of a coil) of circuit parameters there will be two steps involved. The first will be with the work coil mounted on the Q meter, where the resonating capacitance (C_1), circuit Q (Q_1), and frequency will be recorded. The second will be with the unknown connected in addition to the work coil and once again the above reading will be noted, this time as C_2 and Q_2.

From this data the desired parameters can be determined using the appropriate formula selected from those shown in figure 2. High impedance circuits are measured by connecting them in *parallel* with the *Q Capacitor;* i.e., across the "Capacitor" terminals, and using the formulas shown under the heading "Parallel Connection to Q Circuit". If the unknown consists of more than one parameter, it should be noted that the *equivalent parallel* parameters are obtained in this manner. Low impedance circuits are measured by connecting them in *series* with the "Low" side of the coil. In like manner the "Series Connection to Q Circuit" formulas are used to yield the *equivalent series parameters* of the circuit involved.

With the above in mind, it might be well to resolve some specific problems.

1. Measurement of Coils.

a. *Coil inductance too great to resonate with minimum capacitance of Q Meter at desired frequency.* Since this can be considered a high impedance measurement, the unknown coil can be connected to the capacitor terminals and readings of C_1, C_2, Q_1, and Q_2 made in two steps. There will actually be two coils involved in this measurement. C_1 and Q_1 will be the values read with the

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work coil only mounted on the coil terminals. C_2 Q_2 are made after the *unknown* coil has been added across the capacitor terminals and the measurement is made with *both* coils connected. Formula 3b will yield the inductive reactance and 2b will give the equivalent parallel resistance. Combining these two parameters, formula 1b will give the Q directly.

b. *Maximum resonating capacitance on Q Meter insufficient at chosen frequency.* If it is desired to measure a coil having so low an inductance that it cannot be resonated with the capacitance available on the Q Meter, additional capacitance can be added across the Q Meter capacitor terminals and the coil measured in a normal manner. The additional capacitance must be known, if it is desired to compute the reactance of the coil.

c. *Measurement of extremely high Q coils.* If the Q of a coil is beyond the range of the Q meter with its highest XQ multiplier, it can be measured using the same procedure described under (a) above.

d. *Measurements at frequencies below 50 kc.* Measurements can be made on both the 160-A and 260-A Q Meters at frequencies down to 1000 cps by disconnecting the Q Meter oscillator and supplying the instrument signal from an external source through a Coupling Unit Type 564-A. A jack has been provided in the top of the 160-A Q Meter for this purpose. The 260-A has also been provided with a removable panel in the rear of the instrument where disconnection of the 50-ohm cable from the thermocouple block by means of a BNC connector allows easy adaption to low frequency measurements. Any audio oscillator having good waveshape and capable of developing a variable voltage up to 22 volts across 500 ohms can be used.

For coil measurements at these lower frequencies, it will probably be found necessary to provide additional capacitance across the capacitor terminals to resonate the inductors frequently encountered in this frequency range. In using an external oscillator, care must be exercised that the output attenuator is turned all the way down before connecting to the Q Meter in order to pre-

SERIES CONNECTION TO Q CIRCUIT

$$Q_x = \frac{(C_1-C_2)\,Q_1 Q_2}{C_1 Q_1 - C_2 Q_2} \qquad (1a)$$

$$R_s = \frac{1.59\times 10^8 \left(\frac{C_1}{C_2}Q_1 - Q_2\right)}{f\,C_1 Q_1 Q_2} \qquad (2a)$$

$$X_s = \frac{1.59\times 10^8\,(C_1-C_2)}{f\,C_1 C_2} \qquad (3a)$$

$$L_s = \frac{2.53\times 10^{10}\,(C_1-C_2)}{f^2 C_1 C_2} \qquad (4a)$$

$$C_s = \frac{C_1 C_2}{(C_2-C_1)} \qquad (5a)$$

When C_1 is:
Greater than C_2, X_s is inductive (+).
Less than C_2, X_s is capacitive (−).

PARALLEL CONNECTION TO Q CIRCUIT

$$Q_x = \frac{(C_2-C_1)\,Q_1 Q_2}{C_1(Q_1-Q_2)} \qquad (1b)$$

$$R_p = \frac{1.59\times 10^8\,Q_1 Q_2}{f\,C_1(Q_1-Q_2)} \qquad (2b)$$

$$X_p = \frac{1.59\times 10^8}{f\,(C_2-C_1)} \qquad (3b)$$

$$L_p = \frac{2.53\times 10^{10}}{f^2\,(C_2-C_1)} \qquad (4b)$$

$$C_p = C_1 - C_2 \qquad (5b)$$

When C_1 is:
Greater than C_2, X_p is capacitive (−).
Less than C_2, X_p is inductive (+).

In the formulas for Q, the quantities (C_1-C_2) and (C_2-C_1) are always considered positive.

The following symbols refer to values of the unknown impedance, Z: The units used are:

Q_x = Q of the unknown impedance	R_p = Effective parallel resistance.	R = Resistance in ohms.
R_s = Effective series resistance.	X_p = Effective parallel reactance.	X = Reactance in ohms.
X_s = Effective series reactance.	L_p = Effective parallel inductance.	L = Inductance in microhenries.
L_s = Effective series inductance.	C_p = Effective parallel capacitance.	C = Capacitance in micro-microfarads.
C_s = Effective series capacitance.		f = Frequency in kilocycles per second.

Figure 2. General formulas for series and parallel connections.

SERIES CONNECTION TO Q-METER

$$R_p = \frac{1.59\times 10^8\,(C_1-C_2)^2\,Q_1 Q_2}{f\,C_1 C_2 (C_1 Q_1 - C_2 Q_2)} \qquad (6a)$$

$$X_s = X_p = \frac{1.59\times 10^8\,(C_1-C_2)}{f\,C_1 C_2} \qquad (7a)$$

$$L_s = L_p = \frac{2.53\times 10^{10}\,(C_1-C_2)}{f^2\,C_1 C_2} \qquad (8a)$$

$$C_s = C_p = \frac{C_1 C_2}{(C_2-C_1)} \qquad (9a)$$

PARALLEL CONNECTION TO Q-METER

$$R_s = \frac{1.59\times 10^8\,C_1\,(Q_1-Q_2)}{f\,(C_2-C_1)^2\,Q_1 Q_2} \qquad (6b)$$

$$X_s = X_p = \frac{1.59\times 10^8}{f\,(C_2-C_1)} \qquad (7b)$$

$$L_s = L_p = \frac{2.53\times 10^{10}}{f^2\,(C_2-C_1)} \qquad (8b)$$

$$C_s = C_p = C_1 - C_2 \qquad (9b)$$

In the above formulas the same units, symbols and conditions stated in figure 2 apply except that these formulas are accurate only for impedances having a Q greater than 10. The formulas in figure 2 are accurate for any impedance.

Figure 3. Series and parallel connection formulas for impedances having a Q greater than 10.

$$Q_x = \frac{X_s}{R_s} = \frac{6.28\times 10^{-3}\,f\,L_s}{R_s} = \frac{1.59\times 10^8}{f\,R_s\,C_s} = \frac{R_p}{X_p} = \frac{159\,R_p}{f\,L_p} = 6.28\times 10^{-9}\,f\,R_p\,C_p \qquad (10)$$

$$R_s = \frac{R_p}{1+Q_x^2} \qquad (11a) \qquad\qquad R_p = R_s\,(1+Q_x^2) \qquad (11b)$$

$$X_s = X_p\frac{Q_x^2}{1+Q_x^2} \qquad (12a) \qquad\qquad X_p = X_s\frac{1+Q_x^2}{Q_x^2} \qquad (12b)$$

$$L_s = L_p\frac{Q_x^2}{1+Q_x^2} \qquad (13a) \qquad\qquad L_p = L_s\frac{1+Q_x^2}{Q_x^2} \qquad (13b)$$

$$C_s = C_p\frac{1+Q_x^2}{Q_x^2} \qquad (14a) \qquad\qquad C_p = C_s\frac{Q_x^2}{1+Q_x^2} \qquad (14b)$$

Figure 4. Series to parallel transfer formulas.

clude possible thermocouple damage. Reasonably good waveshape is required of the supply, since the thermocouple responds to the rms voltage and the Q Voltmeter is a peak-reading device. The presence of harmonics will therefore affect the Q reading.

e. *Dielectric measurements at low frequencies.* We have had many inquiries from customers relative to the use of a Q Meter for low frequency dielectric measurements. On information received from the field, a typical installation for 1 kc operation would include, in addition to a Q Meter and a 564-A Coupling Unit, a Hewlett-Packard audio oscillator Type 200 AB (20 cycles to 40 kc), a UTC Type MQ B-11 inductor and a good variable or decade capacitor having up to approximately 1500 MMFD, or equivalent equipment. Operation at other frequencies would have other inductance or capacitance requirements.

f. *Inductance measurements.* It might be worth noting that in using the L/C dial for direct inductance measurement, the *effective* inductance is given. If it is desired to read the true inductance and the distributed capacitance, C_0, is known, it can be done by *increasing* the capacitance dial setting, after resonance has been established, by an amount equal to the C_0 and reading off the corresponding value of inductance as the true inductance.

2. Measurements of Capacitors.

This is a fairly common operation and there is no point in elaborating on it. Briefly, the Q of a capacitor is evaluated by first selecting a suitable coil and measuring its Q and resonating capacitance on the Q Meter. After noting these as Q_1 C_1, the unknown capacitor is connected across the capacitor terminals and the circuit is once again resonated using the internal Q capacitor. Values of Q_2 C_2 are recorded, and from Q_1, Q_2, C_1, and C_2, the Q of the unknown can be computed using the appropriate equation in figure 2. It might be noted that the accuracy of measuring capacitance can be improved by using an external variable precision capacitor. Since this measurement is essentially one of substitution, the range of measurable capacitance can also be extended by using a coil with less inductance, thereby requiring the addition of more external capacitance.

Large capacitors may be measured by connecting them in series with the low side of the coil. For this measurement, however, a coil is used whose inductance will resonate to the desired test frequency with the Q capacitor set to the high capacitance end, so that the addition of the test capacitor will cause a measurable change. It is also necessary to shunt the capacitor under test with a 3 to 10 megohm resistor to provide a dc return path for the voltmeter.

With large capacitors it is particularly important to minimize lead length. Shorting the capacitor out, rather than removing it entirely, during the first measurement is desirable.

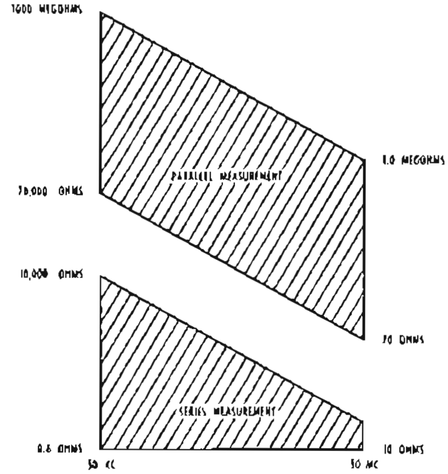


Figure 5. Ranges of measurable resistance on Q Meter, Type 260-A.

Figure 6. Grid electrode holder for measuring mica in a plane parallel to the cleavage plane.
(Courtesy of Bell Telephone Laboratories and American Society for Testing Materials. From ASTM Designation: D748-54T. Reprinted with permission.)

3. Measurements of Resistors at Radio Frequencies.

Resistors are handled either in series with the low end of the coil or in parallel with the capacitor terminals, depending upon whether the resistance value is low or high respectively. There are values of resistance for each frequency, however, that cannot be measured. Some idea of this can be had from figure 5. The use of the low Q scale tends to shrink this area which cannot be measured and there have been articles[1] in the literature describing other techniques that can be employed to surmount this problem.

4. Measurement of Mica.

Q Meters have found wide application in the sorting of mica based on its dielectric properties. The measurement can be accomplished in the perpendicular plane, using a clamp type holder with suitable test electrodes, or in a plane parallel to the cleavage plane using a holder with grid electrodes. The latter is illustrated in figure 6. The actual test methods employed have been adequately covered elsewhere[2].

5. Application of Biasing Potentials.

It is sometimes desirable to investigate circuits with biasing potentials or currents applied during the measurement. Figure 7 illustrates how this can be accomplished using a blocking capacitor and a high impedance power source. Capacitor C must be large enough to present negligible reactance at the operating frequency, and R must be high enough to prevent loading of the circuit by the power supply. The applied voltage will be determined by the resistance, R, and the biasing current. Polarity of the battery and magnitude of the biasing current are inconsequential to the operation of the Q Meter.

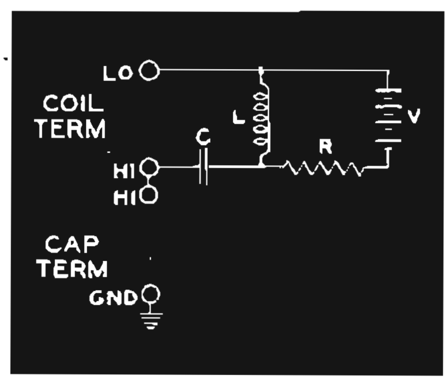


Figure 7. Circuit for applying Q Meter bias.

OWNERS OF SIGNAL GENERATORS, TYPE 202-F FM-AM AND UNIVERTERS, TYPE 207-F PLEASE NOTE

Final Operating Instruction Manuals for these Instruments are now available. If you have not received your copy, please let us know and it will be forwarded immediately.

Figure 8. Q Meter, Type 260-A connected to RF Voltage Standard, Type 245-B.

6. Use of 260-A as a Signal Generator.

The 260-A Q Meter can be improvised as a CW signal source up to approximately 20 mcs, by removing the small rear cover plate, disconnecting the coaxial cable going into the thermocouple block, and connecting it to the input of the 245-B RF Voltage Standard through a 20 db pad. The output of the 245-B then is ½, 1.0, or 2 microvolts depending upon which level is selected on the meter. The X Q control varies the oscillator output for the signal level desired. See figure 8.

7. Use of Q Meter as a Wave Meter.

If the oscillator range switch is set between ranges to turn the oscillator off, a coil can then be connected across the coil terminals and coupled to an active circuit whose frequency is not known. After the Q capacitor has been tuned for a maximum deflection and the active circuit removed, the Q Meter oscillator can then be turned on and its frequency varied for another indication of resonance. The frequency thus indicated, is the frequency of the unknown signal under test.

8. Miscellaneous.

a. *Use of 260-A with frequency counter.* Where greater than 1% frequency accuracy is desired, the Q Meter oscillator can be monitored with a counter. This is accomplished by removing the rear panel, disconnecting the transmission line going to the thermocouple block, and inserting a "tee" fitting to allow the parallel connection of the counter input. Operating the Q Meter with its multiplier set at "X 1" will afford approximately 0.5 volts to a high input impedance counter.

b. *Use of Scale Magnifier.* Where it is desired to read the main capacitor dial more closely than ordinarily allowed, a hemicylindrical magnifier made from plastic, as shown in figure 9, can be used. Constructional details are shown in figure 10.

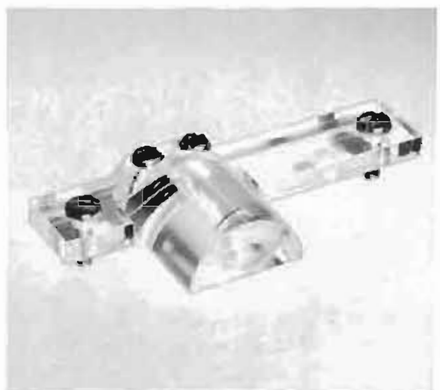
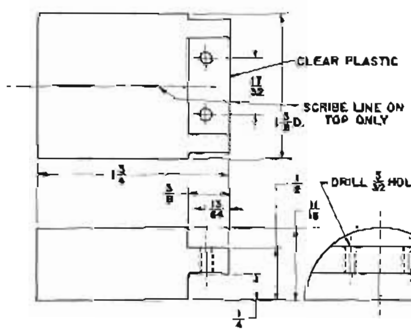


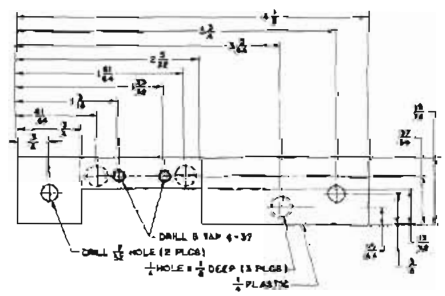
Figure 9. Main capacitor dial magnifier.



MAGNIFIER

ASSEMBLY INSTRUCTIONS

1. Remove (2) #10-32 x 3/8 BH mach. screws on 260-A panel which line up with 7/32 dia. holes.
2. Attach strip with (2) #10-32 x 5/8 BH mach. screws thru 7/32 dia. holes, using flat washers & lock washers. (1/4 dia. spot-face at 3-5/64 will clear fiducial screw on recent 260-A's. Other (2) 1/4 dia. spot-faces will clear fiducial screws on early 260-A's).
3. Mount magnifier with (2) #6-32 x 1/2 BH mach. screws using both flat & lockwashers.



ATTACHMENT STRIP

Figure 10. Construction and assembly details of the main capacitor dial magnifier.

c. *Use of "delta Q" scale to refine resonance.* The "delta Q" scale can be used to obtain a very precise resonance by taking advantage of its greater sensitivity. After carefully resonating the Q Circuit in a normal manner, set the "delta Q" dial to the Q indicated on the Q Meter, and depress the key. By adjusting the "delta Q" potentiometers to center the needle on the red scale, very fine adjustments can be made to the internal resonating capacitor for resonance.

d. *Use of "delta Q" scale as a "go—no-go" test.* Using the technique shown above under (c), once resonance is established, limits within the confines of the red scale can be established. Centering the needle in the red scale for the nominal value, tolerances can be set up for components in terms of deviation from the center point in either direction.

e. *Use of "Lo Q" scale for zeroing voltmeter.* Since both the "Q" scale and "Lo Q" scale have the same zero point, the voltmeter can be adjusted by making the adjustment on the "Q" scale and depressing the "Lo Q" key to make sure there is no change on the needle. When the instrument is properly adjusted for zero, "pumping" the "Lo Q" key should not move the needle at all.

Conclusion

The foregoing discussion is by no means intended to define the limits of the Q Meter, but instead indicate some of its potentialities. As new problems arise, we hope new techniques will be developed. Our field staff stands ready to assist in these problems.

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7. Instruction Manual for the Q Meter Type 260-A, Boonton Radio Corp., Boonton, N. J.

Glide Slope Generator Tone Signal

Phase Relationships of the 90-150 cps Tones

When the weather gets soupy airliners and military aircraft rely on electronic guidance systems for safe landings[1]. One component of this common navigational system is the glide path established by a transmitter located near the end of a runway and the associated receiving and indicating equipment in the aircraft.

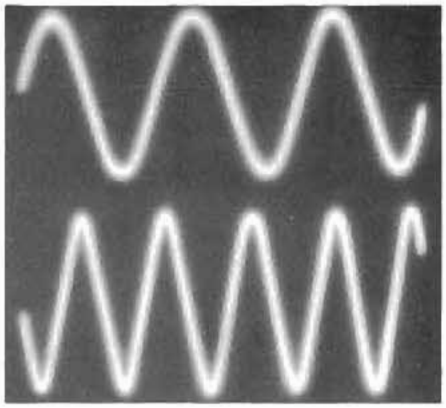
Operation of the system is based upon the ratio of a 90 cps signal to a 150 cps signal appearing at the output of the receiver. The desired course is that inclined plane in which the two modulation tones are of equal intensity and the receiver circuits must be adjusted to produce the corresponding on-course indication.

To insure proper operation of the receivers and indicators it is necessary to check them at frequent intervals with signals corresponding to those received aboard the aircraft in various positions with respect to the desired glide slope. This is usually done with the Glide Slope Signal Generator Type 232-A or its military equivalent, Signal Generator SG-2. This instrument contains a synchronous alternator which supplies 90 cps and 150 cps tones and has provisions for mixing the ratio of the two by predetermined amounts to correspond to different positions of the aircraft with respect to the on-course signal.

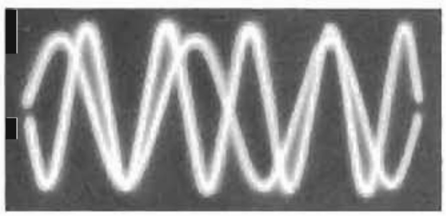
Standard Conditions

In the ground transmitter the relative phase of the 90 cps and 150 cps modulation signals is set so that at no time do peak voltages of both signals occur simultaneously. Otherwise the maximum percentage modulation which could be assigned to either signal would be one-half the maximum total modulation level since each would contribute equally at the in-phase instant as shown in Figure 1.

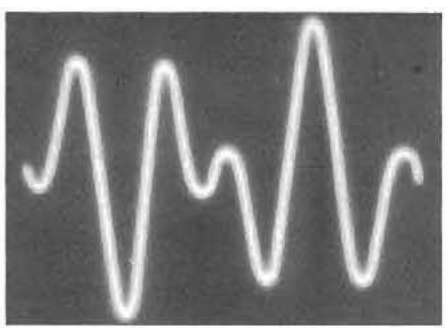
If, however, the relative phase between the alternators is correct, at no time will peaks add in either polarity but will be spaced by a minimum separation of approximately 12° of the 30 cycle repetition rate of the composite pattern. This condition is shown in Figure 2. For a given total peak-to-peak swing of the modulation, it is now possible to deliver a larger percentage modulation of each of the two component signals comprising the composite waveform.



a. Incorrectly-phased 90 cps and 150 cps tones. Leads from 150 cps alternator have been reversed.

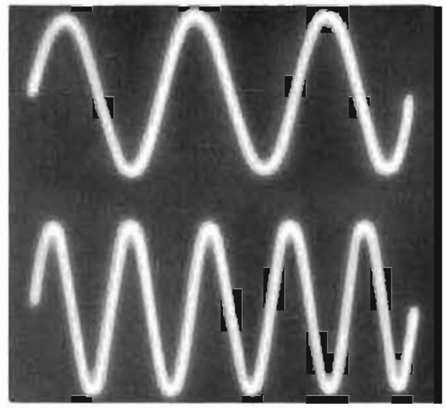


b. Incorrectly-phased tones superimposed.

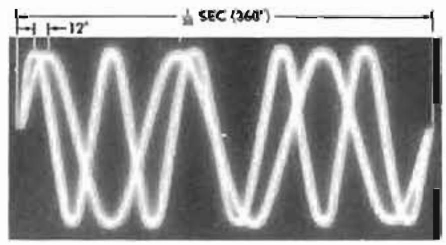


c. Sum of incorrectly-phased tones for "on course" signal as applied to modulator of Glide Slope Signal Generator Type 232-A.

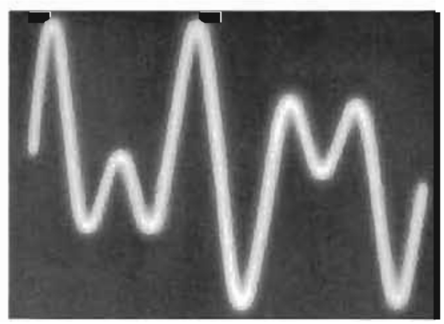
Figure 1. Waveforms for incorrectly phased 90 cps and 150 cps modulation signals.



a. Correctly-phased 90 cps and 150 cps tones displayed separately.



b. Correctly-phased tones superimposed.



c. Sum of correctly-phased tones for "on course" signal as applied to modulator of Glide Slope Signal Generator Type 232-A.

Figure 2. Waveforms for correctly phased 90 cps and 150 cps modulation signals.

The condition pictured in Figure 2 has been selected as standard and is defined on page 41 of reference number 2. This specification defines a certain phase relationship for the signals. There are, however, at least four ways in which the phase relationship can be tested *provided* an initial overall test is made to be certain that the output windings from the 90 cps and 150 cps alternator have been properly interconnected and that the two voltages are approximately equal.

Testing Methods

The definition, which may be used as one basis for a method of measurement, states that the 90 cps and 150 cps signals shall be in phase on the *zero-axis-*

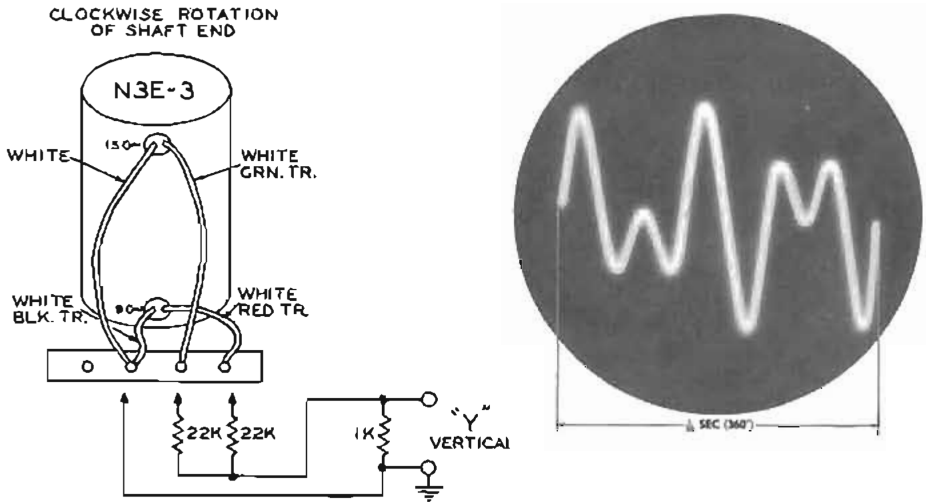


Figure 3. Test connections and correct oscilloscope pattern for the 90-150 cps alternator (Eastern Air Devices Co., Type N3E-3) in the Type 232-A Glide Slope Signal Generator.

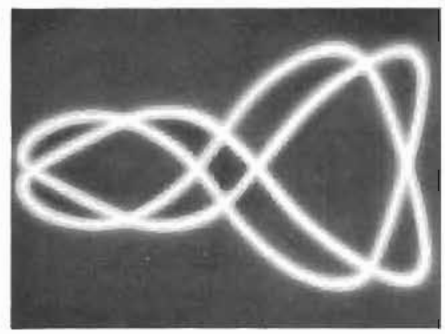


Figure 4. Pattern, obtained from a 30 cps sinusoidal sweep, for testing phase relationships.

crossing on the *positive-going* wave slope. If a dual-channel electronic switch is available for the oscilloscope, it is possible to superimpose the 90 cps and 150 cps signals and obtain the pattern shown in Figure 2b. By the use of suitable techniques and equipment, the region in the vicinity of the zero-axis-crossing can be investigated and the relative phase of the two signals at this point determined.

The second method, which also is based on a definition but requires a somewhat more difficult oscilloscope technique, consists of measuring the separation of the peaks of the two signals to insure that they are not closer than a minimum separation of 12° for either positive or negative pairs as shown in Figure 2b.

In the third method, the two tones are added to give a composite signal as shown in Figure 3. With proper phasing there are pairs of peaks which have the same amplitude. A slight shift in the relative phase will increase the amplitude of one of the peaks while decreasing the amplitude of the other on both the positive and negative pairs. In order to increase the sensitivity of this measurement it is common practice to blow up the image and depress the zero axis by means of the centering control so that a magnified portion of the tips of a pair of peaks appears on the screen. This is a very sensitive test and will yield good accuracy in testing for correct phase angle provided the two signals are of approximately equal magnitude.

A glance at the display will show that it consists of a "W" connected with an "M" in which pairs of legs on both letters are equal. A display in which an "M" is followed by a "W", is equally valid depending on the phase of the synchronizing voltage used for the oscilloscope. Improper connection of the alternator windings gives the waveform of Figure 1c.

A fourth method is based upon forming a Lissajous pattern by using a properly-phased 30 cps sine wave to drive the horizontal oscilloscope amplifier. This has the practical effect of taking the pattern of Figure 3 and wrapping it around a transparent cylinder so that the two ends connect as shown in Figure 4 which strongly resembles a large pair of ice tongs. Since the amplitude of the two peaks depends on the relative phasing of the 90 cps and 150 cps signals, the adjustment for which points of the ice tongs just meet is a very sensitive measure of the phase relationship of the two signals.

Summary

Of the four methods described above two are based on measuring relative phase of two signals in a superimposed display and two are based upon relative amplitude measurements of a composite signal.

When using expanded displays of either of the methods in which amplitude is used as a measure of phase relationship, care must be exercised to insure that the magnitudes of the 90 cps and the 150 cps signals are approximately equal and that the interconnections from the generator are correct as indicated by the "W-M" display shown in Figure 3.

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The Need for Special Instruments

It has always been our policy to develop and design instruments which can be applied under a wide variety of conditions. This flexibility tends to broaden the market thus lowering the price. It also makes it possible for a given laboratory to sustain its work with a smaller number of instruments. For instance the Q Meter Type 260-A covers a frequency range of 50 kc to 50 mc and can be extended down to 1 kc with standard, readily available, external equipment. The range of Q measurements is 10 to 625 and the capacitance range is 30 to 460 micromicro farads. A single-frequency Q meter to measure one level of Q at a given capacitance could be built to very high accuracy but since few would be needed, the price would be high and the owner would soon find many problems not covered by the instrument.

Our policy has also always provided for a standard cabinet and cabinet finish. In addition we have not provided any means for furnishing instruments with performance characteristics or specifications different from those standardly advertised. These policies, also, are based on our strong desire to hold the price to all at a minimum. Our time and money has been concentrated on providing the highest average usefulness to the greatest number of our customers. We have always felt that providing more service to the few would result in less service for the many.

Recent changes in the electronic field have indicated the need for some additions to this policy. Electronic instruments are being used as parts of large assemblies of test equipment instead of as individual laboratory equipments. Since each company's instruments are a different color these assemblies are by no means uniform in appearance. Customers who assemble these instruments quite naturally want them finished in a uniform manner. Commercial equipment in some cases does not exactly fit the performance requirements in other ways. As long as substantial redesign is not required these changes can be made. The importance of the application very frequently justifies the added charges for making minor changes and special arrangements.

To better serve the special requirements discussed above Boonton Radio Corporation has recently set up a Special Devices group. This group will handle finishing and small changes such as relocation of connectors, special cables, and other minor changes to accommodate special requirements. Our internal methods for handling these orders have been simplified so as to give rapid service. Arrangements have been made to assure that none of the charges for this special work appear in the price of our standard instruments. We would be happy to hear your special needs for our equipment with the type of slight modification discussed here.

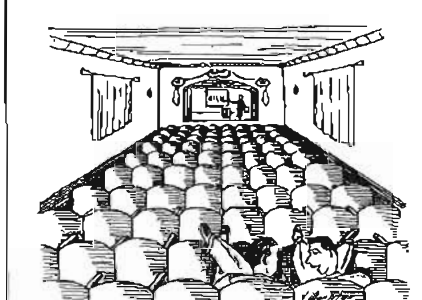
Conventioneering with Cartoons

JAMES E. WACHTER, *Project Engineer*

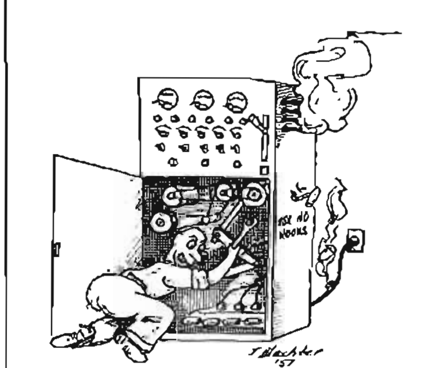
EDITOR'S NOTE

Anyone who has ever attended an IRE show doubtless knows that Jim Wachter, Project Engineer and amateur cartoonist extraordinary at BRC, has covered the field with his "Conventioneering with Cartoons" series illustrated here. Jim saw the convention from "both sides of the fence" so to speak. He served his time in the BRC booth and joined the throngs to view other exhibits. We think the cartoons are an authoritative sampling of what one might expect to encounter at a typical IRE show.

YOU FELLA'S GOT TO LEAVE. THE SHOW'S OVER !!

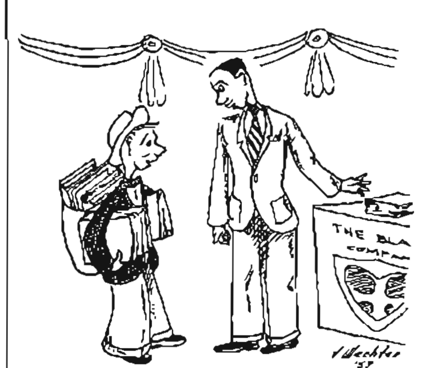


THESE PAPERS ARE THE BEST THING THAT EVER HAPPENED TO MY FEET !!



TEN MINUTES BEFORE THE SHOW OPENS ?

WHAT'S WRONG WITH OUR DISPLAY ?



JUST LITERATURE PLEASE ?!

Q METER CONTEST AWARD

Q of the coil displayed at the IRE show is 336.7. The winning estimate (338) was submitted by Mr. George S. Scholl, Research Engineer with the American Machine and Foundry Co. of Alexandria, Va.

The coil in question was displayed in the BRC booth at the IRE show in New York during March 18 - 21. Anyone visiting the booth was invited to estimate the Q of the coil in competition for a Q Meter which was also on display. Entries were submitted on specially prepared forms. These entries have been tabulated and set up in graph form below to give an indication of the distribution of estimates.

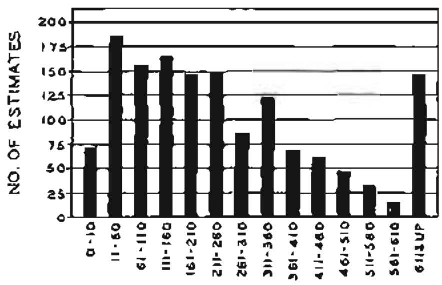
Measurement of the coil was made at BRC by our Quality Control Engineer, on March 25, under the following conditions:

1. The coil was conditioned for 2 hours in the Standard Room with the atmosphere maintained at 73 $\pm 2°$F, relative humidity 50 $\pm 5\%$.
2. Measurement was made on a BRC Type 260-A Q Meter which was checked by Q Standards 513-A and 518-A.
3. The coil was dismounted from the display case and connected to the Q Meter with the coil axis vertical. The winding ends were clamped by the Q Meter coil binding posts; spacing between the winding ends being the same as it was while the coil was on display.
4. The coil measurement frequency of 12.5 mc was checked against a crystal calibrator.

Following measurement, the coil was disconnected from the Q Meter, then reconnected and measured again in the same manner. Readings obtained for both measurements are shown in the table below.

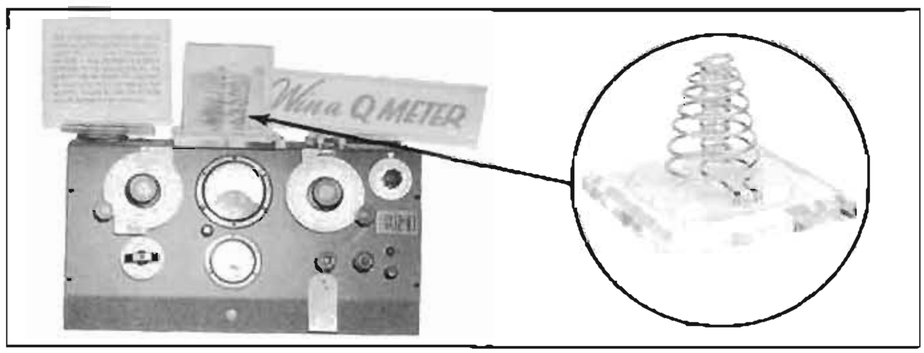
Readings	1st Meas.	2nd Meas.	Average
Frequency (mc)	12.5	12.5	12.5
Q Meter Multiplier	1.4	1.4	1.4
Q Voltmeter	241.0	240.0	240.5
Q Indication	337.4	336.0	336.7
Q Meter Capacitance ($\mu\mu$f)	74.2	74.0	74.1

Other Q estimates worthy of honorable mention were submitted by J. C. Clements, Raytheon Mfg. Co. Ltd. (334) and J. F. Sterner, RCA, Isidore Bady, U.S.A.S.E.L., and S. Krevsky, Evans Sig. Lab., (all with 333).



GRAPH SHOWING DISTRIBUTION OF Q ESTIMATES

Graph showing distribution of estimates in the Q Meter contest.



Shown above are the Type 260-A Q Meter and the controversial coil as they were displayed in the BRC booth at the IRE convention. An enlargement of the coil is shown in the insert.

Printed in U.S.A.

SUMMER, 1957

NUMBER 14

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Calibration of An Instrument for Measuring Low-Level R-F Voltages

CHARLES G. GORSS, *Development Engineer*

The sensitivity of a radio receiver is one of its most important attributes. Often, this sensitivity is in the order of one microvolt or less. Unfortunately, the only devices capable of detecting the presence of these low-level voltages are the receivers under test. Using a receiver of unknown sensitivity to measure these voltages would not be an accurate process. What is needed, is a device which will provide a source of r-f voltages, at microvolt levels, which can be established with a definite and reasonable accuracy without measurement at low levels. This article presents a discussion of some of the techniques employed and problems encountered in designing and evaluating such a device. Basically the device under discussion combines a carefully calibrated r-f voltmeter with a very fine attenuator. See figure 3.

Figure 1. The author uses the 245-B to perform receiver sensitivity measurements.

Voltmeter Design

The voltmeter selected for this device is a reasonably straight forward UHF germanium cartridge diode. The only innovation is that the diode operates with an accurately monitored bias current at zero signal. The bias is such that the input voltage swing is always on the square-law portion of the diode characteristic and never crosses the zero voltage axis. This tends to make the sensitivity relatively independent of temperature and aging effects.

What must be determined however, is the frequency characteristic of this diode voltmeter. At the design stage it was quite obvious that the series resonance of the voltmeter was above 1,000 mc, but the exact frequency was not known. As the series resonance is approached, the diode will increase in sensitivity and the output of the instrument will drop. On the low frequency end, the sensitivity will decrease as the impedance of the by-pass capacitors increase. Since it is most desirable to calibrate this voltmeter at 1,000 cps, where accurate voltages are available, the low frequency characteristics must also be known. A factor to be considered, is that two capacitors are used as a filter over the entire frequency range and that they have a 5-ohm damping resistor between them. Below a certain frequency, this 5-ohm resistor is in series with the r-f circuit of the diode and decreases the sensitivity slightly. For accurate work this must be evaluated.

Voltmeter Performance Checks

The basic problem at the input of the voltmeter system is to accurately measure the voltage applied to the attenuator system over a range of 1 kc to 1,000 mc for a constant indication of the output meter. In order to extend the frequency response downward to 1,000 cycles, an additional 60 μf had to be added to the by-pass circuit. To be sure that this had no effect on the normal calibration, it was necessary to check the meter indication at the lowest operating frequency of 100 kc, with and without the 60 μf capacitor. It was found that this had no effect. It was also found that 60 μf was adequate for 1,000 cps. This was checked by adding more capacitance and noting that no change in sensitivity took place. Operation of the by-passing was observed by evaluation of the effect of the 5-ohm damping resistor at frequencies where the high frequency by-pass has no effect. This was accomplished by a relative check of sensitivity at 1,000 cycles, with and without the damping resistor. Results showed about a 1% change. This 1% increase in sensitivity occurs between 10 and 20 mc as the test frequency is raised. Since this is known, it can be taken into

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account. This condition precludes the possibility of sharp, uncontrolled changes occurring unexpectedly due to series resonance of the two capacitors and their accumulated inductances.

DC to Audio

The most accurate place to begin determination of standard voltage is at a Weston standard cell. In this case, dc was passed through a stable glass film type resistor of 50-ohms dc resistance. This current was monitored with a Weston thermo milliammeter whose calibration was known up to 2 mc. An L and N laboratory potentiometer was used to compare the standard cell voltage with the voltage developed across the resistor. Accurate readings of the thermo milliammeter were taken at the voltages desired for calibrating. In as much as the thermo milliammeter and the resistor are flat to at least 1,000 cycles, the dc source was replaced with a low distortion, 1,000-cycle, power source. By producing the same currents at 1,000 cycles that were produced at dc, the same voltages which were accurately checked with the standard cell at dc were now being produced at 1,000 cycles. These standard levels were used at this frequency to check the accuracy of a Ballantine ac-dc precision calibrator which would serve as a convenient stable 1,000-cycle standard for further use in the testing. The calibrator is continuously variable in level up to 10 volts, (rms, peak, or dc) and is accurately read out to 4 significant figures. This instrument operates well within its rated ½% accuracy.

Audio to 2 mc

The 50-ohm resistor used in the transfer test was known to be flat to well above 2 mc and the thermo milliammeter was nearly flat with a known slope supplied by the manufacturer. By varying the frequency into the system, a calibration curve was produced for a Ballantine Model 310 vacuum tube voltmeter up to 2 mc. This calibrated meter was then used to monitor the input of the voltmeter in the range from 1 kc to 2 mc. For purposes of these tests, the nominal calibration voltage was required to produce the proper meter reading at 1,000 cycles. This nominal voltage was derived by calculating the attenuation of the attenuator to be calibrated from its measured dc values. The voltage was measured at the input end of the input cable. Since the cable length of 30 inches is quite short at 2 mc, it was not necessary to consider any change due to cable mismatch. However, the voltmeter and attenuator in combination were adjusted so that they presented a 50-ohm load having very low VSWR through most of the 1,000-mc band. The input cable is a special 50-ohm cable made to close tolerances for this application. It is necessary to repeat for emphasis at this point, that in any use of the low-level r-f measuring device and throughout all tests and calibration procedures, the dc resistance of the external circuit feeding the device's input cable is 50 ohms, because part of the dc return for the diode is through this path. The procedures to this point have an absolute calibration up to 2 mc, leaving 998 mc to be calibrated.

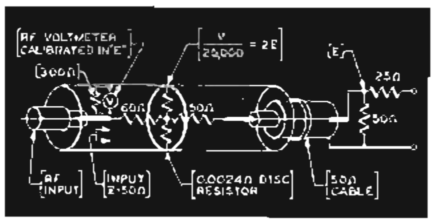

Figure 2. RF attenuator and voltmeter.

2 mc to 1000 mc

Voltage levels at higher frequencies are best measured with a bolometer bridge. In this way the accurately known 1,000-cycle voltages can be compared in their effect to the higher r-f voltages. There is in use in the BRC laboratories a specially constructed bolometer bridge which operates with more than normal sensitivity. This instrument is usable to compare levels of voltage down to 0.03 volts. Space is not available to describe the construction of this special instrument except to say that it compares the heating effect of accurately measured 1,000-cycle voltages to the heating effect of r-f voltages up to 1,000 mc. Since this bridge presents a load resistance of 50 ohms to a coaxial cable and its response is due to a heating effect, it is really a power measuring device and must be considered as such. The reason for this will be developed.

The r-f voltmeter is fed from a 50-ohm source having low VSWR and the level is adjusted until the meter gives standard indication. The signal generator output is then transferred without change to the bolometer bridge. The bridge is then balanced and the r-f is removed and replaced with enough 1,000-cycle energy to rebalance the bridge. The 1,000-cycle voltage is simultaneously measured on a Ballantine voltmeter whose calibration has been verified by the precision ac-dc calibrator. This voltage will then be a measure of the absolute voltage at the diode when certain corrections are applied. The output of the signal generator used was connected, by means of a specially adjusted terminating pad at the end of the coaxial cable, to the input cable of the low-level r-f voltage device.

Corrections

This cable and terminating pad were then transferred to the input jack of the bolometer. Since this jack is connected directly to the bolometer element, loss due to attenuation in the input cable of the low-level r-f voltage device had to be accounted for and subtracted from the indicated bolometer reading. This loss in the cable was calculated from the cable manufacturers rated loss per 100 feet and a knowledge of the cable length. The bolometer element is not exceptionally well matched compared with the low-level r-f voltage device, but since it is a power sensitive device and not a voltmeter, a VSWR as high as 1.5 does not cause an appreciable error. The amount of power reflected from a load having a mismatch of 1.5 is 4%. The resistance of the bolometer is precisely 50 ohms at the 1,000-cycle comparison frequency, so that the error in voltage is only the square root of the power error, or 2%. However, for precise work, the VSWR characteristic of the bolometer must be determined and this error taken into account. The indication will always be lower than actual, because the bolometer rejects some of the power delivered to it. Therefore, the error must be added to the indicated results.

Additional data was derived for design of an accurate voltmeter correction curve by actually determining the resonant frequency of the voltmeter. This was done by scanning the band from 800 to 3000 mc with a microwave signal generator which was known to be reasonably flat. A pronounced minimum in the output was observed at 1400 mc which represented the series resonant frequency. This information was used to confirm the slope of the curve obtained from previously used methods. While this type data does not give actual quantitative information as to the actual magnitude of the resonance, it is used to add credence to the previous quantitative measurements.

Attenuator Design

The basic concept of the attenuator is shown in figure 2. It is a voltage divider composed of a 60-ohm resistor in series with a 0.0024-ohm resistor. The input voltage is fed in across the series combination and the output is taken across the 0.0024-ohm resistor. Since the 0.0024 ohms is not a significant part of the total resistance, the attenuation ratio can be taken merely as the ratio of the two resistors, which in this case is 25.000. Of course, each of these two elements must be the same value from dc to 1,000 mc in order to obtain the desired results.

For the larger 60-ohm resistor, a natural solution was presented in an article by D. R. Crosley and C. H. Pennypacker.[1] It was demonstrated mathematically in this work that if: (1) the central conductor of a coaxial transmission line is a uniform resistive cylinder, (2) this transmission line is shorted on one end, and (3) the geometric dimensions of the line are such that its characteristic impedance X $\sqrt{3}$ is equal to the total series resistance of the central conductor; this section of line, when

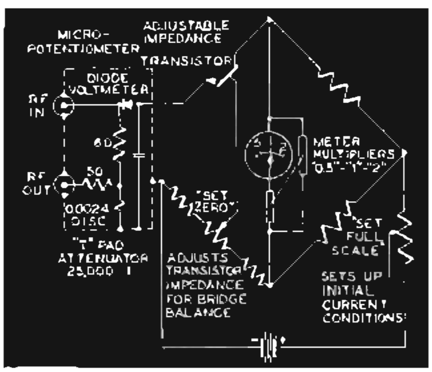

Figure 3. RF voltage standard-basic circuit.

viewed from the open end, will look like a pure resistor equal to the total series resistance of the central conductor. Compared to its dc value, this resistor would have a VSWR of 1.01 when the length of the line is less than 1/100 of the wave length, or less than 1.03 when the length of the line is less than 1/30 of the wave length. The central conductor of this line in practice is a glass rod onto which has been evaporated, in a vacuum, a thin film of metal. In this case, the film is thin enough to be considered without thickness for skin depth considerations. The line is 1 cm long, or 1/100 of the wave length at 300 mc and 1/30 at 1,000 mc. Therefore, in theory, the resistor is within 1% of the dc value at 300 mc and within 3% at 1,000 mc.

The 0.0024-ohm resistor becomes the short circuit at the end of the transmission line. A natural resistor for this type of use is suggested in an article by M. C. Selby.[2] This resistor consists of an annular film of conducting material which bridges the gap between the inner and outer conductors of the coaxial line. For purposes of evaluation, this film can be considered to be a series of square bars whose width is the film thickness. Assuming, for now, uniform penetration of current, the inductance can be evaluated. Consider one bar to be called a. Bar a will have mutual inductance with all other bars. For each bar at an angle ϕ from a, there will be another bar at $-\phi$ from a. These bars will have mutual inductance of equal value, but will have opposite sign and cancel. Bar a will then have mutual inductance with a bar 180° from it. Assuming that the inductance of the disc is the result of all bars in parallel that approximate the disc, (Inductance may be actually less than this because some area is unaccounted for.) total inductance is computed as follows:

$$\text{L Total} \leqq \frac{d}{2\pi r} \left(\text{La} + \text{Maa}_1\right)$$

$$\text{La} \leqq 0.002l \left[\log \frac{2l}{0.447d} - 1 + \frac{0.447d}{l}\right] \mu h$$

Where:
d = thickness of bar = 2.5×10^{-4} cm
l = length of bar = 0.25 cm
r = radius of inner conductor = 0.36 cm

$$\text{Maa}_1 = -0.002 \left[(2l + 2r) \log (2l + 2r) + 2r \log 2r - 2 (l + 2r) \log (l + 2r)\right] \mu h$$

$$\text{La} \leqq 1.329 \times 10^{-3}\ \mu h, \quad \text{Maa} \leqq -5.6 \times 10^{-5}\ \mu h$$

$$\text{LA} \leqq 0.1406\ \mu\mu h$$

The inductive reactance is $4.36 \times 10^{-4}\ \Omega$ at 500 mc and $8.72 \times 10^{-4}\ \Omega$ at 1,000 mc, and is in quadrative with the resistance. This results in a 1.65% error in impedance at 500 mc and a 6.25% error at 1,000 mc.

Uniform current in the bar is indicated, because the effective skin depth in the film at 1,000 mc is roughly 2 times the actual film thickness. It develops that when the skin depth is equal to the film thickness, the resistance is equal to 103% of the dc value, and when the skin depth is equal to twice the film thickness, the resistance is equal to 101% of the dc value. Since the current tends to be more dense on the input side of the annular resistor, this slight inequality of distribution tends to reduce the voltage appearing on the side of the film opposite the side from which the output is taken. This reduction tends to offset the increase due to the disc inductance.

Attenuator Performance Checks

In addition to checking the voltmeter characteristics it is desirable to check the attenuation at various frequencies. The theoretical attenuation is of course derived from accurate and careful dc measurements, but the r-f attenuation must be ultimately checked by judicious comparison with a precision piston attenuator. The piston attenuator can be a rigorously accurate device if used carefully, but it can also be a totally inaccurate device if used improperly. The attenuation of the useful mode in the circular wave guide is well known, but other modes are also propagated into the tube under some conditions. These modes are all attenuated at a rate higher than the normally used TE_{11} mode. If one does not use the attenuator with the probe too close to the driven end of the tube, and if a driving element is chosen which is of such geometry as to favor the TE_{11} mode, the calculated attenuation rate can be used quite safely. Careful checking of small increments of the attenuator output in the high output regions against small precision pads, should reveal the region where the attenuation rate starts to decrease as the driven end of the tube is approached. This region should be avoided. The TE_{11} mode is produced most purely by a symmetrically placed driving loop whose plane is precisely coincident with the plane of the pick-up loop. The pick-up loop should also be symmetrical in the tube.

In order to check the low-level r-f voltage device's attenuator, a signal generator was used to feed a precision piston attenuator. This attenuator output fed into the device's attenuator which in turn fed a very sensitive, stable receiver equipped with an easily read output meter. The piston attenuator output was increased to a level which gave a good sound reading above the noise on the carefully tuned receiver. Care was used to avoid the inaccurate region of the piston attenuator. The attenuator under test was then removed and the piston attenuator withdrawn until the receiver was observed to give the output reading it formerly had given. The measured attenuation was then the total length traveled by the piston times the attenuation per unit length. Using this procedure, the absolute attenuation of the piston is of no interest. In order that this replacement be valid, the piston attenuator output must be adjusted to 50 ohms. Fortunately, receivers operating at these microvolt levels are square-law detectors and therefore power measuring devices. This being the case, slight differences in mismatch between the output of the piston attenuator and the device's attenuator do not matter. The low-level r-f voltage device used as a standard at BRC checked against the piston attenuator within the readability of the measurement.

Standard Unit

The above tests were performed on a number of low-level r-f voltage devices and the best unit, in our judgment, has been retained as a standard. To control the quality of further units, it was necessary to determine to what precision the outputs of the various production units could be compared with this standard unit, considering the equipment to be used and the personnel who would be likely to make the tests.

The meter of the standard unit differs from a production model in that it is calibrated in percentage deviation from standard input. This is used to indicate how much the input voltage of the standard must be changed to produce an output which will have the same effect in a receiver as a unit under test. If the meter reads zero error, the unit is considered to be identical w·h the test unit at the test frequency. Of course, test frequencies are spotted all through the 0.1 through 1,000-mc band. To evaluate the precision to be expected, a considerable number of units were run through the same comparison tests three times by four different persons who are expected to run these tests during production. The results of these tests, shown in figures 4 and 5, were used to improve the operation of the receiver equipment in regions where the spread was unreasonable, and to incorporate the subsequent reasonable spread in the accuracy specifications to be published on the instrument. This tends to make the accuracy rating worse than it probably is, but in a device such as this these findings should be considered.

RF Voltage Standard Type 245-B

Articles concerning the design and application of the Type 245-B RF Voltage Standard have appeared in previous issues of the Notebook.[3,4] Briefly, the instrument is a very fine attenuator used in combination with a carefully

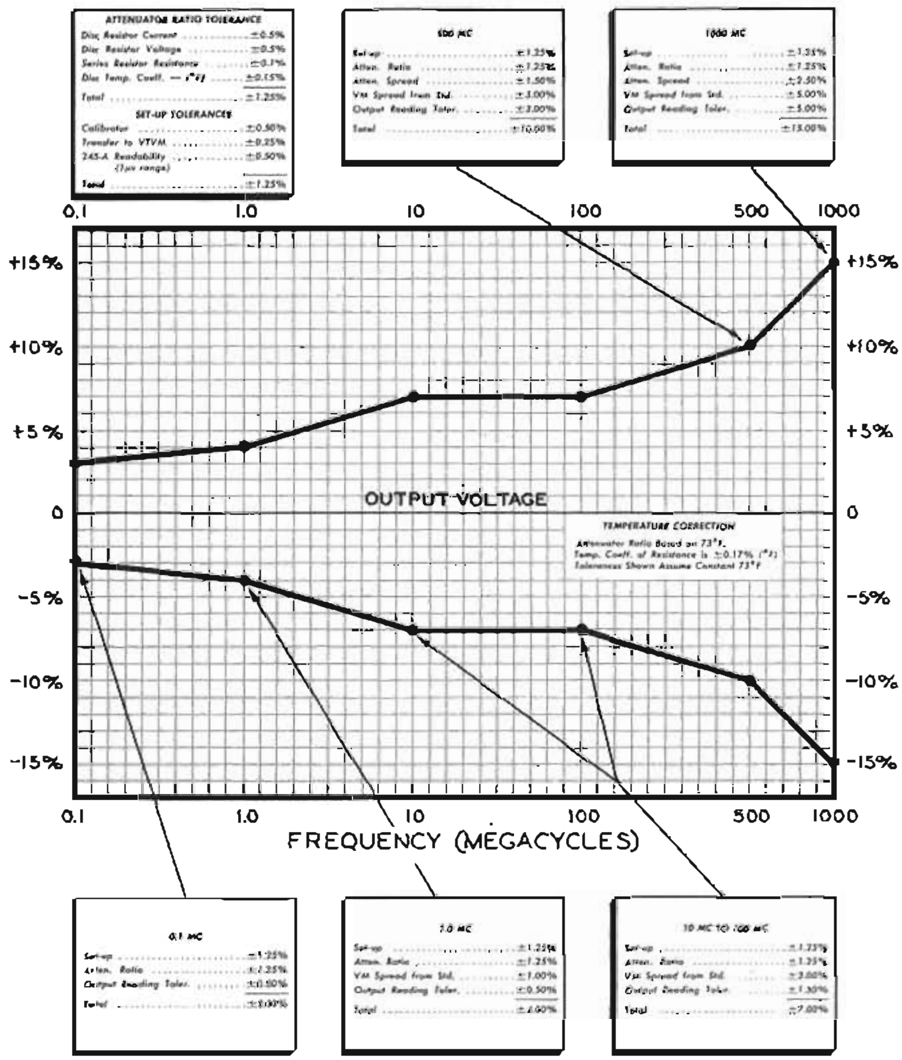

Figure 4. Production accuracy tolerance for RF Voltage Standard Type 245-B.

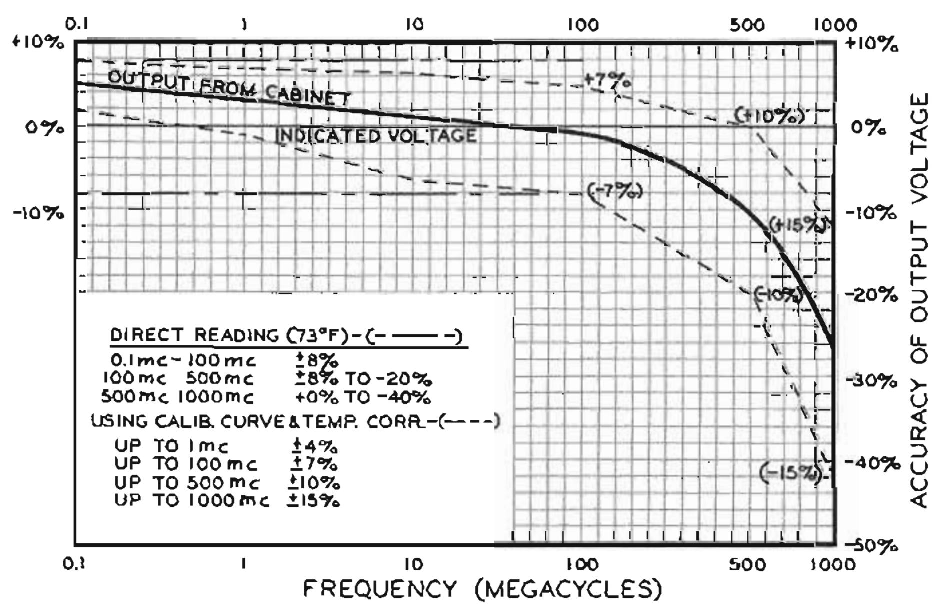

Figure 5. Accuracy of output voltage from RF Voltage Standard Type 245-B.

calibrated r-f voltmeter. When used in conjunction with a signal generator capable of producing 0.1 volt across a 50-ohm load, this device can serve as an indicator of the proper level which is to be fed to the fixed precision attenuator built into the device. The 245-A will deliver 2, 1, or 0.5 μv (depending on the voltmeter range selected) from the 50-ohm output cable of the internal precision attenuator. These levels can be considered standard levels, which are independent of the age, condition, or state of calibration of the signal generator used. The only limitation to be placed on the signal generator is that it have a dc output resistance of roughly 50 ohms (30-70).

The instrument is light-weight, operated from battery power, and small enough to be carried in a shoe box to the most remote locations. Using whatever generator may be on hand, the performance of the generator, or more important the performance of the receiving station, can be evaluated and compared with equipment in other locations.

Conclusion

It is apparent that measuring accurate voltage levels at frequencies of UHF and below is tedious and time consuming. The care which has been taken in its calibration should serve as encouragement to those who are willing to accept the Type 245-B RF Voltage Standard as a standard.

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The Use Of Standards With A Film Gauge

ANTS PIIP, *Development Engineer*

The Film Gauge, Type 255-A can be used for measuring film thicknesses of a variety of film-basis combinations, whether they be conductive films on nonconductive basis, conductive films on conductive basis, or nonconductive films on conductive basis. However, a calibrated standard is required for nearly every film-basis combination, if absolute measurements are to be undertaken. The preparation of these standards can usually be carried out by the user without too much difficulty. This article describes a few new kinds of standards, and gives a few pointers on how to increase the utility of prepared standards. The actual preparation of standards has already been covered in previous issues of the Notebook. [1, 2]

Actual Basis Material Slightly Different From The One Used On The Standard Card

The basis and plating materials of the piece to be measured and the standards should be identical if the meter readings are to have any value. However, if the basis materials are not too different, the instrument can still be made to give useful readings with the balancing procedure slightly modified as follows:

Set up and calibrate the instrument in the normal manner, using the samples on the standard card.

Without touching any controls, move the probe to a sample of the bare basis material actually used in the work. If the meter reading is not more than approximately one-half scale, the standard can be used with the new basis material.

With the probe on the new basis, rezero the instrument using only the SET BASIS control. The errors introduced by this small shift in the zero point are negligible. This method is applicable both on combinations having the same kind of conductive plating, or where the coating is a nonconductor (i. e., paint, ceramic, plastic, anodizing, etc.). In the latter case, the basis materials can differ by somewhat more than one-half scale; e.g., standards with an aluminum basis work perfectly on brass. The same procedure can also be followed for work which is not perfectly flat.

Inhomogeneous Basis Material

The situation is somewhat similar if the actual basis materials happen to be nonuniform (cold formed steels are notorious in this respect). The uniformity and variations-from-norm of the basis material can be checked by noting the 255-A readings on different spots or pieces of basis material. If a piece of coated material and one of the bare basis (identical to the basis on the first) are available, the feasibility of using the Film Gauge for film thickness measurement can be ascertained as follows:

Set up the instrument and adjust the sensitivity by means of the SET STANDARD control, until a nearly full-scale deflection is obtained with the thickest expected coating.

With these adjustments, analyze the actual basis materials to be used in the coating process. If the readings on the various basis pieces do not vary more than 10° from zero (basis), then the errors in thickness measurement should not become excessive. The readings will be unreliable for very thin coatings, where the deflections due to nonuniformities of the basis are comparable to those due to the coating. The readings above half-scale do however, give a reasonably true indication of film thickness.

Ferromagnetic Materials

Care should be exercised when measuring ferromagnetic basis and coating materials, to make certain that readings are ever increasing with coating thickness. Use a series of samples of known coating thicknesses for this purpose. If there is a dip, or even an apparent plateau in the thickness-reading curve, a reversal or a loop in the unrectified thickness-reading curve is indicated for this range of thicknesses and for the frequency being used. The meter circuit of the 255-A contains a rectifier, and therefore the instrument is incapable of distinguishing between positive and negative readings: both show up as positive. This results in ambiguities in calibration, producing identical rectified readings at three different thicknesses. If similar results are obtained at the other frequency position on the 255-A, the combination cannot be handled by the instrument. It should be noted however, that these ambiguous

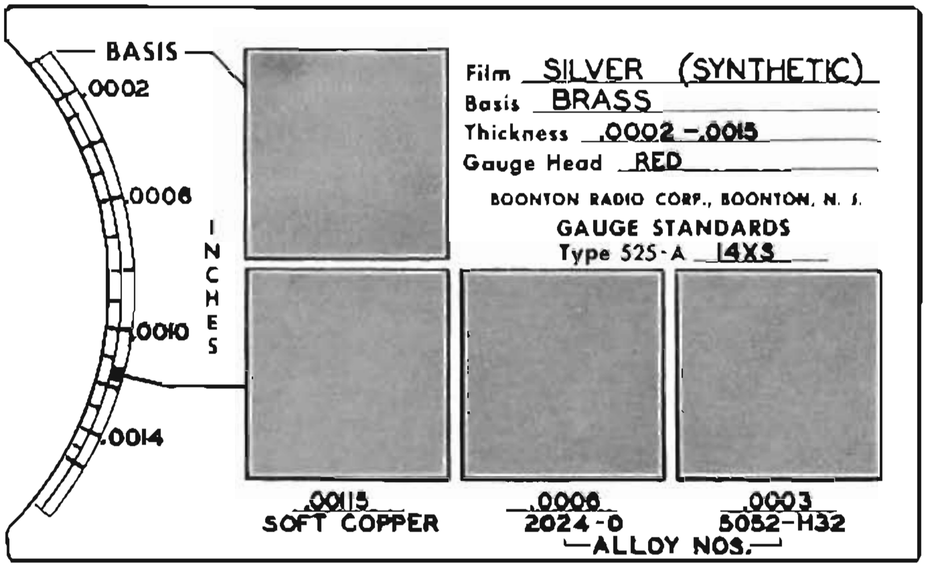

Figure 1. Gauge standard card utilizing synthetic material for plated samples.

loops in the calibration curve do not normally show up in both positions of the GAUGE HEAD selector switch; at least not for the same thicknesses.

Synthetic Standards

Several of the common plating metals are rather soft; e.g., silver, cadmium, etc., and calibration standards using thin films of these materials will have a limited lifetime of usefulness because of wear at the point of contact with the probe spacer rod. However, these plated samples can be simulated by homogeneous specimens (see Figure 1).

Once the calibration of the 255-A has been determined using the actual plated standards, a piece of a third material can usually be found that will give a deflection close to the *thick* end of the scale. After *apparent* thickness of this piece of material is noted, it can be used, together with a sample of the actual basis material, for calibrating the instrument. The original plated samples can be filed as "prime standards" and used only for preparing and checking calibration curves.

Since there is a multitude of alloy materials available (e.g., aluminum, brasses, bronzes, nickel silver, etc.), it should not prove too difficult to find suitable *synthetic* standards.

Because the *synthetic* standards are homogenous, wear caused by the probe tip will not change their conductivity and their "apparent thickness".

If possible, the *synthetic* piece should have a conductivity between that of the basis and plating materials. The "synthetic thickness" holds only for the frequency at which it was calibrated. A change in test frequency, will change the "apparent thickness" appreciably.

Extremely Thin Conductive Films

Measurement of extremely thin conductive films (less than one-tenth maximum measurable thickness) by the conventional method; i.e., with the instrument balanced on the bare basis, usually does not give good results, particularly if the basis is also conductive. One of the main reasons for this is the reduced sensitivity of the instrument at low meter readings.

Improved sensitivity can be obtained by using a modified calibration technique. Instead of initially balancing the instrument on the bare basis, a sufficiently thick (more than the penetration depth) film of the coating material is used for the reference point. The instrument is balanced first on this thick sample, then the probe is moved to the thinnest sample and the sensitivity adjusted to give a reading near

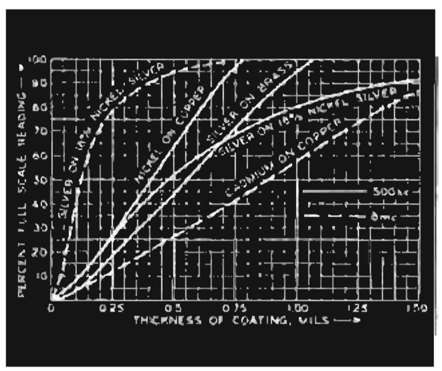

Figure 2. Typical samples of calibration curves.

full scale. With this setting the two intermediate thicknesses required for the generation of a calibration curve (figure 2) are measured. After the calibration has been established, the thick film can be cemented on the card in the space provided for the basis, and the instrument can be used in the usual fashion. Using this technique, the meter readings will be "upside down" (see figure 3.) compared to the normal method; i.e., thickest film at the top, thinnest at the bottom.

A thick, plated or deposited film is used for the reference instead of a solid slab of the film material, because deposited films are apt to differ somewhat from solid stock, leading to inaccuracies in measurement.

If it is desired to measure thin plated films; i.e., conductive films on conductive basis, a further refinement is advisable. The thinner the plating, the more sensitive the instrument becomes to variations in the basis metal. Therefore, to eliminate reading errors that

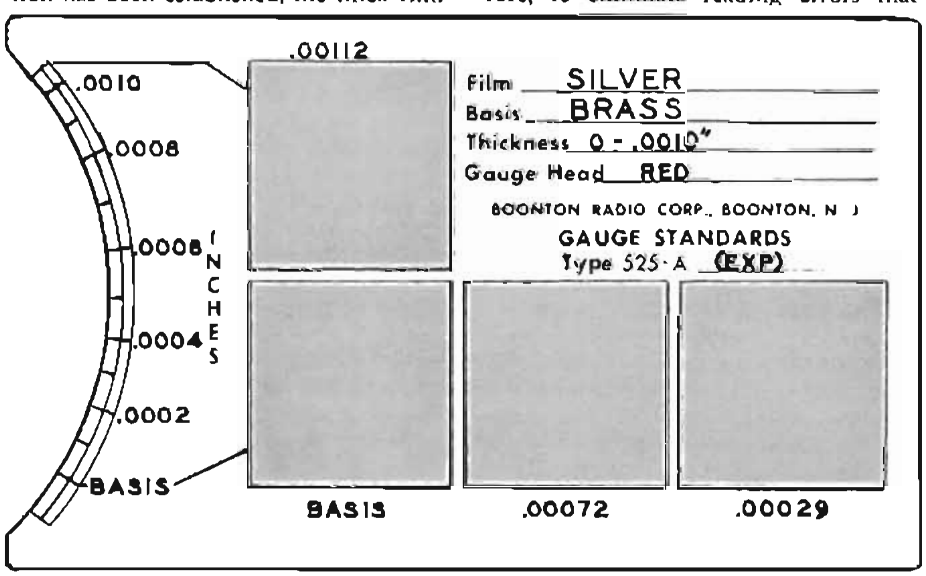

Figure 3. Gauge standard card used for measuring extremely thin films.

may be introduced because of variations of the basis, the last (thinnest) specimen of plating should be replaced with a piece of the bare basis. The calibrating or measuring procedure will be as follows:

Balance the instrument on the thickest film, using the SET BASIS control.

Adjust the sensitivity on the bare basis with the SET STANDARD control. If the actual basis should be slightly different from the one used in the standards, the instrument can be "touched up" by placing the probe on a piece of actual basis material and adjusting the SET BASIS control until the meter reading is the same as required by the standard. In doing this, be careful to keep the instrument tuned to the proper side of zero, i.e., in the direction where the meter does not pass through zero at the top end of the scale.

Thin Conductive Films On Nonconductors

Metal foils and metallizing are considered thin conductive films on nonconductors. It has been found that readings obtained for this type of film are rather insensitive to the probe-foil spacing. The readings remain unchanged from contact between probe and foil to a clearance of about 1/32 inch. Calibration of the instrument for this type of film can be performed simply by using multiple thicknesses of the foil to establish the calibration point. Imperfect contact between layers does not show up on the 255-A. This insensitivity of spacing makes possible the use of the instrument as a noncontacting foil thickness gauge.

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BRC Celebrates Shipment Of Its 10,000th Q Meter

FRANK P. MONTESION, *Editor, The Notebook*

On May 10, 1957, Boonton Radio Corporation commemorated its 23rd year in the instrument design and manufacturing field with the shipment of its 10,000th Q Meter. The occasion was marked at a special ceremony held at the Company's plant. Highlighting the ceremony were brief congratulatory talks by Mrs. W. D. Loughlin, widow of the founder of the company, and Dr. G. A. Downsbrough, President and General Manager of BRC. Dr. Downsbrough told the company employees that the 10,000th Q Meter would be given to Rutgers University, the State University of New Jersey, for use in its engineering laboratories. "It is befitting," he said, "that this instrument be given to an institution of higher learning, and that the institution be located in New Jersey, where BRC was established and still makes its home."

Following the talks by Mrs. Loughlin and Dr. Downsbrough, was a talk by Mr. Lawrence Cook, Quality Control Engineer and BRC's senior employee. Mr. Cook related some interesting and amusing facts about the company's rise from a six-employee, one-telephone concern to a full-grown manufacturing organization. The celebration ended with the serving of refreshments to all company employees.

Q-Meter History

The Q Meter was the first instrument to be designed and produced by BRC after the company was established back in 1934 by Mr. William D. Loughlin and several of his associates. Since that time, the words "Q Meter" and "Boonton Radio Corporation" have become almost synonymous.

Figure 1. Prior to shipment to Rutgers University in New Jersey, BRC's 10,000th Q-Meter is viewed by, left to right, Mr. L. Cook, Quality Control Engineer, Dr. G. A. Downsbrough, President and General Manager of BRC, Mr. B. Barth, Inspection Foreman, and Mr. T. O'Grady, Shipping Foreman.

At the time BRC was established, Q measurement methods were complicated, time-consuming, and often unreliable. The need for improved techniques was certainly eminent. Actually, the design of the Q Meter was undertaken to solve a specific problem encountered by a local concern engaged in the manufacture of hard-rubber coil forms. These coil forms were inspected by this company and met all of its requirements. However, when the forms were inspected by the purchasers, many were returned because they did not meet requirements. Investigation revealed that test instruments and techniques used by the manufacturer of the coils and the purchasers were different and therefore produced results which were not always the same. The problem of BRC's engineering staff then was to establish standard techniques for measuring Q. This was accomplished, and the operating principal and unique possibilities of the first Q Meter, Type 100-A, were presented in November 1934, at the Institute of Radio Engineers' fall meeting held in Rochester, N. Y. Soon afterward, the instrument was accepted as a standard by industry and research

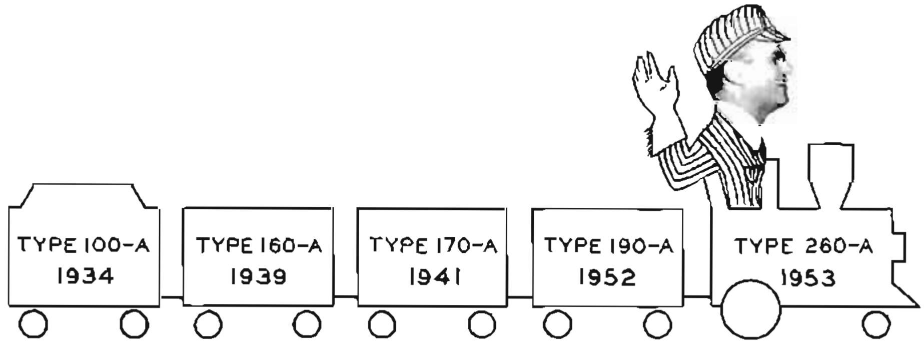

Figure 2. The Q-Meter has come a long way since 1934.

laboratories. Engineers and technicians in the growing radio industry received it enthusiastically.

With the advent of the Q Meter, simple, rapid, and accurate Q measurements became a reality. In the years that followed, improved models and broadened applications were introduced to keep pace with a rapidly growing industry (See figure 2). The Type 160-A Q Meter introduced in 1939, featured improved thermocouple shielding, more sensitive meters, and improved tuning capacitor design, adding together to provide for a much higher degree of accuracy in the high frequency range In the early years of World War II, another Q Meter, Type 170-A, was designed to handle measurements in the very high frequency range. More recently, the 160-A and the 170-A were superseded by the 260-A and 190-A respectively, the latter instruments including such modifications as: "Lo Q" and "ΔQ" scales, protection against thermocouple overload, power supply regulation, improved accuracy through the use of a newly developed annular insertion resistor, and other useful features.

Other Instruments

From this article, one might suppose that all of BRC's efforts during the twenty-three years since its establishment have been directed toward the development of the Q Meter. This is not the case. The engineering staff at BRC has been engaged in the development of numerous other electronic instruments which have found their way to electronic laboratories around the world.

EDITOR'S NOTE

Q Meter Winner

"The Q Meter is one of the basic instruments for any electronic laboratory and many hours of use have taught me to respect its accuracy and adaptability." Music to the ears of the BRC Sales and Engineering Departments were these words written by Mr. George S. Scholl winner of the Q Meter contest sponsored by BRC during the IRE convention in New York City last March. With his estimate of 338, Mr. Scholl outguessed almost 1,600 other hopefuls in trying to guess the Q of a coil displayed at the convention. Actual Q of the coil, as measured at BRC, was 336.7.

Mr. Scholl writes that he was born in Charleston, West Virginia and raised in Charlotte, North Carolina. After serving in the U. S Army during World War II, he earned his BS degree in Physics in 1948 at the University of North Carolina. During the next few years, he was employed as a physicist at the U. S. Naval Ordnance Laboratory. He returned to UNC for two years graduate work, earning his MS in Physics in 1953. He worked again at NOL, this time as an electronic engineer, until 1956 when he joined the American Machine and Foundry Co., Alexandria, Virginia, where he is currently engaged in the development of instrumentation for measurement of air blast pressures. Mr. Scholl is married, has no children, and asserts that his chief hobby is "just relaxing."

Mr. George S. Scholl, Research Engineer with the American Machine and Foundry Co. of Alexandria, Va., is shown with the Type 160-A Q-Meter he won with his Q estimate of a coil displayed in the BRC booth at the IRE show in New York last March.

FALL, 1957

NUMBER 15

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Calibrating An Inductance Standard

JAMES E. WACHTER, *Project Engineer*

Definition of Inductance Standard

Ask an engineer for the definition of an inductance standard and he will probably tell you that it is a coil or inductor having an accurately known, highly stable inductance. This is the definition implied. However, a better definition would be, an inductor having highly stable and accurately determined parameters; i.e., inductance (L), distributed capacitance (C_d), and resistance (R).

In such a standard, the parameters L and C_d are readily measured but the accurate measurement of R is extremely difficult. This is true because, in general, the more useful coils have a relatively high Q and consequently a very small value of R. R is so small in fact, that it is often swamped by the losses in any measuring equipment used, and therefore is very difficult to isolate. Since Q is a function of R ($Q=\frac{\omega L}{R}$), it follows that if Q and L can be determined, the value of R is firmly fixed.

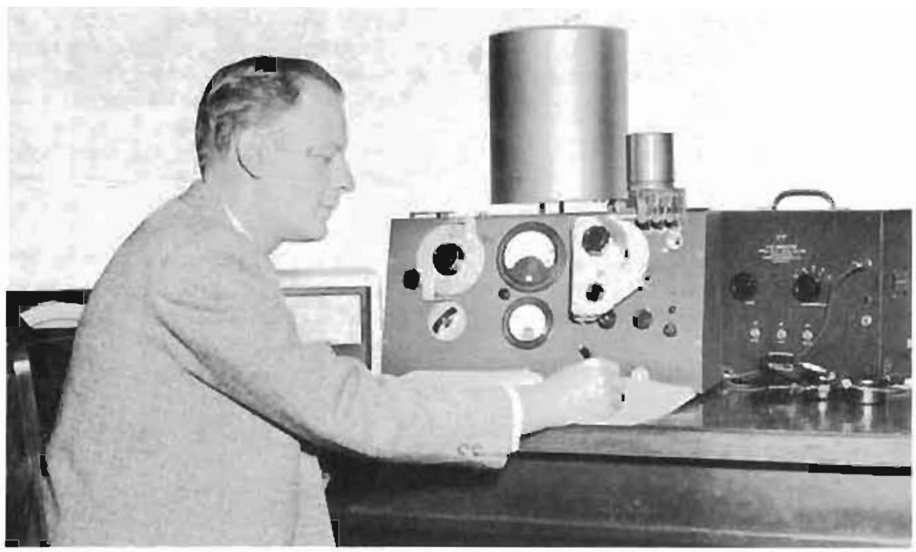

Figure 1. The author shown determining C at a V_o/V ratio, using the modified and specially calibrated Q-Meter.

Methods for Measuring Q

The problem now is to measure Q. An investigation into the possible methods of measuring Q has been made and the following conclusions drawn:

1. The frequency variation method involves the ratio of the frequency at resonance to the difference in frequency between the two half-power points on the Q-versus-frequency curve. This method was found unsuitable because of the variation of impedance of the Q circuit with frequency and the fact that the resonant frequency is different from the frequency for maximum voltage.
2. Injecting an AM signal into the Q circuit and measuring the attenuation of the side bands was rejected because errors in the amplitude of the side bands are caused both by coupling to the Q circuit and any asymmetry of modulation.
3. Injecting an AM signal into the Q circuit and measuring the phase angle of the detected signal was also rejected due to the error introduced in coupling to the Q circuit and the difficulty in accurately measuring the phase angle.
4. Q as determined from measurements on a BRC G-Meter, Type 192-A, is quite accurate and not difficult to determine, but because the G-Meter provides only two measurement frequencies, 1 mc and 30 mc, this means was found unsuitable as a general method.
5. A variation of the "Q by C" method was found to be most suitable, because the frequency remains fixed for these measurements, and the effects on the Q circuit due to varying frequency are eliminated. Also, the measurement requires basically an injection system, a Q circuit with a variable capacitor, and an oscillator, all of which are available in a Q-Meter.

Q Defined

Up to this point, the term "Q" has been used rather loosely. To aid in this discussion, it might be well to define here the several terms of Q with which we will be dealing:

Q — true Q of the inductor; i.e., $\frac{\omega L}{R}$

Q_e — effective Q; i.e., the Q of the inductor mounted on a Q meter, exclusive of all Q-Meter losses.

YOU WILL FIND . . .

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Q_e — circuit Q; i.e., the Q of the Q-Meter resonant circuit, including the inductor.

Q_i — indicated Q; i.e., the Q of the Q-Meter resonant circuit as indicated by the meter. This value includes the calibration errors of the Q-Meter.

Determination of Q

From the preceding definitions, it is seen that Q_e, the Q that the inductor appears to have when associated with additional circuitry, is the most useful value. If necessary, true Q can be derived from this value. To measure Q_e, a substitution method is used whereby the conductance of the Q circuit is determined first with a well shielded high-Q coil and again with the same high-Q coil plus the inductor being evaluated. The difference between the two determinations is the conductance of the unknown. This can be shown mathematically. Referring to the voltage-versus-capacitance curve (Figure 2) of the Q-Meter tank circuit, the conductance of the Q circuit with the high-Q coil can be expressed as:

$$(1)\quad G_1 = \frac{\omega C_{o1}}{Q_{e1}} = \frac{\omega C_{o1}}{\frac{2\,C_{o1}}{\Delta C_1}\sqrt{\left(\frac{V_o}{V}\right)_1^2 - 1}} = \frac{\omega \Delta C_1}{2\sqrt{\left(\frac{V_o}{V}\right)_1^2 - 1}}$$

The conductance with the two coils is then:

$$(2)\quad G_2 = \frac{\omega C_{o2}}{Q_{e2}} = \frac{\omega C_{o2}}{\frac{2\,C_{o2}}{\Delta C_2}\sqrt{\left(\frac{V_o}{V}\right)_2^2 - 1}} = \frac{\omega \Delta C_2}{2\sqrt{\left(\frac{V_o}{V}\right)_2^2 - 1}}$$

If the ratio $\frac{V_o}{V_1}$ is made equal to $\frac{V_o}{V_2}$, then the conductance of the unknown is:

$$(3)\quad G_x = G_2 - G_1 = \frac{\omega(\Delta C_2 - \Delta C_1)}{2\sqrt{\left(\frac{V_o}{V}\right)^2 - 1}}$$

With the conductance of the inductor known, it is a simple step to compute the effective Q:

$$(4)\quad Q_{ex} = \frac{\omega C_o}{G_x}$$

It should be noted that the inductor should have a Q of more than 100, since the foregoing equations have been derived using this assumption.

While the preceding analysis is straightforward, the actual measurements are involved. From equations (3) and (4) it is seen that because ω is readily determined to a high degree of accuracy using a crystal frequency calibrator or frequency counter, the overall accuracy is dependent upon the determination of C_o, ΔC, and V_o/V and the ability to repeat specific ratios of V_o/V.

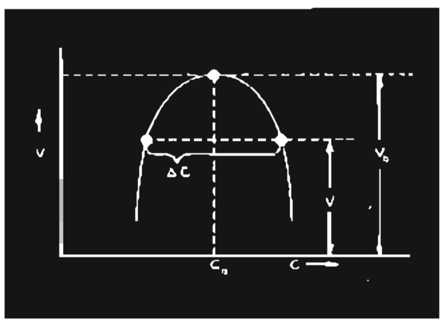

Figure 2. Voltage versus capacitance curve of a Q-Meter tank circuit.

To make the measurements we have used a modified and specially calibrated Q-Meter (Figure 1). Additional binding posts have been added to permit mounting two coils; one in the normal manner and the other from the HI post to ground. A means has been provided whereby the Q-Meter B+ voltage is externally regulated and monitored. Direct connection may be made to the cathode of the Q-voltmeter tube. In addition, a calibrated high-ratio gear drive is used to operate the main Q-capacitor and a parallel group of three micrometer type vernier capacitors, having a total range of about 100$\mu\mu$f, replaces the usual vernier. This permits the main tuning dial to remain in a fixed position while a wide range of ΔC readings is made.

Voltage Ratio

Setting up the V_o/V ratios necessitates the use of two specialized pieces of equipment; an instrument to provide precise levels of a 1000 cps signal (voltage calibrator) and an instrument to provide several very stable DC voltages (bucking voltage source). The 1000 cps source is used to calibrate the DC source in the following manner: The 1000 cps source is connected to the capacitor terminals of the Q-Meter, whose oscillator is made inoperative by setting between ranges. The DC source is connected through a microammeter to the cathode of the Q-voltmeter tube (See figure 3). Q-Meter zero is accurately set and periodically checked. The Q-Meter B+ supply voltage is also checked periodically to insure stable conditions. The level of the 1000 cps signal is adjusted to give a reading well up the Q scale of the Q-voltmeter. This is the resonance reference $V_o = 1$.

The DC source is then adjusted, in the 1 position, to give a zero reading on the microammeter in the lead to the Q-voltmeter tube, indicating that the voltage is of the same level as the voltage delivered to the Q-voltmeter. The 1000 cps source is now set to give exactly 0.9 of the previous signal and the DC source is switched to the next output and adjusted for zero meter reading. This is repeated for several levels; i.e., 0.8, 0.7, 0.6.

It can be seen that the DC source is now a memory for the various voltages appearing across the Q-Meter tuning capacitor, enabling the user to set and reset precisely to any desired voltage.

With the 1000 cps source removed and the Q-Meter oscillator set to the desired test frequency, the inductor (or inductors) to be measured is connected and the Q-Meter is resonated with the main and vernier capacitors. Capacitor settings are recorded and correspond to the resonating capacitance C_o. The DC source is set to the 1 reference position and the Q-Meter XQ level is adjusted to give a zero reading on the microammeter. The DC source is then switched to the 0.9 reference position and the Q-Meter is detuned on either side of resonance, using the vernier capacitors. The difference between the capacitor settings, at the point on either side of resonance where the current meter reads zero, is the ΔC value for the ratio $V_o/V = 0.9$. These capacitor settings are also recorded, and the procedure is repeated for the V_o/V ratios of 0.8, 0.7, etc.

Determining Capacitance

All that remains is to accurately determine the capacitance at the recorded capacitor settings before applying equations (3) and (4).

Modifications made to the Q-capacitors permit settings to be repeated to a very fine degree. This is necessary, because it is required that the Q-Meter be turned off to calibrate the capacitor.

The actual capacitance is measured by connecting a sensitive capacitance bridge to the capacitance terminals of the Q-Meter whose capacitors are set to a previously recorded value. The bridge is then balanced using a precision capacitor. All known corrections to the precision capacitor are applied and correction for the leads from the bridge to the Q-Meter is made.

The preceding is sufficient for the difference in capacitance (ΔC) data, but for absolute capacitance (C_n) data; additional corrections are necessary. A correction for the Q-voltmeter level is required, because the Q capacitance is measured with the Q-Meter turned off. This correction is determined through the use of a second Q-Meter (No. 2). The oscillator of the Q-Meter to be checked is disabled (set between ranges) and the capacitance terminals of this Q-Meter are connected to those of Q-Meter No. 2. Number 2 Q-Meter has a coil connected in its tank circuit and its oscillator is operated at the test frequency. The coil is selected so that some low value of capacitance is required to resonate with it. If the capacitance required is $80\mu\mu f$, then about $40\mu\mu f$ will be supplied by the Q-capacitor of each Q-Meter.

The Q-Meter under test is turned off and its vernier capacitors are used to resonate the tank circuit of Q-Meter No. 2; then it is turned on again, and the process is repeated. The difference in the two settings of the vernier is the capacitance attributable to the Q-voltmeter at the particular voltage (or Q) level. Different voltage levels can be selected by changing the setting of the XQ control on Q-Meter No. 2 and the capacitance correction versus Q level can be plotted as shown in Figure 4.

A final correction to the capacitance values may be necessary. When the Q-capacitor of a Q-Meter is calibrated, some small capacitance existing between the Q-Meter HI terminal and the cabinet (ground) is included in the calibration. A shielded coil connected to the coil terminals of the Q-Meter (shield connected to the LO terminal), causes some

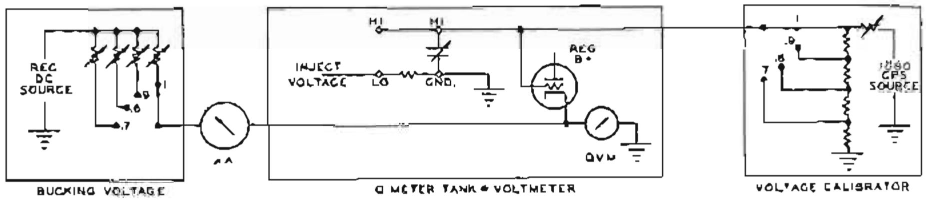

Figure 3. Connections used for setting up V_o/V ratios.

of the capacitance to shift from HI terminal to cabinet, to HI terminal to shield. This is called the proximity effect and is exceedingly difficult to define. In all but the most rigorous determinations, this effect may be neglected without seriously affecting the accuracy of the result. Without developing a lengthy and involved procedure for determining this effect, it may be said that generally any shielded coil having overall dimensions similar to the shielded coils manufactured by Boonton Radio Corporation (3-inch diameter shield cans) will produce a proximity effect of approximately $-0.4\mu\mu f$ when mounted on a Q-Meter with the coil base about 1 inch above the Q-Meter top panel. The figure will decrease with a decrease in the shield can diameter or an increase in the distance between the shield can and the Q-Meter.

Recapitulating the capacitance corrections:

C_i — Capacitance indicated by Q-Meter Q-capacitor.

$\pm C_c$ — Correction indicated by precision capacitor.

$\pm C_l$ — Correction for leads from capacitance bridge to Q-Meter.

$\pm C_r$ — Correction for Q-voltmeter level.

$-C_p$ — Correction for proximity effect.

It should be noted that for best results all measurements should be conducted in a temperature and humidity controlled atmosphere so that both the inductor under test and the measuring equipment are not affected by these conditions.

The ΔC values derived using the described procedures are used in equation (3) and a value of G_x is obtained for each V_o/V ratio. An indication of the care and accuracy of measurement is apparent by the degree of coincidence of the several values of G_x for each test frequency. The C_n value measured at each test frequency is used in conjunction with the average G_x value for that frequency in equation (4) to yield the effective Q of the inductor.

True inductance and distributed capacitance of the inductor can be found by applying data obtained from the previous measurements to the following equations:

$$(5)\quad L = \frac{\left(\frac{1}{f_1^2} - \frac{1}{f_2^2}\right)}{4\pi^2(C_{o1} - C_{o2})}$$

$$(6)\quad C_d = \frac{C_{o1} - n^2 C_{o2}}{n^2 - 1}$$

Where:
C_{o1} and C_{o2} are the capacitances necessary to resonate the coil at frequencies f_1 and f_2 respectively, and n is the ratio of f_2 to f_1.

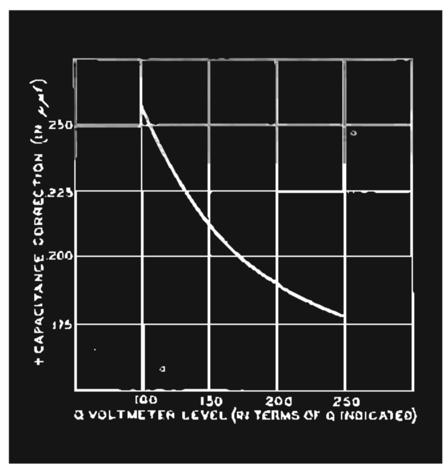

Figure 4. Capacitance correction versus Q-Level curve for a Q-Meter.

All of the true and effective parameters may now be determined by applying the following equations:

$$(7)\quad \text{Effective inductance } (L_e) = \frac{L}{1 - \omega^2 C_d}$$

$$(8)\quad \text{Effective resistance } (R_e) = \frac{\omega L_e}{Q_e} = \frac{R}{(1 - \omega^2 L C_d)^2}$$

$$(9)\quad \text{True } Q = \frac{\omega L}{R} \approx Q_e\left(1 + \frac{C_d}{C_o}\right)$$

BRC Q-Standard, Type 513-A

A specific example of an inductor developed through use of the described methods is the BRC Q-Standard, Type 513-A. The nameplate information for this inductor includes L, C_d, Q_e at three frequencies, and another term, Q_i, at the same three frequencies. In this case, Q_i is the Q that would be indicated by a Q-Meter having average loss and zero calibration error. The Q_i information was derived through an analysis of production indicated Q checks made on Q-Meters manufactured by BRC and relating the result to the measured Q_e value. Of course, in production the procedures just outlined would be impractical, therefore, a comparison method was developed. At each of the three frequencies involved, the production coil is compared to a standard, which has been established using the rigorous method described above. A precisely calibrated Q-Meter is used for the comparison, although its accuracy has only higher order effects on the results.

Suppose that at one of the frequencies, the difference between the production coil and the standard, as compared on a Q-Meter, is given by ΔC and ΔQ. The functions of standard and production coils would then be defined as follows:

Function	Known Standard Coil	Unknown Production Coil
Indicated Q	Q_i	Q_{ix}
Resonating C	C_o	C_{ox}
Effective Q	Q_e	Q_{ex}

Where:

(10) $Q_{ix} = Q_i + \Delta Q$

(11) $C_{ox} = C_o + \Delta C$

(12) $Q_{ex} =$

$$\frac{\omega(C_o + \Delta C)}{\frac{\omega C_o}{Q_e} + \frac{\omega C_o}{Q_i + \Delta Q}\left[\left(1 + \frac{\Delta C}{C_o}\right)\left(1 - \frac{\Delta Q}{Q_i + \Delta Q}\right) - 1\right]}$$

Using the same process to obtain difference data at the other two frequencies, Q_{ex} can be found at each frequency. Distributed capacitance and inductance of the unknown are given by:

(13) $C_{dx} =$

$$\frac{(C_{o1} + \Delta C_1) - n^2(C_{o2} + \Delta C_2)}{n^2 - 1} \text{ and}$$

(14) $L_x =$

$$\frac{\frac{1}{f_1^2} - \frac{1}{f_2^2}}{4\pi^2\left[(C_{o1} + \Delta C_1) - (C_{o2} + \Delta C_2)\right]}$$

Where the subscripts 1 and 2 refer to measurements at 2 frequencies.

To reduce the possibility of errors in manipulation, equations (12), (13), and (14) have been transformed to nomograms for use by production personnel (See Figure 5).

Figure 5. BRC inspector shown using a nomogram to determine the effective Q of a Type 513-A inductor.

A great deal of care has been taken in the physical design and manufacture of the Q-Standards to insure their stability. The coil form is mounted on a copper base, which is fitted to a shield can. The unit is hermetically sealed, evacuated, and filled with an inert gas to a pressure of 1 psi above atmospheric pressure. Leads are brought through the base to banana plug connectors, which may be replaced without breaking the seal. The high potential connector is isolated from the base by a low-loss ceramic seal.

The care taken in determination and production of the Q-Standards is attested to by the fact that not a single coil has been returned for mechanical failure or deterioration of electrical specifications. In some cases, Q-Standard users involved in government work have been required by the cognizant government agency to have their Q-Standards periodically rechecked by BRC. In each such case it has been found that the nameplate information was well within the original specification tolerance and no corrections of any kind were required.

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5. "A Standard for Q and L", BRC Notebook No. 12, Winter, 1957.

BRC manufactures two Q-Standards; the Type 513-A, discussed in the above article, and the Type 518-A. Each Type 513-A Q-Standard is individually calibrated and marked with its true inductance (L), distributed capacitance (C_d), effective Q(Q_e), and indicated Q(Q_i) at 0.5, 1.0, and 1.5 megacycles. Because these parameters are accurately known and highly stable, this standard may be used for providing precisely known supplementary Q-circuit inductance desireable for many impedance measurements by the parallel method, as well as a means for checking the Type 260-A and 160-A Q-Meters. The Type 518-A Q-Standard on the other hand is a precision inductor designed primarily for use in checking the overall operating accuracy of Q-Meters Type 260-A and 160-A.

A Linear Detector for FM Deviation Measurements

FRANK P. MONTESION, *Editor The Notebook*

With the development of the Type 202 FM Signal Generators in 1946, there arose a need for a device for the precision measurement of frequency deviation. Such a device would be required to provide laboratory accuracy and, at the same time, had to be simple and convenient for direct use on the production line. A detailed survey of available instrumentation revealed that such an instrument was not available and work was carried out at BRC on the development of this laboratory tool, concurrent with the development of the FM Signal Generator.

The result of this project was an early prototype unit, which over the years has undergone constant redesign and improvement to become known as the BRC Type 208-A FM Linear Detector. These instruments have been in constant use in our standards laboratory and engineering and production departments during this period, and are currently used to calibrate the Types 202-E and 202-F, FM-AM Signal Generators.

At the request of several customers who had a need to perform similar measurement of frequency deviation, the Type 208-A FM Linear Detector has been put into production and is now available for sale.

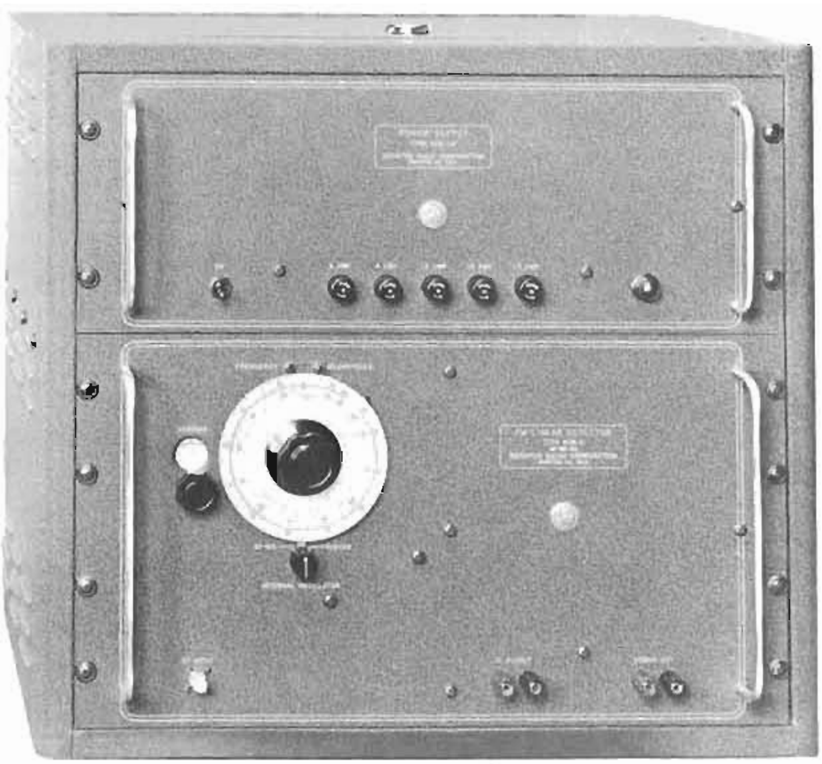

Figure 1. Type 208-A FM Linear Detector.

Operating Principles

The basic circuit elements of the FM Linear Detector are shown in block diagram form in Figure 2. The RF oscillator, doubler, amplifier-doubler, mixer, and RF amplifier stages of the Linear Detector are conventional circuits that operate to produce a signal usable for detection purposes. Actual detection is accomplished in the limiter and discriminator circuits shown in simplified schematic form in Figure 3. A type 6C4 triode, tuned over a frequency range of 27 to 54 megacycles is used as an RF oscillator. The output from the RF oscillator is fed to a Class C frequency-doubling stage tuned to the second harmonic of the oscillator frequency (54 to 108 megagcycles) which drives a Class C stage operating as a frequency doubling stage on the high frequency range (108 to 216 megacycles). The RF output from the doubler-amplifier stages, together with the output of the FM signal generator under test, are fed into a mixer stage. The difference frequency produced by the mixer is then fed through three stages of IF amplification to the limiter stage. After the first stage of IF amplification, a signal is fed through a cathode follower stage to the IF terminals for use in AM measurements.

The limiter stage squares the top and bottom of the signal wave. This square wave signal is then fed to the discriminator where it is converted to a single uniform pulse of current for each cycle. The current pulse rate follows the repetition rate or frequency of the incoming signal.

A low-pass filter is connected across the output of the discriminator to remove undesired signal frequency components and to allow the instantaneous potentials to rise and fall with each discharge of a current pulse. The demodulated output of the detector is a varying unidirectional potential directly proportional to the IF frequency.

Discriminator and Limiter

Referring to Figure 3, the input e_{in} is an FM signal with a carrier frequency between 100 and 300 kc The amplitude of this signal is sufficient to overswing the cutoff and zero bias limits of tube 807. During the part of the cycle when the tube is cut off, the plate potential will rise to the level of E_B. When the grid is positive, the plate current will rise to a maximum value. Any further increase in positive grid swing will not increase the plate current. It can be seen then that the minimum and maximum values of instantaneous plate potential in tube 807 are held constant, producing an output wave which is squared off, top and bottom, at definite fixed potentials and which is unaffected by possible variations in grid-swing amplitude. This square wave of plate voltage has a repetition rate equal to that of the input signal, e_{in}.

During the part of the cycle when the plate potential of tube 807 reaches its peak value (E_B), capacitor C_1 is charged through diode d_1 of tube 6H6 to a potential equal to E_B. When the plate potential of tube 807 swings toward its lowest value, capacitor C_1 discharges through diode d_2 of tube 6H6 in series with its load, R_3. This action causes one pulse of current to flow through resistor R_3 for each cycle of operation.

The total charge taken by capacitor C_1, once each cycle, is CE. (A small bias voltage in series with the charge diode d_1 effectively overcomes the contact potential of both diodes d_1 and d_2, therefore, the effect of this potential may be discounted.) The portion of this total charge which passes through diode d_2 and resistor R_3 of tube 6H6 is equal to the total charge (CE_{ll}) minus the residual charge (Ce_{min}).

With E_{ll} and e_{min} held constant, and the time constants of the charge and discharge circuits sufficiently small compared to the input wave period, the total quantity of current flowing through R_3 during each cycle is constant. An increase in the repetition rate (frequency) of the incoming signal will increase the number of current pulses per unit of time, thereby increasing the average value of current flowing through R_3. Conversely, a decrease in the repetition rate of the incoming signal will decrease the average current through R_3. The average potential across R_3 then is a perfectly linear function of the frequency of the incoming signal and the dynamic operation of the detector will result in essentially distortionless detection.

Calibration

The Linear Detector is accurately calibrated at the factory to provide a voltage versus frequency deviation coefficient for frequency deviations up to ± 300 kc at an intermediate frequency of 350 kc. However, it is advisable, because of component aging, to recalibrate the instrument periodically during normal use. Either of two methods, the Static Method or the Bessel Zero Method, may be used to accurately calibrate the Linear Detector.

Static Method

The Detector is interconnected with an FM signal generator, a frequency calibrator, and an accurate DC measuring device. The generator is connected to the Detector's RF INPUT terminals, the frequency calibrator is connected to the IF OUTPUT terminals, and the DC measuring device is connected to the

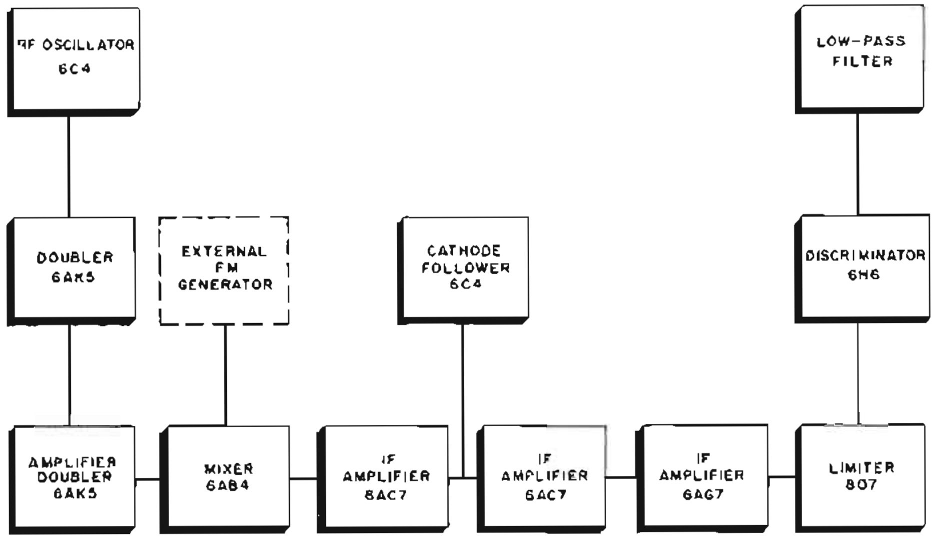

Figure 2. Basic circuit elements of the Type 208-A FM Linear Detector.

DEMOD OUT terminals.

With the frequency dials on the signal generator and the Detector set to the same frequency, the DC measuring device will read zero. Advancing the Detector frequency dial to provide 100 kc difference frequency, as indicated by the frequency calibrator, will cause a voltage reading to be indicated by the DC measuring device. This reading is noted and the Detector frequency dial is advanced again until a 200 kc difference frequency is indicated by the frequency calibrator. The voltage reading at the DEMOD OUT terminals is again noted. This procedure is repeated at each 100 kc increment until a 1 megacycle signal is indicated by the frequency calibrator, the output voltage being recorded for each step. The voltage readings obtained are then plotted on a graph with voltage as the "Y" axis and frequency as the "X" axis. The resultant curve will yield a straight-line section, whose upper and lower limits indicate the frequency points in the frequency spectrum between which linear operation of the Detector should be expected. With the DC voltage for these two frequency deviations known, any deviation (within the linear limits) may be ascertained by using the frequency versus voltage coefficient.

Bessel Zero Method

This method of calibration requires the use of a signal generator, an accurate 10-kc audio signal source, and a heterodyne-type receiver containing a BFO. The RF output of the signal generator is connected to the receiver and the receiver is tuned to the unmodulated carrier frequency of the generator to obtain a beat frequency of several hundred cycles, using the receiver's BFO. This frequency is monitored with earphones or a voltmeter. With the signal generator modulation control set to produce a 10-kc FM modulating signal, the FM deviation control is advanced slowly.

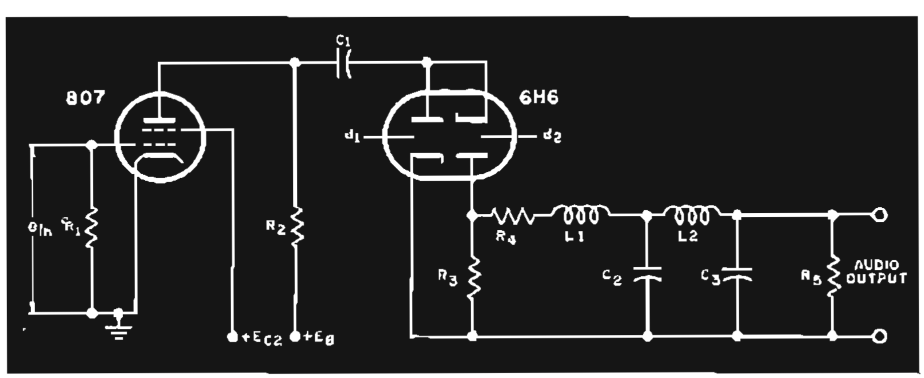

Figure 3. Basic detector circuit.

The beat frequency will disappear at certain points as the deviation is increased. These null points correspond to specific modulation indices, the first several being: 2.405, 5.520, 8.654, 11.792, and 14.931. Frequency deviation values at these null points are then calculated using the Bessel function as follows:

$$B = \frac{\Delta F}{f}$$

Where: B = modulation index
F = frequency deviation (kc)
f = modulating frequency (kc)

After the first null is detected, the receiver is disconnected and the generator signal is fed to the RF INPUT terminals on the Linear Detector. The Detector is tuned to 350 kc IF frequency deviation to insure operation within the linear region. A peak-reading AC voltmeter connected to the DEMOD OUT terminals on the Detector will indicate the voltage output for the deviation calculated for the first null. The procedure is repeated for each null point, and the voltage output obtained for each calculated frequency deviation is recorded. These voltages are then plotted against frequency deviation to produce a curve whose straight-line portion again indicates the linear limits of the Detector.

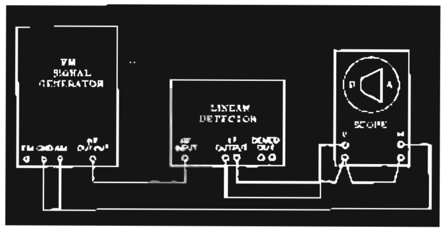

Figure 4. Connections for percent AM measurement showing trapezoidal display.

Application

The FM Linear Detector, as was previously explained, has been designed especially for the measurement of FM frequency deviation. With the instrument calibrated as explained above, its frequency deviation versus voltage output coefficient is known. Measuring frequency deviation becomes merely a matter of multiplying the voltage output reading at the DEMOD OUT terminals of the Detector (produced as a result of the FM signal fed into the Detector) by the frequency deviation versus voltage coefficient.

The Detector may also be used to measure the degree of amplitude modulation of a signal source. For this measurement, an external signal source is mixed with a signal produced by the Detector to provide a difference frequency of approximately 100 to 150 kc at the Detector's IF OUTPUT terminals. This difference frequency is then displayed on an oscilloscope as a trapezoidal pattern, similar to the trapezoidal pattern shown in Figure 4. The lengths of the vertical sides of the pattern (A and B) are then applied to the following equation to indicate the percentage of amplitude modulation.

$$\text{Percent AM} = \frac{A - B}{A + B} \times 100$$

Conclusion

The Type 208-A FM Linear Detector is a precision instrument capable of performing accurate FM and AM measurements in the engineering laboratory or on the production line. It will doubtless become a valuable aid to those customers who have measurement problems of this nature.

BRC Film Gauge Used To Measure Aircraft Organic Finish Thickness

The Glenn L. Martin Company Reports On a New Method Approved by The Navy

At the request of the Bureau of Aeronautics, The Glenn L. Martin Company of Baltimore, Maryland recently prepared a report entitled, "A New Method for Measuring Aircraft Organic Finish Thickness," which describes a new method employing the Boonton Radio Corporation Type 255-A Film Gauge.

The Film Gauge was introduced to The Martin Company by BRC as a means for measuring thickness of metallic plating finishes over nonferrous metal surfaces and its principle of operation was later proposed as a means for measuring aircraft organic finish thickness. The Martin Company subsequently conducted a research program to test the suitability of the Film Gauge for this purpose.

Many finishes were tested by The Martin Company to find a correlation and mode of operation which would prove that practical measurements could be made with the Film Gauge. A correlation was found, preliminary calibrations were performed to confirm it, and a report proposing the new method was issued by The Martin Quality Division Laboratory. This report was later submitted to the Bureau of Aeronautics who gave tentative approval of the method. The method was then evaluated by the Naval Air Material Center Laboratory. Results of this evaluation concurred wih Martin's and full acceptance of the method was published in a report from Naval Air Material Center.

The report prepared by The Martin Company at the request of the Bureau of Aeronautics was prepared by Norman R. Keegan, Chemical-Physical Engineer. In his report Mr. Keegan describes in detail the methods used in determining the accuracy of the Film Gauge for this specific application. Step-by-step procedures are given for operation of the Film Gauge for finish thickness inspection, and special instructions are included for specific applications.

The Martin Report has been reprinted by Boonton Radio Corporation for distribution to interested customers. Copies will be furnished upon request.

EDITOR'S NOTE

BRC Promotions Announced

The appointments of Mr. Frank G. Marble as Vice President and General Manager and Mr. Harry J. Lang as Sales Manager effective July 1st were announced by Dr. George A. Downsbrough, President of Boonton Radio Corporation.

Mr. Marble, formerly Vice President-Sales at BRC, succeeds Dr. Downsbrough as General Manager. Dr. Downsbrough will continue as President, Treasurer, and a Director.

Frank G. Marble

Mr. Marble has been associated with Boonton Radio Corporation since 1951. He served as Sales Manager until 1954 when he was appointed Vice President-Sales.

Prior to his associaion with BRC, Mr. Marble's career covered a broad field of engineering experience. He held design and development posts for seven years with Philco Radio and Television Corporation and Western Electric's electrical research division. Later, he served in engineering administrative positions with Bell Telephone Laboratories and Pratt and Whitney Aircraft Corporation. During the three-year period just preceding his association with BRC, Mr. Marble served as Sales Manager for Kay Electric Company.

In 1934 Mr. Marble received his BS degree in Electrical Engineering from Mississippi State College. He earned his MS in Electrical Engineering from the Massachusetts Institute of Technology in 1935.

Mr. Lang joined Boonton Radio Corporation in 1949 as a production engineer and successively served as project engineer, contracts engineer, and sales engineer. In 1954 he left BRC to study for his master's in Business Administration at the Harvard Business School.

Before returning to BRC, Mr. Lang served as sales engineer in charge of sales for the newly-formed Industrial Products Department of Airborne Instruments Laboratory.

Mr. Lang received his BS degree in Electrical Engineering from the Massachusetts Institute of Technology in 1949. During his studies there, he was an engineering trainee at Western Electric Co., New Jersey Bell Telephone Co., and Bell Telephone Laboratories.

From 1945 to 1946 he was a junior engineering officer with the U. S. Merchant Marine.

Harry J. Lang

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NUMBER 16

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

A Crystal-Controlled FM Signal Center

CHARLES G. GORSS, *Development Engineer*

Figure 1. The author is shown final checking the RF tuning unit on one of the first production models of the TYPE 242-A FM Signal Generators.

A New Concept

As communications systems get more precise and complicated, the demands placed upon the design of equipment required to test and calibrate these systems become more severe. One soon finds that the old standby signal generator no longer will do the job. Old concepts of high-frequency accuracy and stability are inadequate in systems crammed one next to the other, where each channel carries several subcarriers, and each subcarrier in turn carries its own information. In addition, with technical pursuits being carried on from pole to pole; in the arctic cold, the desert heat, the destructive humidity of the jungle, and the thin air of the higher altitudes; the signal generator must now be designed to perform in environments strangely different than the cozy laboratory or factory. The signal generator described in this article employs design techniques which meet the challenge of these stringent modern requirements.

YOU WILL FIND . . .

Design Considerations

First of all, the signal generator under consideration must be crystal controlled or stabilized with an accuracy in the order of 0.002 per cent. This is a typical long-term accuracy figure for a crystal oscillator. Since many changes of frequency may be required in the set up of multichannel systems, the test frequencies must be selected quickly and easily. Even though frequency control by crystals implies a finite number of channels, provision must be made for easy selection of between-channel frequencies for bandwidth determination and for the selection of off-beat frequencies. Frequencies so selected should also be reasonably accurate and stable. The generator should be frequency modulated and feature a wide deviation range (in the order of 300 KC) with little distortion. In addition, an optional compressor should be provided so that large changes in input modulation, caused by adding several subcarriers at one time, does not result in over modulation. The compressor should not increase distortion appreciably. Besides handling a wide range of external modulating frequencies, the generator should provide at least two different internal modulation frequencies, so that simple tests can be performed with no extra input equipment. Last but by no means least, the instrument must be rugged in design and must be able to survive not only wherever man can survive, but a few places where he cannot survive for long.

Design Techniques

Selection of 1 MC Frequencies

A frequency selection system employing two dials (Figure 2) was chosen to cover the range of frequencies from 400 to 550 MC in one megacycle steps. The dials are mounted side-by-side nearly touching each other. The first two digits are on the left-hand dial and range from 40 to 55 in 16 steps. The right-hand dial carries the last digit ranging from 0 to 9 in 10 steps. Every whole number from 400 to 550 can be selected by rotating both dials and hori-

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zontally aligning the three digits. The dials are masked so that only the numbers in use are clearly seen. However, adjacent numbers, visable through a colored transparent portion of the mask, serve to indicate whether dial rotation is toward the high or low end of the range. Detents are provided on each dial to index each selected digit in proper alignment.

Each dial is mechanically connected to the selector switch in a crystal oscillator which in turn electrically selects the proper crystal as the dial is indexed. The outputs of the two crystal oscillators are added together in a mixer which is designed to minimize spurious signals. This sum is then doubled twice and the resulting quadrupled product is added in another mixer to the output of a fixed-frequency oscillator which can be frequently modulated. If this oscillator output is frequency modulated, the sum frequency will reflect the same amount of modulation. The output of this mixer is doubled twice more to produce the final frequency which has four times the frequency modulation originally present.

Since each step of the left-hand dial spans 10 MC and each step of the right-hand dial spans only 1 MC, the increments (in actual crystal frequencies) in the left-hand or tens oscillator are 10 times the increments in the right-hand or units oscillator. There are 16 crystals in the tens oscillator spaced 1/16 of 10 MC apart, and 10 crystals in the units oscillator spaced 1/16 of 1 MC apart. The 1/16 spacing is a direct result of the fact that the frequency is multiplied 16 times between the oscillators and the output. The actual frequency ranges through the system are shown in figure 3.

There are several tuned stages through the chain of doubling and mixing, all of which must be tracked with frequency. A variable capacitor drive using a unique two-speed differential gearing system has been employed for this purpose. Input from either the tens shaft or the units shaft will turn the tuning capacitor shaft. One step on each shaft is originally the same angular displacement. However, the angular rotation of the variable capacitor caused by one step of the tens shaft is ten times the angular rotation of the variable capacitor caused by one step of the units shaft. The drive from the two dials to the differential is accomplished by two very low-backlash belt-drive systems using beryllium copper belts and spring-loaded phosphor bronze antibacklash cables. The differential system itself and all associated gears are also spring loaded to prevent backlash and assure optimum tracking.

Selection of Inbetween Frequencies

The combination of switching and capacitor-drive mechanisms allows selection of each even megacycle from 400 to 550, but an additional device is required to provide selection of all of

Figure 2. Frequency selection dials on the Type 242-A FM Signal Generator.

the inbetween frequencies. This is accomplished through the use of a highly stable "Clapp" type variable frequency oscillator which is coupled to the units oscillator dial by means of another beryllium copper belt system. The oscillator produces the same frequencies which the units crystal oscillator produces at any given dial setting and tracks right along with the units crystal oscillator. An internal beat detector coupled to the output monitor meter allows the user to bring the VFO into exact zero beat with the crystal oscillator by tuning a trimmer in the VFO circuit. Normally the crystal oscillator is coupled to the first mixer, but a panel switch allows the VFO to be substituted. A clutch connected to this switch shaft disconnects the normal switch detent and substitutes a stop device which permits 1 MC rotation either side of the point of crystal calibration. The smooth curve of the VFO capacitor and the direct calibration with the crystal oscillator at a near frequency combine to give a relatively high degree of accuracy to all frequencies selected by the VFO.

Frequency Mixing

The general subject of mixing in an FM signal generator could well be the subject for an entire article but a brief explanation of the problem is certainly worth outlining here. The two inputs to the first mixer are in the general ranges of 3.5 to 4 MC and 10 to 18 MC, and it is entirely possible that harmonics of the 3.5 to 4 MC input might lie close to the desired sum output which ranges from roughly 13 to 22 MC. Should this harmonic be present in the plate circuit of the mixer which would be tuned to the sum, and should it be perhaps within 100 KC of the sum, it is entirely possible that this harmonic would not be far enough off the sum to be rejected. The unfortunate result is the familiar sum of two vectors whose frequencies differ somewhat; i.e., an amplitude-modulated wave whose phase is advancing and retarding at the difference frequency (Figure 4). This shifting phase can be considered as phase modulation which at a given modulating frequency can also be considered as frequency modulation.

The amplitude modulation is not very alarming because it can be removed by limiting. Frequency modulation, on the other hand, is not easily erased and becomes a problem in an FM signal generator because extra modulating signals, which have no relationship to the desired modulation signal, are undesirable. Since erasing is difficult, the best way to avoid this unwanted modulation is to avoid the circuit conditions that cause it.

First of all the frequencies must be selected which will produce an absolute minimum of possible harmonics approaching the sum frequencies. This may however leave some possible trouble situations even after the best possible choice of frequencies.

Next, make sure that the lower frequency input is as pure as it can be. A low-pass filter cutting off sharply above 4 MC proved to be a very effective way of accomplishing this. Since even a very clean signal can be distorted in the mixer, it is very important that the lower frequency be introduced on the grid or element whose transfer characteristic is the straightest. The signal level of the lower input should also be kept as low as possible and the element should be so biased that the total swing takes place on the most linear portion of the transfer characteristic. It was found desirable to use a 6AS6 tube as a mixer with the 3.5 to 4 MC signal fed into grid number 1 and the higher frequency fed into grid number 3. While this is not the most efficient operation (with only ½ volt on grid number 1) it does greatly reduce the distortion of the signal on grid number 1. Any resulting harmonic content can be additionally reduced (if the tuning is accurate enough) by tuning the plate of the 6AS6 to the sum and following this stage with another sharply tuned stage. In fact, it is necessary to tune these stages for minimum spurious FM in the final set-up procedure.

Carefully following the above principles results in a minimum of spurious FM. Ignoring these facts would easily result in 100 KC or more spurious FM deviation in the output.

Frequency Doubling

The need for doubling to frequencies up to 550 MC with a minimum of harmonic production and a maximum of power production led to the use of a fairly interesting technique. Electron tubes are not suitable for this purpose because they produce a fairly strong string of higher harmonics with little energy where it is wanted. Theoretically the commonly used full-wave rectifier circuit, operating from a center-tapped transformer, is ideally suited as a doubler because it produces a great amount of second-harmonic voltage in its output and very little of anything else. In fact, it produces almost half of the voltage at the second harmonic as is put into it at the fundamental frequency. The problem however is to set up these ideal conditions at UHF frequencies.

First of all, the impedance of these balanced transformer secondaries must be kept low to avoid unwanted resonances and capacitive shunting. In practice the untuned balanced secondary is mounted on the grounded end of the tuned primary winding. Each half of the winding is approximately a single turn, with the center tap slightly off center to compensate for the fact that one turn is closer to the primary.

The Transitron type T-6 diode proved to be very desirable for this use. It has a very high forward conductance and apparently switches rapidly enough. The forward conductance is important because these doublers must drive the cathode of a grounded-grid amplifier which does not present much more than 100 ohms resistance. Diodes which are normally useful to much higher frequencies were not useful in this case because of their poorer forward conductance and a resulting high internal loss.

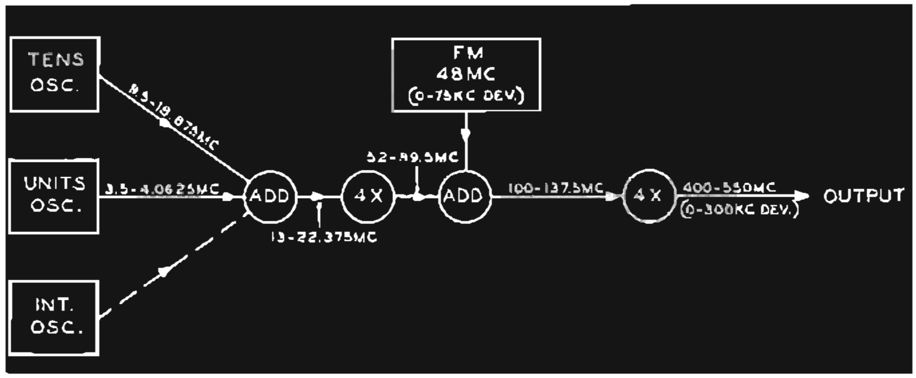


Figure 3. System for providing Type 242-A output frequencies.

The output of the doubler circuit is capacitively coupled to the cathode of the following grounded-grid stage and the diodes are allowed to develop their own DC bias in a load resistor on their side of the coupling capacitor. A choke in series with the DC load prevents loading of the RF circuit. Bias is adjusted to a compromise optimum value. The diodes must be allowed to conduct enough current to insure a low forward conductance, but must not be allowed to conduct so much current that the barrier capacitance is greatly increased and the switching rate slowed down.

Amplifiers

All of the lower frequency amplifying stages are fairly conventional plate-tuned 6AK5 stages, but from 200 MC up, the 6AN4 tube has been employed in a grounded-grid configuration. These grounded-grid stages are somewhat unique in that they are series tuned in the plate circuit. This takes the form of a few turns of wire from the plate connected in series with a tuning capacitor to ground. This type of tuning increased the highest frequency attainable from a given tuning capacitor by separating the plate capacitance of the tube from the minimum capacitance of the capacitor by means of a coil connected between them. The tank can be tuned almost up to the frequency where the minimum value of the variable capacitor alone resonates with the tank coil. However, the lowest frequency obtainable is limited by the condition where the variable capacitor reaches infinity and becomes a short, causing the plate capacitance to resonate with the tank coil. This puts two very rigid limits on the tuning range and calls for an unusual shape plate for the tuning capacitor. Since the minimum capacitance of the tuning capacitor can generally be made much lower than the tube plate capacitance, this technique allows tuning to higher frequencies with lumped circuit components than are obtainable with parallel tuning of the plate. Power is coupled from this type of device by a turn or two of wire around the tank coil.

Frequency Modulation

As shown in Figure 3, the frequency modulated oscillator is injected into the second mixer and always operates at a center frequency of 48 MC. This oscillator and the control circuit which is required to hold it within 0.001 per

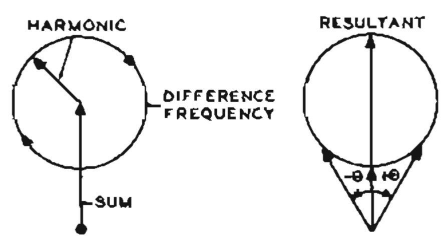


Figure 4. Vectors illustrating how certain harmonic frequencies may cause frequency modulation.

cent, is very complex and would very likely be ample subject matter for another article. Briefly, the oscillator is a triode oscillator which has a pentode reactance tube tied across its tank circuit. A part of the output of the oscillator is amplified and mixed with a very stable 47.9-MC crystal oscillator. The 100-KC IF frequency is then amplified and coupled into a frequency counting circuit which translates frequency shifts to DC level. This DC level is amplified and connected back to the reactance tube grid by means of suitable bucking voltages. The audio modulating voltages are also fed to this grid. In order that the reactance tube may be operated as closely as possible to the center of its linear characteristic, an electron-ray indicator tube in conjunction with a trimmer is provided on the front panel. With these, the operator can set the oscillator as close to center frequency as he can with the automatic frequency control circuits defeated. This precludes the need for compensating for long-term drift on these circuits. A check of this electron-ray indicator after every two or three hours of actual use will usually suffice to maintain low distortion. A spring-loaded switch on the AFC defeat prevents accidental defeating of the AFC when the generator is in use. A suitable indication on the indicator assures the operator that the control circuits are locked in.

Audio Circuits

The audio circuitry in this generator must do several things. First, it must supply internal modulating signals at 400 cycles or 1,000 cycles as required. Next, it must handle external signals between 300 cycles and 100,000 cycles and, at the same time, maintain a phase-shift characteristic which is essentially linear with frequency. There must be facility to handle a number of subcarriers and optional compression circuitry which will maintain constant deviation as subcarriers are switched on and off, without substantially increasing the distortion in the audio channel. The deviation indicator meter on the panel must accurately indicate peak deviation regardless of the subcarriers present. To accomplish this, the meter circuit must be a peak-to-peak rectifier with a long enough time constant to hold between peaks and with a low enough impedance in the driving circuit to assure full charge of the capacitors in the time allowed by short peaks. The driving voltage must be amplified to a high enough level so that the diodes are essentially linear, otherwise, they will not act as an ideal switch and peak detector.

The details of the compressor circuit are too lengthy for a survey of this nature. Briefly, it is essentially composed of pentodes whose gain has been controlled by varying the supressor-grid voltage. Additional circuitry has been included in parallel with each tube to balance out the DC transients which could result from this type of control. This additional circuitry simply takes the form of another similar pentode with its plate drawing current from the first tube's screen supply and its screen drawing current from the first tube's plate supply. It is similarly driven by the DC control voltage, and tends to counteract the DC current charges in the compressor tube. This counteraction occurs because the cathode current of each tube varies very little. As the supressor voltage changes, all that occurs is a current flow which is diverted from plate to screen or vice versa.

The balance of the audio unit is fairly conventional broad-band audio circuitry of careful design.

Power Supplies

A generator of this nature requires DC power with little voltage fluctuation, low impedance, and good filtering. Two very similar series regulated power supplies were selected for this purpose. Two supplies are required because the current requirements are too much for a single 6080 tube to handle. Also, it is desirable to adjust one supply to match the compressor operating point without affecting the DC levels in the FM oscillator control unit. The only difference between the two power supplies is that the unit supplying the FM oscillator control circuit uses wire-wound resistors to immunize it against temperature changes. This supply also delivers power to the crystal oscillators.

Some of the more critical tubes in the FM control unit also have filament voltage stabilization. This is achieved by means of a thermal resistor (in series with these tubes) which essentially maintains a constant current in spite of normal line fluctuation and is independent of line frequency. A ferro-resonant transformer would regulate as well but would not have this immunity to line frequency.

In order to assure proper stability in all temperatures from −40°F to +137°F, several methods of temperature control are employed. First of all, the crystals are contained in a pair of ovens which maintain the internal temperature at 75°C ±1 degree. This is always above expected ambient. The entire RF tuning unit is contained in a large shield can which is heated by thermostatically controlled heaters. These heaters are turned on whenever the internal temperature falls below 50°C. A cooling fan is turned on whenever the temperature outside the can is above 15°C. Together these elements hold the temperature inside the can to fairly close limits (approximately 40° to 50°C).

Ruggedness

Great care has been taken to use materials and finishes which will withstand fungus and moisture. The instrument is not sprayed with moisture fungus varnish but rather uses materials in the construction which are nonfungus nutrient and moisture resistant.

Conclusion

All of the above-mentioned characteristics have been included in the Boonton Radio Generator Type 242-A. For those interested, advertised specifications for the production version of this instrument are available. It is our hope to explore some of the finer points of this device in further articles.

OWNERS OF TYPES 160-A AND 170-A Q METERS PLEASE NOTE

Radical changes in design changed considerably the parts make-up of Q Meters Type 160-A and Type 170-A after serial numbers 2000 and 700 respectively. Owners of instruments bearing these or lower serial numbers are advised that BRC's stock of spare parts for these instruments has been depleted and therefore all service of these instruments must be discontinued.

The RX Meter or the Q Meter?

NORMAN L. RIEMENSCHNEIDER, *Sales Engineer*

The question of which instrument, the RX Meter or the Q Meter, is best suited to a particular measurement problem frequently arises in our field work, since both instruments are designed for general impedance measurements. From the customers' point of view, this question is particularly timely since he must decide which instrument will yield the greatest utility for the expenditure involved. The purpose of this article is to set forth clearly the basic differences between the RX Meter and the Q Meter and further to establish how these differences affect measurements in areas of immediate interest.

In order to simplify our discussion of this rather broad subject, it might be well to consider only the basic ranges of the instruments involved, without attempting to include extended range measurements through the use of external accessories or special modifications. However, inasmuch as the Q Meter is basically designed to operate with a "work" coil, we shall allow the use of this particular accessory to lie within the boundaries of our approach.

Figure 1. The author loads one of BRC's station wagons before taking off on a field trip. This field service brings our full line of instruments to the customer's door. Actual demonstrations and technical consultation assist the customer with problems and in selecting the correct instrument for the job to be done.

Basic Differences

The fundamental differences between the RX Meter and the Q Meter may be briefly listed as follows:

1. The Q Meter provides a direct indication of Q over a range of 5 to 1200 or 10 to 625, depending upon the model used. Higher Q's, up to the order of several thousand, can be readily measured by indirect methods. The RX Meter is basically a low-Q device and will measure Q to zero. RX Meter readings are direct in equivalent parallel capacitance and resistance, and Q must be computed from these readings.
2. In the transition from direct to indirect measurements on the Q Meter, there is a gap in the range of resistances which can be measured. The RX Meter is on the other hand, calibrated directly in parallel resistance, and resistance measurements can be made in a continuous range of from 15 to 100,000 ohms. A comparison of the values of resistance that are measurable on the RX Meter and the Q Meter is shown in Figure 2.
3. The 260-A Q Meter operates in a range from 50 kc to 50 mc and the 190-A Q Meter operates in a range from 20 to 260 mc. Operating in a range from 500 kc to 250 mc, the RX Meter frequency range is almost equivalent to the total range of both Q Meters.
4. The resonating capacitance of the RX Meter goes down to zero, whereas the minimum capacitance of the 190-A Q Meter is 7 $\mu\mu f$ and the minimum capacitance on the 260-A Q Meter is 30 $\mu\mu f$. Coils requiring very low values of resonating capacitance can not be resonated on either of the Q Meters, but can be resonated on the RX Meter.
5. Voltage applied to the specimen in making measurements on the Q Meters is a function of the setting indicated by the XQ Meter, and is for the most part determined by the Q of the coil to be measured. The applied voltage of the RX Meter varies from 0.1 volt to 0.5 volts, but can, with the addition of an external potentiometer, be made to vary to a minimum of 0.02 volts. This feature of the RX Meter is particularly desirable when it is necessary to measure input circuits to tube grids or transistors, where the impedance can be influenced by the level at which it is measured.[1]
6. Since the insertion voltage of both types of Q Meters is developed between the low coil side and ground, neither side of the inductance is at ground RF potential. This, as a consequence, does not permit series type measurements to be made with one side at ground RF potential. The RX Meter is a grounded-type bridge and therefore measurements can be made with a common RF ground established between the bridge and the circuit under consideration.

Common Measurement Problems

It might be well at this point to consider how the fundamental differences between the Q Meter and RX Meter affect components and circuits under investigation. With these differences under consideration, let us list the several major measurement categories and determine which instrument is best suited for a particular measurement job.

Coils

As pointed out above the Q of high-Q coils can be best and most expeditiously

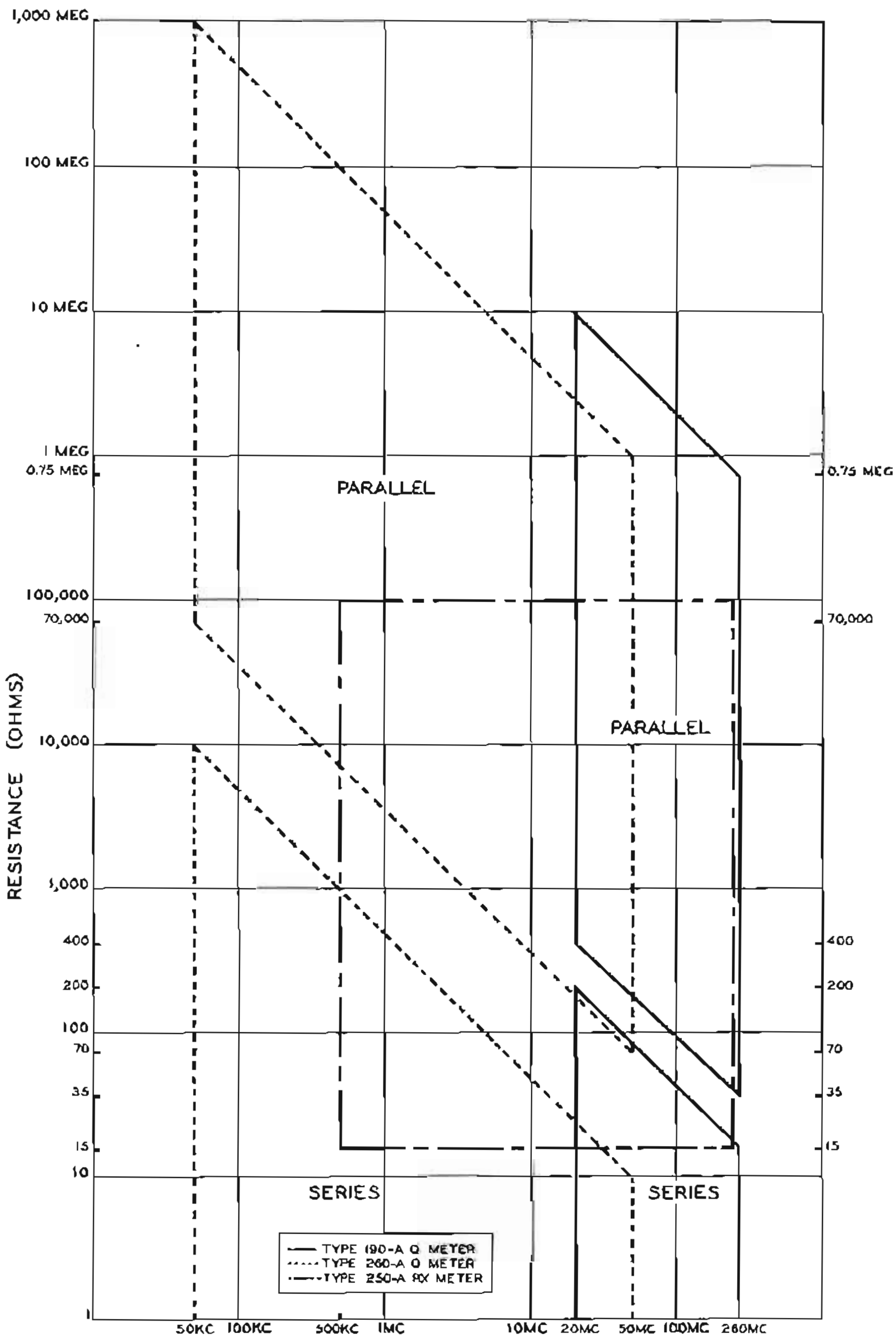


Figure 2. A comparison of measurable resistances on the Q Meter and RX Meter.

measured on the Q Meter. However, coils with very low Q's or coils with cores exhibiting extremely high losses can be measured more conveniently on the RX Meter because the latter instrument has the ability to measure Q's down to zero. (The permeability and losses of some core materials vary widely with dc and rf driving levels.)

As covered in a previous article[1] the RX Meter will measure the self-resonant frequency of coils, and distributed capacitance can very easily be determined from self-resonant frequency. This same measurement can be accomplished on a Q Meter; however, it is a somewhat more lengthy procedure to determine the self-resonant frequency on the Q Meter.

Capacitors

The capacitance of capacitors can be measured directly on the RX Meter within its range. However, unless the capacitor is extremely "lossy", its Q and/or dissipation factor can only be measured on the Q Meter because of the high value of parallel resistance involved.

Semiconductor Devices

Customers in the field advise BRC sales engineers that the RX Meter lends itself extremely well to semiconductor measurements because: (1) measurements can be made directly in one step, (2) the instrument can very easily be modified to operate at reduced level, and (3) the instrument lends itself to biased circuits.

Antennas

Either the Q Meter or the RX Meter can be used for antenna measurements, but the RX Meter lends itself somewhat more readily because it is an unbalanced, one side grounded, device. Use of the 515-A Coaxial adapter augments the convenience of the RX Meter in this application.[2]

Transmission Lines

While the characteristic impedance, attenuation, and velocity of propagation of transmission lines can be measured on the Q Meter, it can be accomplished more accurately and more expediently on the RX Meter. A description of the methods and techniques involved has been covered elsewhere.[3]

Vacuum Tubes

The RX Meter is better suited to vacuum tube measurements than the Q Meter for the same reasons outlined under semiconductor devices.

Conclusion

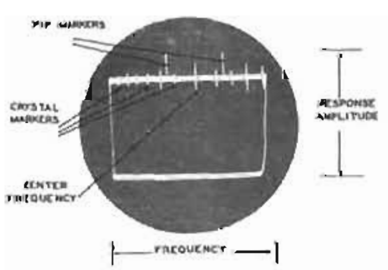
In this article we have attempted to make a comparison of the Q Meter and RX Meter, and to outline the areas of their application. It should be remembered, that with either type instrument, other measurements not mentioned can be made using suitable but perhaps more indirect techniques. Generally speaking, it can be said that measurements that can be handled on either instrument, with the exception of dielectric measurements and measurement of high-Q coils, can be performed more expediently on the RX Meter.

References

1. Riemenschneider, N. L., "Some VHF Bridge Applications," BRC Notebook No. 6, Summer 1955.
2. Gorss, C. G., "A Coaxial Adapter for the RX Meter," BRC Notebook No. 3, Fall 1954.
3. Riemenschneider, N. L., "Transmission Line Measurements with the RX Meter," BRC Notebook No. 3, Fall 1954.

Use Of Markers On Sweep Signal Generator Type 240-A

The Type 240-A Sweep Signal Generator is equipped with a built-in marker system which produces crystal-referenced "birdie-type" markers and adjustable pip interpolation markers. These markers are added to the response of the system under test, (see illustration below) producing a composite signal which may be displayed on an oscilloscope. Proper use of the Sweep Signal Generator, especially the marker system, will produce sharp, easy-to-interpret displays. Some helpful hints, aimed at providing optimum operation of the marker system, are contained in the following paragraphs.



Oscilloscope display showing markers added to response curve.

Test in Amplitude Control

The rectified response from the circuit under test by the Type 240-A Sweep Signal Generator is fed to the TEST SIGNAL IN jack, through the TEST IN AMPLITUDE control, and then to the Test Signal Amplifier where it is combined with the birdie markers to produce a composite signal output which may be displayed on an oscilloscope. When adjusting the vertical deflection of such a display, it will be found that best results are obtained when adjustment is made by means of the vertical deflection control on the oscilloscope, with the TEST IN AMPLITUDE control on the Sweep Signal Generator turned full on.

Verification of Center Frequency

Crystal markers, as they appear on the response curve of a circuit under test, have no significance unless a value can be assigned to the center marker. All that is known at this point in the measuring procedure, is that the markers are frequency spaced as indicated by the CRYSTAL MARKER SELECTOR. In order to assign frequency values to these markers, identification of the center marker must be made. When the center frequency marker is identified and its value ascertained by the CENTER FREQ control setting, it is a simple matter to determine the frequency values of other markers based on the frequency spacing setting of the CRYSTAL MARKER SELECTOR. Identification of the center frequency marker can be made as follows:

1. Set the CENTER FREQ control to the desired frequency.
2. Turn the CRYSTAL MARKER SELECTOR to the 2.5MC position and observe the markers on the oscilloscope display.
3. Narrow the sweep width by turning the SWEEP WIDTH control on the Sweep Signal Generator in a counterclockwise direction until only one birdie marker remains on the display. This marker is the center frequency marker and may be assigned the frequency value indicated on the CENTER FREQ dial.

After the center frequency marker has been identified, turn the SWEEP WIDTH control clockwise until the desired sweep width is obtained and all of the desired 2.5 MC markers are visible on the oscilloscope display. At the same time, note the position of the center marker with respect to the other markers. With the center frequency value assigned to this marker, it is now possible to determine frequency values of other points along the response curve, because the frequency spacing between the markers is known (in this case 2.5MC). If desired, .5MC and .1MC crystal markers may be introduced by repositioning the CRYSTAL MARKER SELECTOR. This will cause more closely spaced markers to appear on the display, and will permit further delineation of frequencies along the response curve.

Pip Markers

Once the frequencies along the response curve have been identified by means of the tuning dial on the Sweep Signal Generator and the crystal markers, the pip markers should be positioned to mark the desired frequencies, and the crystal markers should be turned off. This procedure provides a means for marking any two of the frequencies along the response curve, and at the same time, removes the possibility of interference with the curve caused by the crystal marker frequencies.

Attenuation of Crystal Markers

When using the Type 240-A Sweep Signal Generator to test a circuit which has a high-gain characteristic, it may be found that the crystal markers tend to become attenuated if the TEST IN AMPLITUDE control is in its full-on position and the vertical gain of the oscilloscope has been turned down to reduce the high level signal put out by the circuit under test. Since the amplitude of the markers is insufficient in the face of the high test circuit output, the input should be reduced or the "Test In" amplitude reduced to obtain the proper signal-to-marker ratio for satisfactory marker display.

EDITOR'S NOTE

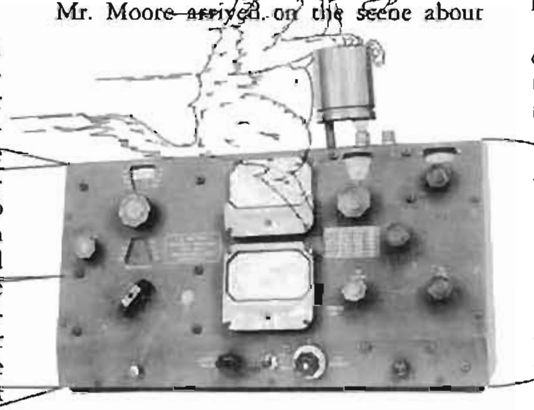
Q Meter Survives Re-Entry

The 8-mile drive west on Route 46, from Mountain Lakes to Dover, New Jersey is an exciting safari. Deluxe beaneries, vari-colored gas pumps, sign-infested discount stores, glowing billboards, and other scenic wonders, too numerous to mention, are breathtaking, even on a dull day. On December 2, during one of these treks from desk to dinner table, your editor was eyewitness to something he knows will be of extreme interest to Notebook readers, especially rocket enthusiasts and moonwatchers.

It was a little after 5 o'clock; the sun was not more than a few minutes below the horizon; and a group of other homeward-bound auto pilots and myself were jockeying for position for the 8-mile run to Dover. Suddenly a red-glowing object fell from the sky onto the roadway ahead, and jarred me from my hypnotic stupor. I was about third in line and had to lean heavily on the brake to avoid crashing into the automobile ahead of me. With the other drivers, I got out to investigate. The heat was so intense, we could not get within twenty feet of the object. After a few minutes however, the object no longered glowed, and appeared to be cooling rapidly.

The small crowd that had gathered by this time, edged closer to the object, until those of us in the lead were within 5 feet of it. It was then that I got the surprise of my life. The object in the road was unmistakably a BRC Type 260-A Q Meter. This didn't seem possible; after all, the object had fallen from the sky; but there it was.

When I was able to regain my composure, I dashed to a nearby public telephone, and phoned Mr. C. Moore, head of the BRC engineering department. Mr. Moore arrived on the scene about ten minutes later, attired in volunteer firemen's garb, and convinced the local police that the instrument should be transported to the BRC laboratory for examination.



The Type 160-A Q Meter shown soaring through space was actually received for repair after it had been damaged by fire. Inspection personnel at BRC were amazed to find the instrument in operating condition.

The instrument was examined with utmost care by Larry Cook, Quality Control Engineer, Bob Barth, Inspection Foreman, and other BRC engineers, and a full report was prepared and presented at a special board meeting held on Wednesday, December 4th. Copies of the report have been distributed to capital hill and interested government agencies. The contents remain confidential as this account goes to press, but I have been authorized by proper authority to disclose two unusual facts: the 260-A Q Meter was tested and found to be in good working order, (incredible as this may seem); the examiners noted what appeared to be the paw marks of a dog on the instrument's front panel.

There are some Notebook readers, I'm sure, who are writing this off as just another flying-saucer story; but let me assure these skeptics that this account is factual, and will be substantiated just as soon as the security ban is lifted. In the meantime, we have been authorized to publish the photograph shown on the left. The Q Meter shown is the object that fell from the sky. Note the damaged front panel. Naturally, for the sake of security, the paw marks have been cleverly removed by a retouch artist.

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NUMBER 17

The
NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

A Q Comparator

JAMES E. WACHTER, *Project Engineer*

It would be difficult to imagine that any industry has expanded more than the electronic industry during the past decade. The production of television receivers, electronic control devices, computors, and many other electronic devices has soared during this time. With this increased production of electronic devices came ever increasing demands for enormous quantities of electronic components including coils, capacitors, resistors, and these components wired in combination. This brought about the inevitable need for faster methods of producing these components.

Developers went to work on the problem and today the fruits of their labor are evident in the numerous automatic production machines which are standard equipment in the plants of many of the electronic component manufacturers. These same manufacturers are producing thousands of coils, capacitors, and resistors daily. Productionwise the challenge has been met.

Reports from some of the electronic components manufacturers indicate, however, that not all is well. It appears that, in some areas, instrumentation for checking these components against manufacturer's tolerances has not kept pace. Today, manufacturers are interested in getting trend and reject information from the inspection operation back to the production operation as quickly as possible to keep production within design tolerances and to keep rejects to a minimum. This cannot be done with available instruments.

This article describes and delves somewhat into the design principles of an instrument which will meet present day requirements for a fast and accurate means for determining relative Q, inductance, capacitance, and resistance. The instrument, BRC's Type 265-A Q Comparator, is now in production.

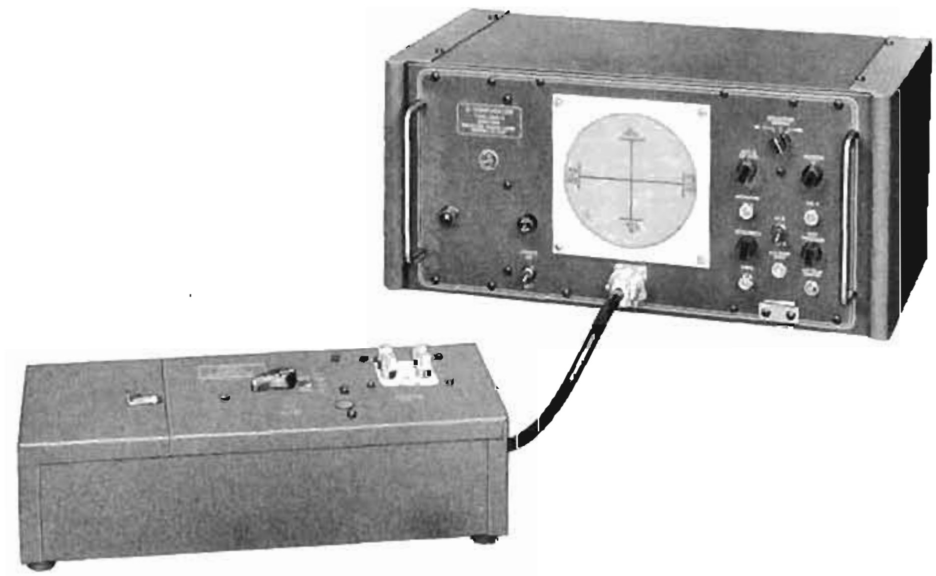

Figure 1. Two-unit design of the Type 265-A Q Comparator puts the test circuit portion of the instrument at the front of the test position where it is needed, and the indicator portion at the rear of the test position, out of the way but within easy view.

GENERAL DESCRIPTION

The Q Comparator is comprised essentially of a swept-frequency oscillator, Q meter-type measuring circuit with detector, vertical amplifier, differentiator, spot generator, horizontal amplifier with blanking circuit, cathode-ray tube, and power supply, (Figure 2). The swept-frequency oscillator and measuring circuit are assembled into a relatively small unit which is cable-and-plug connected to the main unit containing the cathode-ray tube and remaining circuitry. With this arrangement, the portion of the instrument on which the test connections are made can be used at the front of an inspection bench, taking up no more space than most inspection gauges or fixtures, while the main indicator portion of the instrument is placed at the rear of the bench or rack mounted off the bench altogether, out of the working area, (See Figure 1).

The initial set up of the Q Comparator is performed by a test engineer or other suitably skilled person, using a standard component having the characteristics that are desired in the production components. The set-up procedure, which will be explained later, results in a dot at the center of the CRT when the standard component is connected to the Q Comparator test circuit. Following this set up, comparatively unskilled personnel can rapidly check production components by simply connecting them to the test circuit and observing the position of the dot on the CRT. These are no tuning operations to be performed and no meter readings to be evaluated.

Any dot which does not appear at

YOU WILL FIND . . .

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the center of the CRT indicates that the component under test is different than the standard. Deviation along the vertical axis shows a change in Q and deviation along the horizontal axis shows a change in L (inductance) or C (capacitance). It becomes apparent that any desired limit conditions could be grease penciled on the CRT, and any dot falling outside these boundaries would then indicate a reject component. In addition to performing this "go — no-go" function, the instrument will supply trend information quickly so that in many cases production can be altered before rejects occur.

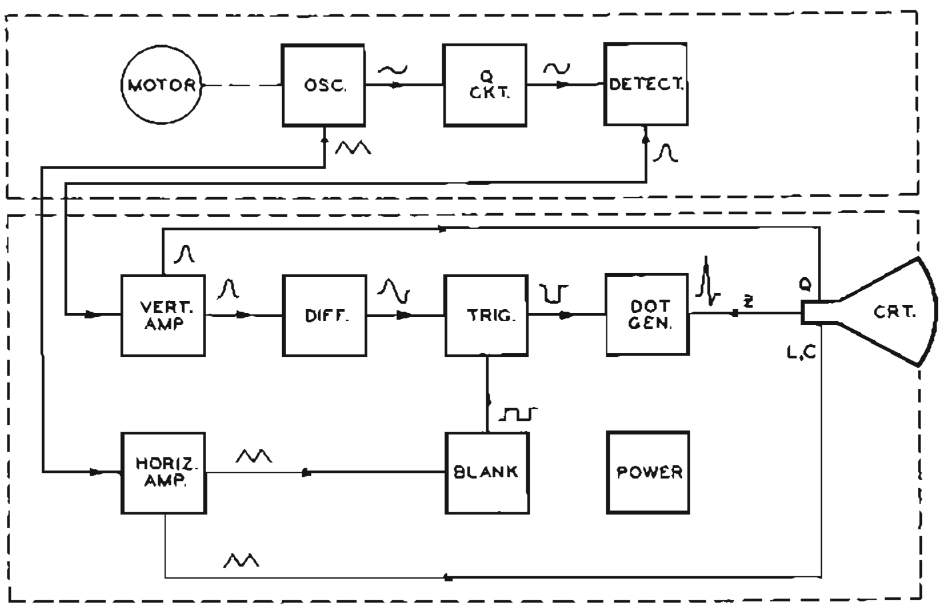

Figure 2. Q Comparator Block Diagram

SET-UP PROCEDURE

The standard component is connected to the Q Comparator test circuit, which is a basic Q-meter resonant circuit containing an injection impedance and resonating capacitor. With the oscillator tuning capacitor motor turned off, the capacitor is set to its center frequency position by means of a detent in its shaft. An oscillator coil covering the desired test frequency is selected and plugged into the oscillator circuit and tuned to the desired frequency by means of its calibrated adjustment screw. The capacitor detent is released and the motor actuated. By means of a switch, either of two sweep widths may be selected; ±25% or ±5% of the center capacitance in the oscillator circuit. It is simpler to use the wider sweep during the set up and then reduce it later if desired.

The adjustments made thus far are performed on the test circuit unit. The remaining adjustments concern the CRT display. The intensity is turned up and a fixed dc voltage is applied to the vertical deflection amplifiers by means of a function selector switch. Under this condition, a horizontal trace appears on the CRT. The trace is vertically and horizontally centered on the CRT by adjusting the vertical and horizontal deflection voltages respectively. The length of the trace, corresponding to the sweep width, is set to the desired size by varying the gain of the horizontal amplifier. The Q calibration of the CRT is made by varying the gain of the vertical amplifier. With the foregoing procedures completed, the function selector switch is put in the test position and the resonating capacitor (in the test circuit) is tuned for resonance.

Either of two resonating capacitors may be used, depending upon whether a high or low value of capacitance is required. Resonance will be indicated on the CRT by the appearance of a resonance curve or a portion of a resonance curve. The RF level of the oscillator is adjusted so as to raise or lower the peak of the resonance curve to the vertical center of the CRT. The resonating capacitor is finely tuned to locate the peak at the horizontal center of the CRT.

Adjustment of the "Dot Position" control will cause an intensified dot to appear on the resonance curve and to be moved across the peak of the curve. The dot is positioned at the highest point of the resonance curve; i.e., the center of the CRT (Figure 3). A "Hi Q - Low Q" switch permits the "Dot Position" control to function properly under either high or low-Q conditions. With the dot in position, the intensity of the display is reduced with the "Intensity" control until the resonance curve disappears and only the dot remains. The "Astigmatism" and "Focus" controls are adjusted for the sharpest and most symmetrical dot. The Q Comparator is now ready for use.

DESIGN TECHNIQUES

Oscillator and Sweep Generator

The oscillator is a simple one operating in the 200-kc to 70-mc range and consisting of a cross-connected 12AU7 double triode. Electrically the oscillator section is quite conventional. The tuning arrangements of the oscillator on the other hand are somewhat unconventional.

Tuning for center frequency is accomplished by means of variable inductances in the oscillator tank circuit, while the variable (swept) tuning is accomplished by a motor-driven capacitor. With this system, the ratio of the capacitance variation to the total average capacitance is independent of the coils used, and the same relative reactance variation can be obtained for any center frequency. The output of the oscillator is coupled via a loop to the injection impedance of the test circuit detector. Amplitude is controlled by varying the plate voltage of the oscillator tube.

Simultaneously with providing a swept frequency, one section of the motor-driven capacitor serves to generate a sawtooth voltage for the horizontal sweep (corresponding to the X axis on the cathode-ray tube) which is always in synchronism with the instantaneous frequency deviation.

Detector

In addition to the detector itself, the detector section contains the test circuit which is a Q-meter arrangement with a two-section tuning capacitor for high and low frequencies. The detector proper is an "infinite impedance" type (cathode follower). A 12AU7 double triode is used for this purpose with one of the tube sections being used for compensation of heater voltage changes.

Vertical Amplifier

The vertical amplifier consists of two cascaded differential dc amplifiers. DC amplification is required in order to keep the base line (corresponding to a no-input signal) in place so that the height of the resonance curve can be measured from this base for any position or width of the resonance peak. The differential amplifiers provide a simple means for balancing level variations caused by heater voltage changes, and depressing the base line so that only the peaks of the resonance curves show on the CRT screen.

A voltage divider stabilized with a neon lamp provides a means for setting up and checking the vertical amplification at the input of the vertical amplifier. The amplification of the vertical channel can be adjusted by varying the amount of negative feedback in the first stage.

The vertical amplifier also delivers the input signal to the differentiator.

Differentiator and Trigger

The differentiator-trigger circuit, consisting of a feedback-pentode differentiator between two cathode followers, provides a trigger pulse for the dot generator at the precise moment of the resonance curve peak. The differentiator produces first and second derivatives of the resonance curve which, in turn, are used to anticipate the peak of the resonance curve to allow for circuit delays.

In addition to acting as a source of feedback voltage to the differentiator proper, the second cathode follower drives the input amplifier to the trigger circuit. A minimum number of coupling capacitors have been used in the circuit in order to keep down any charging effects.

The trigger circuit comprises two cascaded modified "Schmitt" triggers and are essentially overdriven amplifiers delivering a well defined voltage step of constant amplitude at the moment the input voltage reaches a predetermined level.

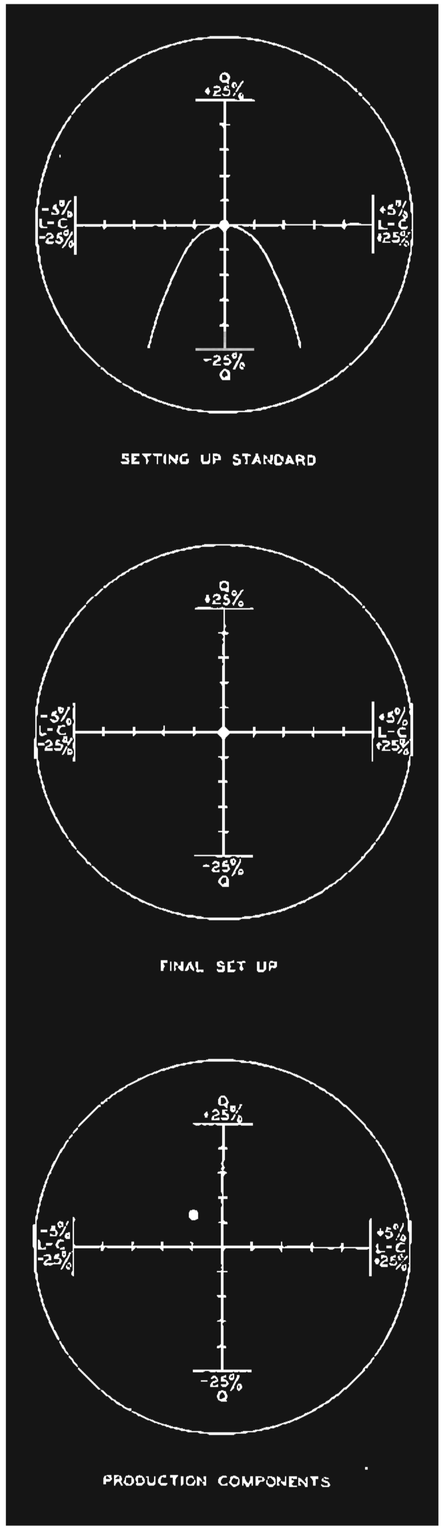

Figure 3. Q Comparator CRT Displays

Spot Generator

The step voltage from the trigger is differentiated and serves to initiate the final brightening pulse. This pulse is generated in a monostable multivibrator. In order to obtain a pulse of high amplitude and very short duration without excessive power consumption, high value load resistors are used and feedback and coupling capacitors of the multivibrator consist of the distributed and tube capacitances.

Horizontal Amplifier

A small sawtooth voltage is obtained from an RC network containing a rotating capacitor and amplified to a sufficient level to cover the entire CRT screen. Filters isolate the sawtooth generation from the oscillator. The first two stages of this amplifier are conventional RC-coupled triodes, with time constants that keep the sawtooth from deteriorating in shape. The final stage is a differential amplifier which provides a push-pull output and an easy means for centering the trace on the CRT screen.

One of the sawtooth outputs is differentiated into a square wave and used for actuating the blanking circuit. Blanking is necessary because of the slight nonlinearity of the sweep voltage which produces two resonance peaks side-by-side on the CRT screen. Eliminating the brightened dot from one of these results in the display becoming single-valued.

Cathode-Ray Tube

A cathode-ray tube is used as the display element, with the X axis horizontal and the Q axis vertical. The type 5ABP1, a 5-inch post-deflection accelerator tube, was chosen for this purpose because it requires the simplest amplifying and power supply circuitry.

The tube is operated with its cathode at —1700 volts, the deflection system at a few hundred volts positive, and the post accelerator at +1700 volts. An astigmatism control is required in addition to the usual focus and brightness controls, in order to provide a brightened dot that is reasonably small (about 1 mm) and circular. The brightness of the dot is considerably higher than would be used in a line display.

A magnetic shield isolates the CRT from the transformers and chokes in the power supply and from the driving motor in the oscillator and sweep generator sections.

Power Supply

The power supply supplies all of the voltages required by the Q Comparator. Certain of these voltage lines are stabilized with voltage regulator tubes because they feed stages sensitive to dc level changes. In some instances tubes will occasionally have their cathodes operating at close to +150 volts or −150 volts. In these cases only tubes with high permissible heater-cathode voltage ratings have been employed. This eliminates the need for separate filament windings.

Because of the number of tubes used in the instrument, some drawing relatively heavy currents during short intervals (pulses), thorough filtering is employed in the power supply in addition to sectionwise decoupling.

CONCLUSION

In the Type 265-A Q Comparator, BRC believes it has developed an instrument which will be invaluable to manufacturers of electronic components. The comparator provides comparison measurements of relative Q, inductance, and capacitance easily and accurately. Since these measurements require merely the visual inspection of a dot on a cathode-ray tube screen, the instrument can be operated by unskilled personnel once it is set up.

BRC customers interested in the Q Comparator may obtain information from one of the BRC representatives listed on page 8 of the Notebook. Also, the instrument will be on display in the BRC booth at the IRE show in New York City during March 24 through 27.

Remote Measurements with the RX Meter Employing Half-Wavelength Lines

ROBERT POIRIER, *Development Engineer*

Previous Notebook articles, Fall 1954, issue number 3, page 7 and Summer 1956 issue number 10, page 5 broached the subject of measuring impedances which are necessarily located some distance from an RX Meter but connected to the RX Meter through a constant impedance transmission line. The procedures to be described in the present article while not necessarily limited to constant impedance transmission line circuits will be restricted thereto because of the great simplification in data processing which results from the use of constant impedance transmission lines. It is the purpose of the present article to carefully consider the use of half-wavelength transmission lines of constant characteristic impedance, Z_o for remote impedance measurements with the Type 250-A RX Meter. The viewpoint is toward establishing simple procedures for interpreting the RX Meter readings through either lossless or lossy transmission lines to an unknown terminal impedance via the Smith Chart.

Transmission Line Equations

The equation which relates input impedance, Z_1 to the characteristic impedance Z_o of a transmission line and the load resistance Z_l is given for the general case by

$$Z_1 = Z_o \frac{Z_l \cosh \gamma l + Z_o \sinh \gamma l}{Z_o \cosh \gamma l + Z_l \sinh \gamma l} \qquad (1)$$

where the complex propagation constant

$$\gamma = \sqrt{(R + j\omega L)\ (G + j\omega C)}, \qquad (2)$$

for the distributed series resistance, R; inductance, L; shunt conductance, G; and capacitance, C per unit length.

Some special cases are considered as follows:

For $Z_l = Z_o$

$$Z_1 = Z_o$$

for any distance l between the load impedance and the measuring point.

$$\text{When } Z_l = \infty \quad \left(\frac{Z_o}{Z_l} = 0\right)$$

as in the case of an open circuit terminal

$$Z_p = Z_o \frac{\cosh \gamma l}{\sinh \gamma l} = \frac{Z_o}{\tanh \gamma l} \qquad (3)$$

and when $Z_l = 0$ as in the case of a short circuit terminal

$$Z_s = Z_o \tanh \gamma l \qquad (4)$$

By manipulating equations (3) and (4) we have

$$Z_o = \sqrt{Z_p Z_s} \qquad (5)$$

and

$$\tanh \gamma l = \sqrt{\frac{Z_s}{Z_p}} \qquad (6)$$

In this article we are primarily concerned with half-wavelength lines for which Z_p and Z_s are pure resistances, R_p and R_s respectively. Also from (5) and (6)

$$\tanh \gamma l = \frac{Z_o}{R_p} \qquad (7)$$

and separating γ into $\alpha + j\beta$ we have for ½ wavelength lines

$$\tanh 2\alpha l = \frac{2 R_p Z_o}{R_p{}^2 + Z_o{}^2} \qquad (8)$$

and

$$\tan 2\beta l = 0 \qquad (9)$$

Equation (8) may be written as

$$\alpha l = \tfrac{1}{2} \tanh^{-1} \frac{2 R_p Z_o}{R_p{}^2 + Z_o{}^2}$$

which for $R_p \gtrsim 5Z_o$ reduces to

$$\alpha l \approx \frac{Z_o}{R_p} \text{ neper} \quad \text{or} \quad 8.69 \frac{Z_o}{R_p} \text{ db}$$

Equation (8) relates the input resistance, R_p measured at integral half-wavelength distances from the open or short circuited end of a Z_o characteristic impedance transmission line to the total transmission loss αl of the length, l transmission line. We note that for lossless lines ($\alpha l = 0$) R_p may be either 0 or ∞ but for transmission loss in excess of 20db R_p approaches Z_o for either open- or short circuited terminals or any termination in between open or short. Equation (8) is plotted in Figure 1 for the open and short circuit limits and from these a family of curves is generated to show the effect of transmission loss in db on the measured resistance of various resistive terminations separated from the measuring point by integral half wavelengths of transmission line. The characteristic impedance, Z_o of the transmission line has been assumed 50 ohms for this illustration.

The Smith Chart

The Smith Chart is a family of solutions contained within a unit circle for the transmission line equations in terms of the series orthogonal components and may be interpreted for $R + jX$ of impedance or $G + j\beta$ of admittance. The polar coordinates of the unit circle are reflection coefficient, ρ and line length, βl in terms of wavelength. The maximum radius of the unit circle therefore corresponds to a reflection coefficient of 1 and the center to a reflection coefficient of 0. This is related to VSWR as

$$\text{VSWR} = \frac{1+\rho}{1-\rho}$$

So far as the series components of impedance or admittance are concerned, one-half wavelength of βl corresponds to a complete cycle and to 360° or one full turn around the chart. Smith Charts are available with the orthogonal coordinates calibrated in resistance and reactance for 50 ohm characteristic impedance, conductance and susceptance for 20 millimho characteristic admittance, or normalized impedance or admittance coordinates;

$$\frac{R}{Z_o}, \pm\frac{jX}{Z_o}, \frac{G}{Y_o}, \text{ and } \pm\frac{jB}{Y_o}$$

Usually included on the Smith Charts are the radially scaled parameters; reflection coefficient and loss, standing wave ratio and transmission loss.

For the purpose of illustrating the effect of transmission loss on Smith Chart plots, Figure 2 shows a normalized coordinate Smith Chart with 1db steps of transmission loss drawn as concentric circles. The intersection of these circles with the resistance/conductance axis corresponds, in Figure 1, to the intersections of the open and short circuit curves with whole integers of db and shows (as in Figure 1) the effect of transmission loss on the measured resistance of open or short circuit terminals at integral half-wavelength intervals in 1db steps.

Consider for example an open circuit termination. This corresponds in impedance coordinates to the extreme right end of the resistance axis and in admittance coordinates to the extreme left end of the admittance axis. An integral number of half wavelengths removed from the termination along a transmission line corresponds in the lossless case to the same integral number of clockwise revolutions around the Smith Chart, at constant radius, where the open circuit termination still appears as an open circuit. But for an integral number of half wavelengths and with 1db removed from the termination, the impedance or admittance locus spirals clockwise toward the center of the chart for the number of complete revolutions corresponding the number of half-wavelength line lengths to the circle representing 1 db transmission loss. In this case, we read from Figure 2 for the open circuit termination; $R = 8.7Z_o$ ohms (435 ohms for $Z_o = 50$ ohms) or $G = 0.115Y_o$ millimhos, or from Figure 1; $R = 435$ ohms.

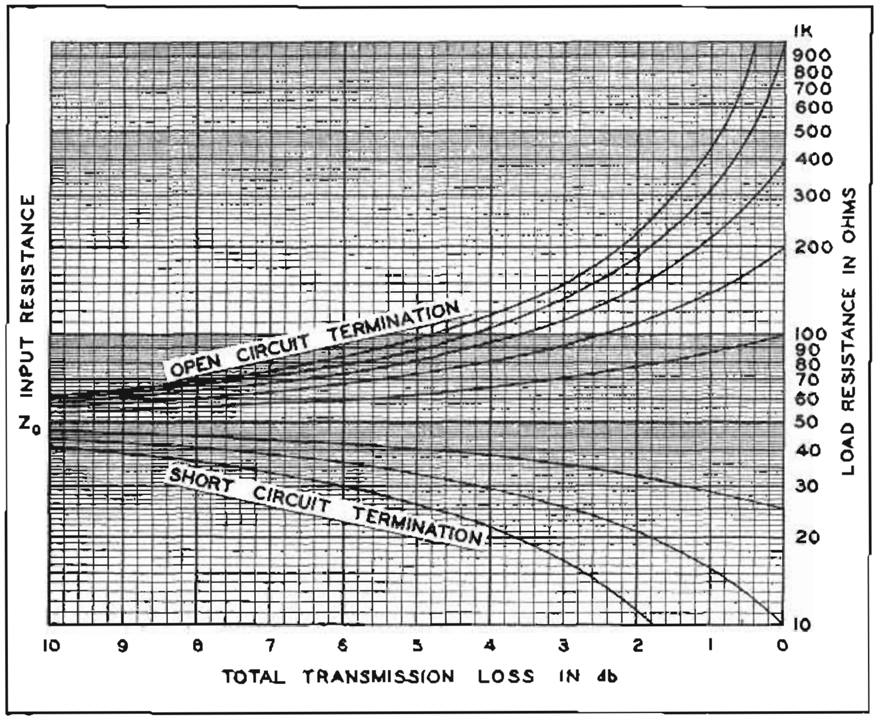

Figure 1. Apparent change of load resistance versus line loss in db when viewed from positions located on integral number of half-wavelengths from the termination.

Measurement Procedure

The basic operations associated with remote impedance measurements employing the RX Meter are outlined as follows:

1. Establish a length of uniform characteristic impedance transmission line between the RX Meter and the measuring point to be exactly an integral number of half wavelengths by providing an open circuit termination exactly at the location where the impedance to be measured will be located. Adjust the line length and/or the measuring frequency so that the RX Meter, appropriately tuned and balanced, reads a resistance; $R_p > Z_o$ and $C_p = 0$. For 50-ohm coax it is expedient to use the Coax Adapter Kit Type 515-A for connecting to the RX Meter. Record R_p. The effective parallel resistance thus obtained for an open-circuited line may be located on Figure 1 or a Smith Chart to determine the total transmission loss between the RX Meter and the open circuit. Record the total transmission loss. As an example, if the RX Meter reads $R_p = 1{,}000$ ohms $C_p = 0$ from Figure 1 or Figure 2, the total transmission loss is 0.4db.

In setting up the proper length transmission line it is preferable to avoid the use of a multiplicity of patch cords and/or constant impedance adjustable lines which are not exactly constant impedance. The reason for this is that adjustable lines and cable connectors (the latter especially when carelessly assembled and/or not kept clear of grease, dirt and the little brass chips and flakes of silver plate which tend to accumulate) produce discontinuities which cause changes in VSWR independently of the VSWR changes due to transmission loss. The operator in this case is faced with the choices of attempting to account for all discontinuities, eliminating them, or neglecting them.

2. Connect the impedance to be measured at the end of the transmission line in place of the open circuit. Rebalance the RX Meter and read and record R_p and C_p. Since the RX Meter

reads out parallel components it is convenient to convert the reading to normalized admittance coordinates,

$$G = \frac{1000}{Y_o R_p} \text{ and } B = \frac{j1000\omega C_p}{Y_o}$$

per unit millimho and locate the coordinates on a Smith Chart.

3. Consider a radius line drawn from the center of the Smith Chart through the located admittance. Since the transmission line has been established exactly an integral number of half wavelengths, the actual admittance at the end of line must be somewhere on this radius, there being no net change in βl for integral numbers of half wavelengths. For lossless or reflectionless transmission the admittance at the end of the line is the same as measured at the input end of the line. For lossy transmission with reflection the VSWR on the line will be increasing toward the load. The actual load admittance may be evaluated by adding the total transmission loss, previously determined, to the measured admittance along the radius vector away from the center of the Smith Chart. That is to say, the total transmission loss in db is added along the radial transmission loss scale in db away from the center of the Smith Chart, which is the direction for approaching the load to locate the point on the chart denoting the admittance (or impedance) of the load proper.

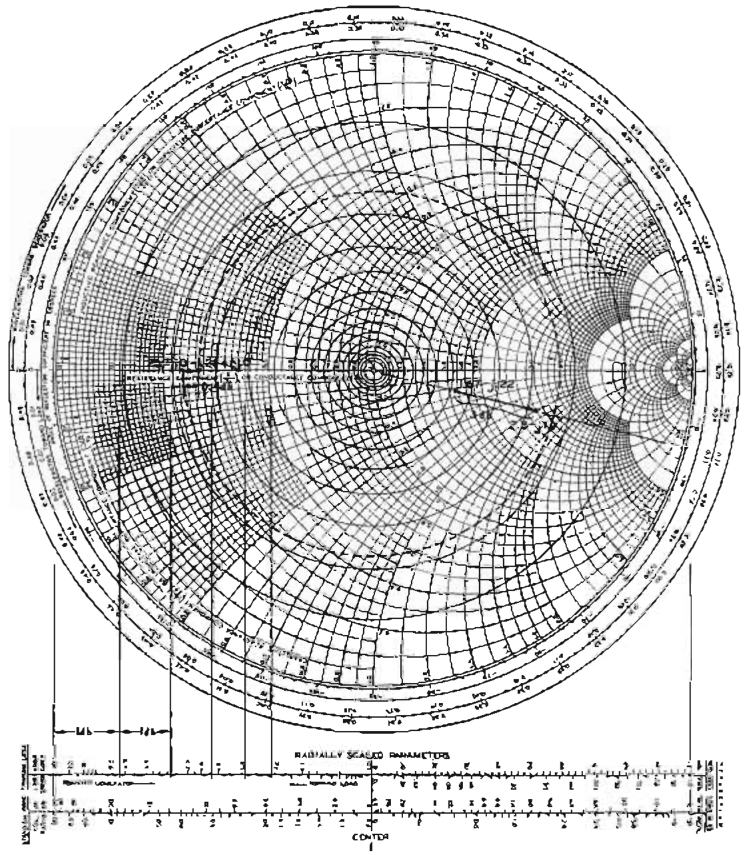

Figure 2. Smith Chart illustrating transmission loss.

Examples

To illustrate step 3, above, let us suppose that the transmission loss of a given cable of one-half-wavelength 50-ohms characteristic impedance has been found to be 0.4db according to step 1, and that an RX Meter has indicated $R_p = 170$ ohms $C_p = 0$ for the termination and line. This may be plotted directly either as normalized conductance or normalized resistance since the series components are the same as the parallel components for pure resistance. On Figure 2 this data is plotted as conductance

$$G = \frac{1000}{20 \times 170} = 0.294$$

per unit millimho. Scaling 0.4db along the conductance axis away from the center from $0.294 \pm j0$ we find $0.25 \pm j0$ per unit millimho or 200 ohms for the termination. This result can also be obtained from Figure 1 which is plotted for pure resistance terminations. The entire admittance locus for any point on the transmission line is shown on Figure 2 as a dashed line spiral of one revolution for the one half wavelength total line length.

A second example is considered as follows:

In this case, step 1 for 50-ohm coax reveals $R_p = 150$ ohms $C_p = 0$. From Figure 1 or Figure 2 the total transmission loss is 3 db. Step 2 reveals $R_p = 30$ ohms $C_p - 7.0\ \mu\mu f$ at 100 mc. Converting to normalized admittance coordinates

$$G = \frac{50}{30} = 1.67 \text{ and } B = -j\,50\,\omega$$

$C_p = -j0.22$. This point is located on the chart and a radial line through this point extended 3db away from center locates $2.8 - j1.0$ per unit admittance at the termination or

$$R_p = \frac{50}{2.8} = 17.9 \text{ ohms}$$

and $C_p = -31.8\mu\mu f$. For the purpose of simplifying the figure, the spiral admittance locus, mainly a matter of academic interest, was omitted in this example. Moreover, to plot the locus it is necessary to know the length of the line in wavelengths which was not given as data.

Some Aspects of Balanced Line Measurements

The foregoing procedures are essentially applicable for either unbalanced coaxial transmission line or dual balanced transmission lines such as "twin lead"; directly so when the balanced to unbalanced transition known as a balun is omitted and the two wires of the balanced line are connected directly to the Hi and Lo binding posts of the RX Meter. This procedure is subject to error resulting from unbalanced capacitance effects at the binding posts which may be minimized by leading the balanced line vertically away from the bridge. If a balun is employed a multiplying factor of 4 is required in going from unbalanced to balanced impedance and from balanced to unbalanced admittance. If the transmission line includes both balanced and unbalanced sections the balanced and unbalanced sections may be any length each, provided that the total transmission length is an integral number of half wavelengths and that the characteristic impedance of the balanced section is exactly 4 times the characteristic impedance of the unbalanced section. The total transmission loss is evaluated, as previously, by measuring R_p, for the open circuited integral half-wavelength transmission line, locating R_p on the Smith Chart, and measuring attenuation on the db scale between $R = \infty$ and R_p. If the balanced to unbalanced characteristic impedance ratio is not exactly 4:1 then the two sections must be each adjusted to exactly an integral number of half wavelengths and the total transmission loss may be evaluated as previously.

Generally, measurements made with balanced lines are more subject to error than unbalanced coaxial. Primarily these errors result from discontinuities in Z_o caused by unmatched cable connectors and the close proximity of the balanced sections to conducting objects. Here again, the choices are to account for all discontinuities, eliminate them or neglect them.

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New Wings for BRC

EDSON W. BEATTY, *BRC Pilot*

Over six years ago, BRC entered the business aviation field with the purchase of a Beechcraft Bonanza Aircraft for executive transportation. Recently we replaced this single-engine plane with a multi-engine Cessna 310, shown in the photograph. Because BRC is located on the fringe of Metropolitan New York, an area offering three major airports and fourteen domestic airlines, one would presume that we should have very little use for a business aircraft. Quite the opposite is true: we find the airplane to be extremely useful for handling our business transportation problems.

BRC's new Cessna 310 shown at the Morristown Municipal airport where it is based.

There are many advantages of owning and operating our own business aircraft; the most apparent being the time saved and the fact that we are not bound by firm schedules. For example, in order to travel via commercial airlines, an executive must depart from his office at least 2 hours before the scheduled departure time. Using the company aircraft, he can be in the air and on his way in 45 minutes. Since he sets his own departure time, he can tend to those "last minute details" before leaving. Also, with this type of operation, we are not limited to airline-served cities, but may use the facilities of most of the military and civil airports located throughout the country.

Our new Cessna 310 may be briefly described as a multi-engine, five-place, low-wing monoplane with retractable tricycle landing gear. At a distance in flight it resembles a jet aircraft. This allusion probably stems from the large wing-tip fuel tanks (each holds 50 gallons of fuel). The engines are a reciprocating type, consisting of two six-cylinder horizontally-opposed engines developing 240 HP each. The fuel tanks, together with the relatively short 25-foot length over-all and 36-foot wing span add greatly to the inflight stability of the aircraft. The 310 is aerodynamically clean in design, offering very low drag resistance. This reduction in drag resistance is accomplished mostly by good engine and cowling design. Each powerplant is installed in a 21-inch deep cowling which is an airfoil section contributing lift to the wing area. The powerplant accessories are completely duplicated on each engine so that in the event of a complete engine failure in either unit the generator output of only one is lost. The output of the other generator is more than ample to operate the electrical system and accessories of the plane.

At full load, the 310 will cruise at a speed of 205 MPH and climb at a rate of 1700 feet per minute near sea-level. It has a range of 875 miles and a service ceiling of 20,000 feet. The gross weight is 4600 pounds and the useful load is 1750 pounds.

Navigation and communication are no problem to the pilot because the BRC plane is amply equipped with the following impressive list of navigation and communication equipment.

ADF Receiver
VHF Navigation Receiver
VHF Communications Transceiver
VHF Standby Transmitter
ILS Glide Slope Receiver
Marker Beacon Receiver
Audio Amplifier

If we seem as enthusiastic about our airplane as a child with a new toy, it is because the aircraft has become an important part of our business operation. A recent business trip made by company executives typifies the ground that can be covered with this type of travel. The BRC group enplaned for a mid-morning departure from Morristown, New Jersey and arrived for a luncheon business meeting in Cleveland. They departed in mid-afternoon, arriving in Chicago late that same afternoon for meetings and dinner. After dinner, they left for Omaha, where they remained overnight. Departing the next morning, they arrived in Denver in time for noon appointments. We doubt that such a schedule could be maintained utilizing public transportation facilities.

ANOTHER Q METER CONTEST

In 1955, and again in 1957, BRC awarded Type 160-A Q Meters to lucky persons who visited the BRC booth at the IRE shows held in New York City in March of those years. The first award was determined by means of a drawing of registration cards filled out by guests at the BRC booth. In 1957 a new twist was added when guests were invited to estimate the Q of a specially wound coil exhibited at the booth, the person guessing closest to the actual measured Q being awarded the Q Meter.

The latter contest stimulated so much interest at the show last year that BRC has decided to sponsor a similar contest this year. A factory reconditioned 160-A Q Meter will be awarded again to the person whose Q estimate is closest to the measured Q of the "mystery" coil. Be assured that the coil will be as weird in shape and dimension as the fiendish minds of the BRC engineers can make it. See the coil, together with the Q Meter, at the IRE show in New

York City during March 24 through 27 and judge for yourself.

Contest rules will be the same as they were last year. The person who submits the Q estimate which is the closest to the actual Q of the coil, as measured by the BRC quality control engineer in the company standards laboratory, will be declared the contest winner. In case of a tie, a drawing will be made to decide the winner.

If you have some measuring problem, drop around to booths 3101 and 3102 at the show, we look forward to serving you. Maybe you will win a Q Meter for your trouble.

Norman L. Riemenschneider

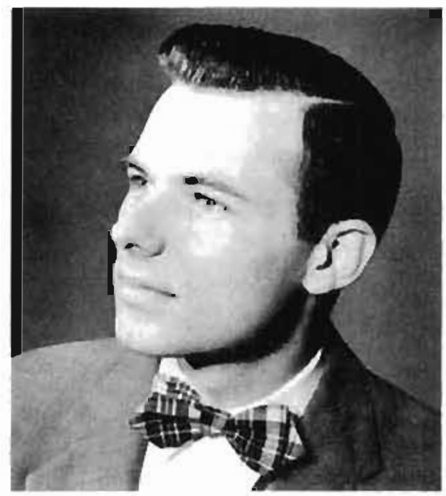
BRC FIELD ENGINEERING STAFF

Pacing the growth of the electronic industry in the various geographical areas of the Country, we have endeavored to provide a staff of capable, well-trained field engineers to serve our customers in each territory. Below you will find a complete listing of our local field offices where, through a simple local telephone call, complete engineering assistance on the application of our instruments may be obtained. In addition, we maintain a field engineering group at our factory in Boonton, New Jersey to serve the local areas and also to assist our representatives in special applications. Our present home office staff includes Norman L. Riemenschneider and George P. McCasland, and we thought we would take this opportunity to introduce these engineers to you.

"Norm" joined BRC in 1953 as Sales Engineer and has concentrated on serving customers in the Metropolitan New York and Philadelphia areas. He received his B.S. in Electrical Engineering from Newark College of Engineering in 1943 and, drawing upon his 17 years of experience in the electronic industry, augmented by amateur radio experience dating back to 1938, has developed a wealth of valuable application information that has continually saved countless time and effort for our many customers.

"George" joined BRC just this past January and promises to be a welcome addition to our staff. He received his B.S. in Electrical Engineering from the University of Virginia in 1952 and his M.S. in Industrial Management from M.I.T. in 1954. With 3 years of experience in electronic systems development and amateur radio experience dating back to 1947, he is well qualified to assist our customers in the solution of their measurement problems. George will be responsible for serving our customers in the Metropolitan Baltimore and Washington areas.

If you're attending the I.R.E. Show, why not drop by at Booth Numbers 3101-2 and let Norm and George know of your instrumentation problems? A telephone call to our plant, at any time, will place these engineers at your service.



George P. McCasland

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

A Control System for an FM Signal Generator

CHARLES G. GORSS, *Development Engineer*

The oscillator in an accurate fm system has two rather incompatable requirements impressed upon it. First, it must be stable: its frequency must not change so much as 0.001% under the influence of environment, input voltage changes, or prolonged usage. On the other hand, it must respond cheerfully and faithfully to a command to deviate from its assigned frequency according to an applied modulating wave and return instantly to its preassigned frequency when the impressed voltage is removed. Of course, these demands upon an oscillator in an fm system are not impossible to meet, nor are they new. FM transmitters, employing many different techniques, have fulfilled these requirements for many years. Some of these methods are much too complicated and inflexible for incorporation into a portable signal generator, and therefore have been discounted for this discussion. Instead, a system will be presented which has been developed specifically for use in a crystal controlled signal generator (Figure 1), but whose accuracy features make it equally suitable for use in a transmitter.

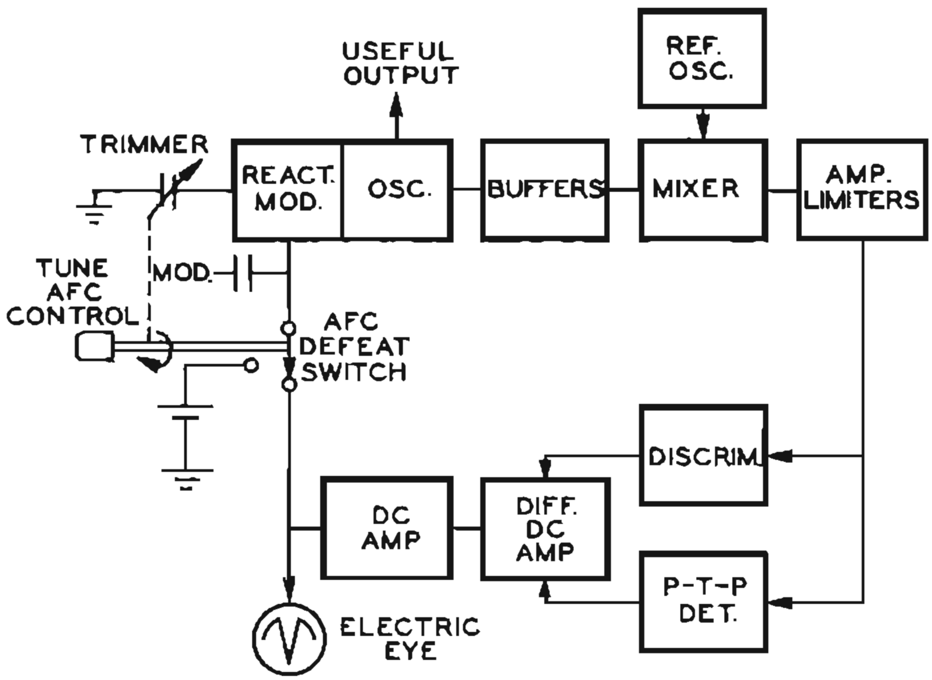

Figure 1. Block Diagram of Overall System.

FM Oscillator

Basically the oscillator to be used is a simple triode oscillator with a tuned plate circuit (Figure 2). A conventional reactance tube circuit has its plate tied to the oscillator plate. In this case the 90° phase shift to the grid is such that the current through the tube lags the voltage across it by 90° and the reactance tube looks like an inductance whose value can be varied by varying the control grid voltage. An ac voltage applied to the control grid modulates the oscillator in the prescribed manner by varying the reatcance tube inductance.

To keep the center frequency where it belongs, a monitoring device constantly monitors the number of cycles of rf occurring in a unit of time. This unit of time is long compared to a cycle of modulation, but short enough to correct for the factors which will bring about unwanted frequency shifts such as thermal changes, filament voltage drifts, and the slow changes that might occur in component values. Fortunately all of these normal instability factors are measured in seconds or longer, while the modulation periods are no longer than 1/20 of a second. The frequency monitor then applies a dc voltage to the reactance tube grid, bringing about a steady-state change in inductance which is of the proper direction and amplitude to correct the center frequency.

Reference Oscillator

The first objective in the frequency monitoring system should be to establish a stable reference to which the oscillator can be compared. A natural choice for this is a crystal oscillator operating near the unknown frequency. The reference oscillator chosen here is a conventional two-tube Butler oscillator (Figure 3) operating on the fifth overtone of a series resonating crystal, which is kept in a temperature-controlled oven for frequency stability. The fm oscillator and reference oscillator frequencies are mixed in a suitable device and the difference frequency is then monitored. Since the unknown varies back and forth with modulation, some offset is required to

YOU WILL FIND . . .

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insure that the difference observed is always in the same direction. If this were not done, and should the difference pass through the zero beat, the result would lead to ambiguity because a certain frequency difference would produce the same beat note, whether it were above or below the reference.

A difference frequency of 100 kc was chosen for this system because the widest deviation to be expected would be 75 kc. This allows 25 kc of safety before zero beat is reached. If broader deviation were required, wider separation would have been considered. To consider specific numbers, the frequencies selected for the system are 48 mc for the carrier and 47.9 mc for the reference oscillator.

By adding the reference, the requirements put upon the frequency monitor are greatly relaxed. For example, a requirement of 0.001% maximum error at 48 mc represents a frequency error of 480 cycles. Four hundred and eighty cycles out of the 100-kc difference on the other hand is roughly ½%. This is approximately a 500 to 1 reduction in the stability which will be required of the frequency monitoring device. Using this approach, it is obvious that it is desirable to reduce the difference frequency to a minimum which is set by the requirements for modulation.

In practice, it is necessary to separate the carrier oscillator from the mixer where the 100-kc difference is generated, in order to prevent modulation of the carrier oscillator by the 100-kc difference frequency. Two cascaded, grounded-grid amplifiers with capacitive voltage dividers are employed to reduce the net stage gain to unity and increase the isolation. This reduces the incidental AM caused by the 100 kc, to a level well below ½%.

Discriminator

A counter-type frequency discriminator (Figure 4) is used to measure the IF and produce a dc output proportional to this frequency. In this circuit, a low-value capacitor is charged to a fixed voltage on each positive cycle and discharged into a high-value capacitor on each negative half cycle. A resistor connected across the high-value capacitor constantly discharges it. The voltage on the high-value capacitor therefore is directly proportional to the rate at which the charging cycles occur, since each charging cycle delivers a constant charge ($Q = CE$; C being the low-value capacitor and E the fixed voltage to which this capacitor charges).

Limiters

To assure constant charging voltage, several stages of limiting are included between the mixer, where the 100-kc IF is produced, and the discriminator. There are four stages which build the 100-kc voltage up and limit it to a constant 200 volts peak to peak. The voltage into the discriminator is necessarily kept quite high to assure a sizeable output from the discriminator. This is necessary to insure that the discriminator output is high compared to thermal voltages and the voltages generated by cathode emission in the dc amplifiers which follow in the circuit.

Detector

There is still a possibility of instability in the output of the discriminator, because the output of the limiters can fluctuate slightly with changes in supply voltages and component values. A change in the limiter output will result in a change in discriminator output

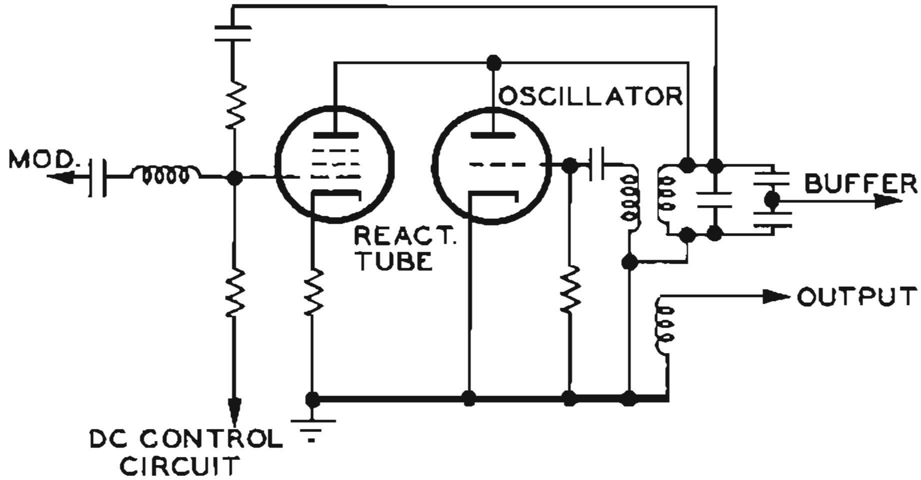

Figure 2. Reactance Tube Oscillator.

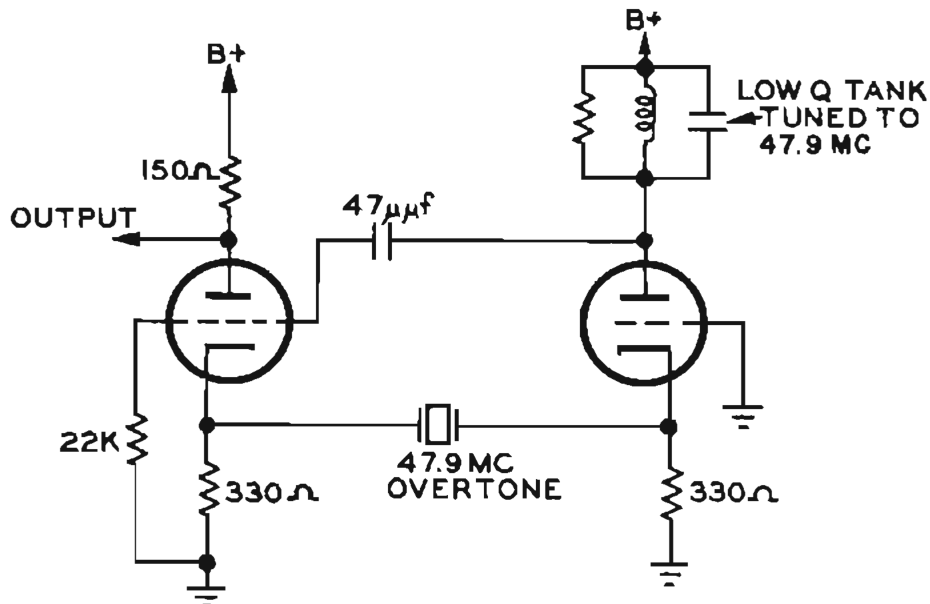

Figure 3. Reference Oscillator.

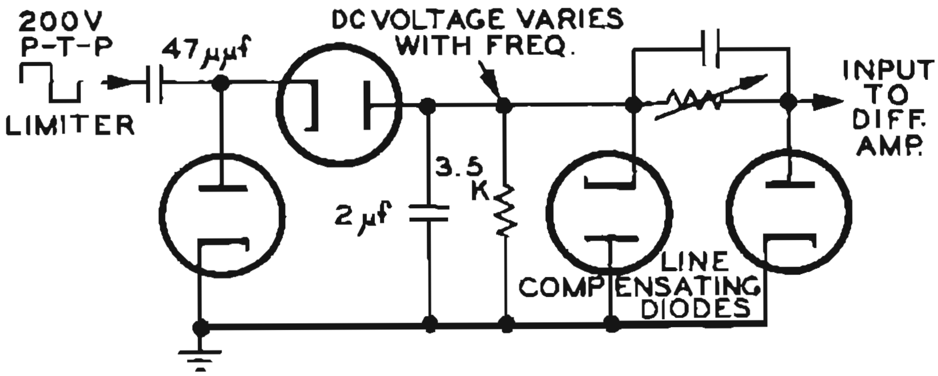

Figure 4. Frequency Counting Discriminator.

which is exactly proportional to the change in limiter output. In order to cancel out any such instability, a peak-to-peak diode detector is also coupled to the output of the limiter. The output of this detector is relatively independent of frequency, but varies directly with peak amplitude.

Differential DC Amplifiers

The discriminator output is connected to one input of a differential dc amplifier, and the output of the peak-to-peak detector is connected to the other input of the differential dc amplifier. Since the output of the peak-to-peak detector is much higher than the output of the discriminator, it must be first divided down to match the output of the discriminator. The differential dc amplifier is so designed that only a difference between the two inputs will result in an output variation. If both inputs vary simultaneously, as would result from a limiter output level change, no change in output results. However, should the frequency change, only the output of the discriminator would vary and an output change would appear amplified in the output of the differential dc amplifier.

The output of the differential amplifier is amplified further in another dc amplifier and then coupled back to the grid of the reactance tube. Inasmuch as the output of the last dc amplifier is at a plate voltage level, it is necessary to refer this output down to the grid voltage of the reactance tube. This is accomplished by returning a voltage divider off the plate of the dc amplifier to a negative supply of roughly —150 volts. Since the plate voltage is about +150 volts in this case, a tap about half way down the divider results in proper grid potential with a loss in forward gain of only 6 db.

With the grid connected to the output of the dc amplifier the system will react strenuously against any change in frequency. The forward gain following the discriminator is sufficient to reduce any change in the natural resonant frequency to less than 1/100 of that change.

AFC Defeat

As might be expected, there is a flaw in this scheme. When the control circuit is first energized and the supply voltages are building up at various rates, it is possible that the 48-mc oscillator, under the influence of a reactance tube warm-up transient, might be forced down to 47.8 mc. This would result in 100-kc difference between the reference oscillator and the fm oscillator; the same difference obtained with the fm oscillator operating at 48 mc. The trouble here is that the phasing in the system is reversed. Normally an increasing frequency would result in an increasing difference. In this case however, an increase in the fm oscillator frequency will result in a decrease in the 100-kc difference. The system is so phased that a decrease in IF results in a voltage coupled to the fm oscillator which will increase the oscillator frequency, and result in pushing the oscillator back to where it belongs. This is true as long as the frequency never gets lower than 47.8 mc. Should the frequency fall below 47.8 mc, the IF would be higher than the normal 100 kc. The discriminator would then operate as if the frequency were too high, resulting in an output to the reactance tube grid which is in the direction to lower the oscillator frequency. Once thi saction starts, there is no stopping it until all of the amplifiers and the reactance tube are saturated and can go no further.

To break this deadlock, a momentary switch is added which breaks the AFC loop (Figure 1) and connects the reactance tube grid to a fixed voltage which is its normal operating voltage. This allows the oscillator to return to a point close to 48 mc. When the loop is closed again, the AFC action returns to normal and the 48-mc frequency is restored.

Electric-Eye Indicator

A 6E5 indicator tube is used as an indication of proper operation. When the output of the last dc amplifier is normal, the eye is approximately half open. If the system is out of control, the eye is completely closed. Therefore, anyone observing the eye will get an immediate indication of proper or improper operation of the system. Since the gain of the dc amplifier is quite high it takes little change in frequency to cause the eye to go from the full-open to the full-closed condition.

Another use is made of this indicator. It is desirable that the reactance tube operating point be fairly constant since it is being used for fm modulation. In time, a drift in component values would result in the operating point of the reactance tube being shifted far off to one side of the linear range in order to hold the frequency accuracy, causing the modulation distortion to increase. To correct for this, an oscillator fine trimmer is coupled to the momentary AFC disabling switch in such a manner that the operator can depress the switch button and then turn it, causing variation of the internal trimmer capacitance. Proper frequency is indicated at the point where the eye begins to change from the open to the closed condition or vice versa. The actual frequency range spanned from the full-open to the full-closed condition is so small that exact adjustment to the half-open condition is not possible. This is of no consequence because exact adjustment will be automatically attained when the control loop is restored. What is important here, is that the frequency is close enough so that no large shift in th enough so that no large shift in the reactance tube operating point is required. The range of the trimmer is small enough and so centered that one could never reach 47.8 mc. If it were possible to reach 47.8 mc, the IF would again be 100 kc and ambiquity would result. Because the drifts being compensated for are relatively small, a range of more than 100 kc is unnecessary.

Voltage Regulation

In order to obtain the desired stability from this system, all dc power supplies are well regulated against line fluctuations and all filaments in the dc amplifiers and discriminator circuit are regulated by a ballast-type regulator. However, even with this regulating system, small variations in frequency result from the small changes in filament voltage and, to a lesser degree, from small changes in dc potentials. To eliminate this problem a simple line voltage compensator (Figure 4) has been employed. The compensator, consisting of two 6AL5 diode sections connected in series, produces a few tenths of a volt as a result of cathode emission; the voltage varying with line voltage. If a part of this voltage is placed in series with one of the dc amplifier grids, and the polarity and level is adjusted properly, all frequency variations due to line voltage can be cancelled. The relationship between filament voltage and this correction voltage is not very linear, but since the 6AL5 tubes are also regulated by the same ballast that regulates the critical stages, their voltage swings very little. A small section of this curve is fairly straight and the correction is quite useful. In practice, the amount of the voltage coupled in series with the grids from the diodes is adjustable and the frequency can be made remarkably independent of line voltage. Since almost all line instability is due to the filament shift, the time constant of the correction is also well matched.

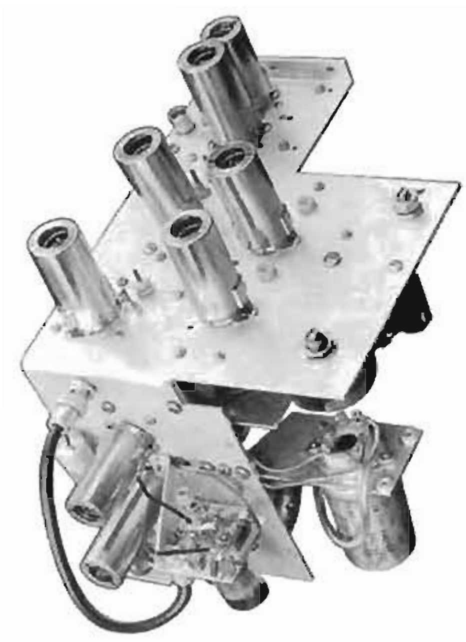

Figure 5. FM Unit for Type 242-A Signal Generator.

BRC Type 242-A Signal Generator

In the BRC Type 242-A signal generator, where it is used, the system discussed here maintains a frequency accuracy of 0.0005%. It produces frequency modulation of 75 kc deviation with audio frequencies well over 100 kc, with less than 1% distortion. The output of the system is added to one of 150 different crystal frequencies. The sum frequency is then multiplied by 4 to produce 1-mc channels running from 400 to 550 mc at a maximum deviation of 300 kc (75 times 4).

Because the Type 242-A is designed to operate in an ambient temperature range of −40°F to +137°F, the entire fm oscillator control circuit is contained in a chamber in which the temperature is thermostatically controlled. This would not be necessary at normal room ambient range as the internal thermostat only holds within ±10°F.

Conclusion

The general principles and circuitry used in the system described here could be easily applied to any simple channel fm oscillator such as a transmitter It is obviously not suited to variable frequency operation, but through the use of the adding system employed in the Type 242-A, could be a part of a flexible system.

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Signal Generator Performance

HARRY J. LANG, *Sales Manager*

As designers and manufacturers of laboratory signal generators since 1941, we have come to establish certain standards of performance for this type of instrument which should be of interest to both the designer and the user. While some of these standards may be termed implied and form a nucleus of "unwritten specifications", they are, in many cases, of paramount importance in the selection of a signal generator for a particular application. In this article, an attempt will be made to outline these standards of performance which actually underlie our design philosophy.

A signal generator, in the broadest sense, may be defined as a precise signal source of known, stable, and controllable characteristics. Basically, the device must generate a signal of known frequency, level, and modulation and may be specified in terms of accuracy, settability, readability, stability, reliability, and convenience.

Accuracy

Here we are concerned with the absolute accuracy of frequency, output, and modulation. For a device to qualify as a signal generator, in terms of the current state of the art, frequency accuracy must, in general, be better than 1% at all points within its range. All unwanted spurious outputs should be minimized to a level at least 30 db below the desired output frequency. Output must be directly calibrated to an accuracy of between 1% and 20%, depending upon the region of the frequency spectrum within which the device operates, with accuracy generally decreasing with increasing frequency. Output level must be continuously variable over a range of from roughly 10% to typical system sensitivities to better than 1000% of this value. Basic accuracy must be maintained at all output levels at a constant impedance, and leakage should be negligible at the lowest calibrated output level.

Modulation, whether amplitude, frequency, pulse, or phase, should be directly calibrated to better than 10% and should be continuously variable over a range covering all current system requirements. Unwanted modulation by-products; e.g., incidental fm, or am on fm, must be reduced to a negligible level and the modulation system must operate with specified accuracy over the range of modulating frequencies normally encountered in system operation. Modulation purity, or distortion, must also be minimized to provide a signal that will permit accurate evaluation of

Figure 1. Type 150-A Signal Generator, First FM Signal Generator to be Manufactured by BRC.

the distortion products generated within the system under test.

Settability

Here, again, we are concerned with the fundamental characteristics of frequency, output, and modulation with basic requirements dictated by the system to be tested. Settability may be generally specified at better than 10 times basic accuracy and modified by special system requirements. For example, narrow-band, crystal-controlled communication receivers require frequency settability in the order of 0.001%.

The matter of repeatability of settings also becomes extremely important since, in many cases, the generator as a development tool, is used to sense minute changes in circuit characteristics rather than absolute values. Furthermore, since certain specialized applications may require individual calibration of generators to a fraction of their generally specified accuracy, the calibration accuracy that can be obtained will be directly determined by both settability and repeatability.

Readability

While this characteristic is closely related to both accuracy and settability, it is important to note that sufficient calibration points must be provided to permit direct interpolation to sepcified accuracy at all points within the operating range. Conversely, the use of an excess number of calibration markings may be misleading and imply a level of accuracy inconsistent with generator performance.

Stability

A signal generator, to perform its intended function, must provide specified accuracy independent of external conditions encountered in normal operation. Factors which must be considered in signal generator design, include the following:

1. *Input Power Supply.*

Since all commercial sources of power provide regulation in the order of 5%, and further, since laboratory or production line applications may be further aggravated by local load conditions, signal generators must be designed to operate with minimum line input variations of ±10%.

2. *Ambient Temperature.*

Since both factories and laboratories may experience ambient temperatures varying between 50°F and 100°F, adequate compensation must be incorporated into the generator design to provide specified accuracy over this range.

3. *Vibration.*

Since many generators will be operated in plants where vibration from heavy production machinery will be encountered, every effort must be made to reduce variations in frequency and output as well as spurious modulation from this source.

4. *Operating Cycle.*

Since most generators are not operated on a continuous duty cycle, adequate provision must be made to minimize the warmup time required to obtain specified performance.

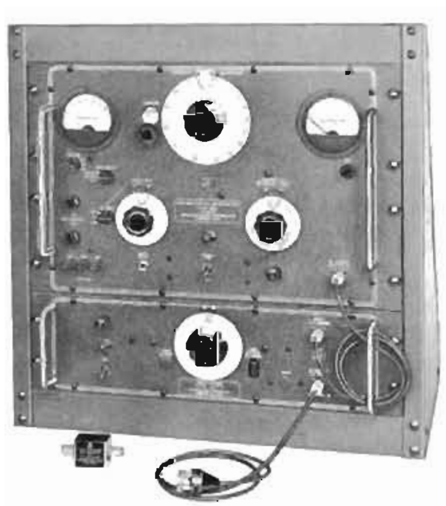


Figure 2. Rack Mounted View of the Type 202-F Signal Generator and the Type 207-F Univerter.

Reliability

Though the area of reliability is currently under continuous discussion in the field of military electronics, it is sometimes under-emphasized with respect to commercial equipment. Downtime and repair expense are not always catastrophic in the case of commercial test equipment, but they do represent lost value to the user. A well-designed signal generator should provide a minimum of ten years reliable service with only the need for occasional replacement of tubes and certain other minor components that may have a shorter life cycle. Rugged mechanical design, coupled with adequate derating of electrical components, is necessary for reliable service.

Figure 3. Type 240-A Sweep Signal Generator.

Convenience

A well-designed signal generator should permit simple, rapid, foolproof opertaion by using personnel. All controls and dials should be legibly marked to show their function; dial scales should be direct reading and readily visible under laboratory conditions with a minimum of eye strain. All front panel controls should be grouped functionally and should be human engineered for ease of adjustment. The use of correction charts and nomographs should be avoided.

While this discussion of signal generator performance is not intended to be all-inclusive, we have tried to outline the major considerations given to the design of our products in this field: these considerations being an outgrowth of experience, gained over the years, in

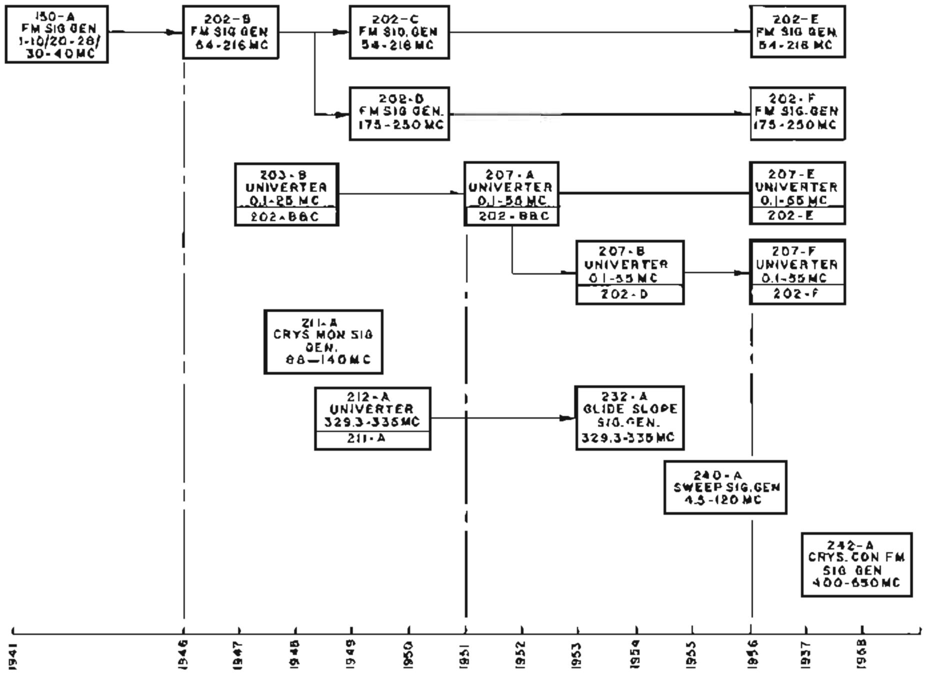


Figure 4. Development and Evolution of BRC Signal Generators.

signal generator design. In 1941, BRC produced the first commercial FM Signal Generator, our Type 150-A (Figure 1). We have continued to specialize in the field of FM and now produce our Types 202-E and 202-F (Figure 2) FM-AM Signal Generators in which we have attempted to provide instrumentation to meet the ever-changing requirements of current day systems. The Types 207-E and 207-F Univerters are companion accessories for these generators and provide extended frequency coverage. Sweep Signal Generator, Type 240-A (Figure 3), offers signal generator performance in a wide-band sweep source. The Type 211-A Crystal Monitored Signal Generator and the Type 232-A Glide Slope Signal Generator are specifically designed for the testing and calibration of aircraft ILS and Omni-Range systems. Our latest, the Type 242-A Crystal-Controlled FM Signal Generator provides fm outputs with crystal accuracy. Figure 4 historically traces the development and evolution of these products.

MEET OUR REPRESENTATIVES

CROSSLEY ASSOCIATES

HARRY J. LANG, *Sales Manager*

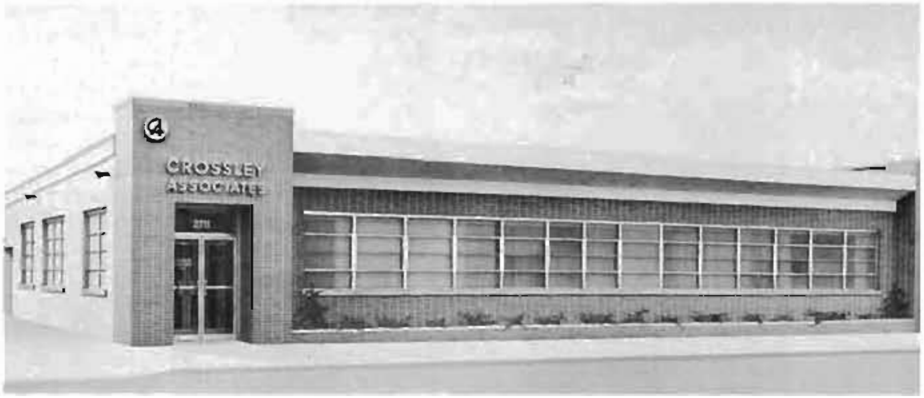
The first in a series of articles intended to provide a capsule history and facilities report on the sales representatives who sell and service BRC products throughout the world.

Founded in 1937, Crossley Associates of Chicago, Illinois, is the oldest engineering sales representative organization in the Midwest specializing in the field of electronic instrumentation. Pioneers in technical customer service, they have helped to establish distribution channels for the entire instrument industry and have developed application engineering techniques and set standards of performance for others in this field.

Alfred Crossley, President and Founder of the company, has been active in the radio and electronics field since 1916 and holds numerous electronic patents. Notable among these is a patent for a crystal-stabilized oscillator. After filling many key engineering posts in the radio industry, he entered the field of consulting engineering and subsequently made the transition to engineering sales representation in 1937. Boonton Radio Corporation, in order to provide a new level of customer service to the growing electronics industry in the Midwest, became the first manufacturer to engage the services of the newly formed engineering sales company. Crossley Associates thus became the first representative for BRC products.

ALFRED CROSSLEY
President and General Manager

Over a period of twenty-one years, Crossley Associates has vastly expanded and modernized both its facilities and services to keep pace with the ever-changing needs of the electronics industry. With headquarters in Chicago, the organization now maintains three branch sales offices in Dayton, Indianapolis, and St. Paul to provide local service for customers throughout the Midwest area. In three modern, well-equipped service laboratories, complete facilities are maintained for the evalua-



Chicago Headquarters Crossley Associates

tion, application engineering, calibration, and repair of all of the products manufactured by fourteen leading producers of precision electronic equipment. Clerical services are provided for the distribution of technical information and a separate department processes all phases of customer orders including placement, expediting, and factory liaison. Direct TWX service between all sales offices and factories offers up-to-the-minute customer information.

Frank Waterfall, the company's Vice President and Sales Manager, joined Crossley Associates in 1946 after post-graduate studies at Indiana University and the University of Minnesota, augmented by extensive instrumentation experience at the Naval Research Laboratory during World War II. Other members of the Crossley family include ten field engineers who devote their entire effort to the solution of customer measurement problems, and an additional twenty office personnel who support all field engineering activities.

A strong believer in education, Crossley Associates maintains scholarship programs with twelve major Midwestern universities. The field engineers operate under a continuous educational program which includes technical field seminars and factory training courses. Two Crossley trainees are currently attending universities to complete and broaden their engineering education.

It is Crossley Associates' policy to apply, sell, and service the finest in precision electronic instrumentation and components for the communications, electronics, and electromechanical measurement fields. Service is perhaps the most important part of the business since it transcends many years beyond the initial sale of a product. Both Crossley and BRC believe they are duty-bound not only to place in the hands of their customers the proper instrumentation for their particular application, but also to ensure that these products continue to provide precise, reliable answers to ever-changing problems in the rapidly expanding field of electronics.

We at BRC proudly salute Crossley Associates for their continuing record of faithful service to our many customers throughout the Midwest.

FRANK WATERFALL
Vice-President and Sales Manager

SERVICE NOTE

Modification of Glide Slope Signal Generator, Type 232-A, For Improved RF Output

Some time ago, BRC received reports of Type 232-A breakdowns which were attributed to loss of rf output. The instruments concerned exhibited a rapid decrease in rf output. This decrease in rf output resulted in the loss of output reserve, and eventually brought about a condition where "red line" operation and consequently satisfactory operation of the signal generator was no longer possible.

After extensive investigation of the problem, BRC engineers traced the trouble to two tubes in the rf generator circuit. It was discovered that if the filament of the rf demodulator tube (V6/6173) is operated above 6.3 volts, the tube will become gassy and load the output tank. Tests also revealed that the plate dissipation in the rf doubler tube (V5/538-B) runs close enough to its advertised limits to reduce the life of this tube.

The following minor circuit changes eliminated this rf output problem and provided satisfactory operation of the Type 232-A Signal Generator.

1. A 2.2-ohm resistor was added in series with the ungrounded filament connection on tube V6 (6173). This resistor is a wire-wound type capable of handling 0.25 watts.
2. Resistors R24 and R25 (10k ohms) were replaced with 18k-ohm resistors. These resistors are composition type, with a ±5% resistance tolerance and a ½-watt power rating.

Owners of Type 232-A Signal Generators are advised that these modifications will be made whenever an instrument is returned for repairs.

EDITOR'S NOTE

Q Meter Contest Award

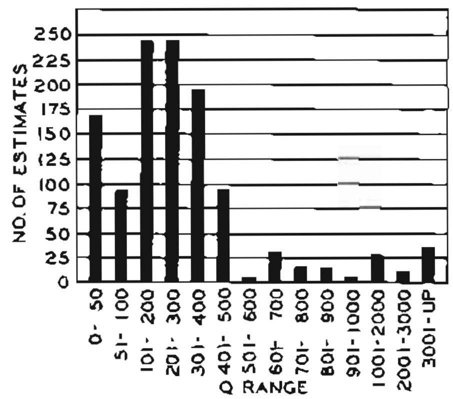
The Q of the coil displayed in the BRC booth at the IRE show is 378. The winning estimate was submitted by Mr. Eugene J. Caron, a project engineer with Radio Condenser Co. of Camden, N. J.

Since Mr. Caron was one of three persons who estimated 378, a drawing was held to determine the winner. The other 378 estimates were submitted by Mr. Chester Warner of Great Neck, N. Y. and Mr. J. Pakan of Forest Park, Ill.

A total of 1198 entries were submitted, with estimates ranging from less than 50 to over 3000. We report, with considerable amazement, that there were twenty one estimates within 1% of the actual measured Q. A list of persons submitting these near misses and a graph showing the distribution of estimates are given below.

ESTIMATES WITHIN 1%

Estimate	Submitted By
375	R. Friedman, Polarad Electronics, L. I. C., N. Y.
375	D. W. McLead, Norden Laboratories, White Plains, N. Y.
375	M. Leonard, Columbia University, N. Y. C.
375	C. Briggs, Mass. Inst. of Tech., Cambridge, Mass.
375	F. A. Blackshear, Sperry Gyroscope, Great Neck, N. Y.
375	R. P. Thurston, Waters Mfg. Co., Wayland, Mass.
375	H. C. Hausmann, Arcs Industries, W. Islip, N. Y.
375	W. K. Springfield, IBM, Endicott, N. Y.
375.5	M. Fischman, Sylvania Electric, Bayside, N. Y.
375.6	H. E. Whitted, Western Electric, Winston-Salem, N. C.
376	T. B. Robinson, National Co., Malden, Mass.
376	M. R. Easterday, Bendix Aviation, Kansas City, Mo.
377	M. Freedberg, N. Y. C. Community College, N. Y. C.
377	D. T. Geiser, Sprague Electric, N. Adams, Mass.
377.1	R. C. Ferris, Lockheed Aviation, Marietta, Ga.
378	C. WARNER, Sperry Gyroscope, Great Neck, N. Y.
378	EUGENE J. CARON, Radio Condenser Co., Camden, N. J.
378	J. PAKAN, A.R.F. Products, Forest Park, Ill.
379	G. Elliott, Rochester, N. Y.
380	S. B. Alexander, Emerson Radio, Jersey City, N. J.
380	W. A. Palmisano, Army Chemical Center, Edgewood, Md.



Distribution of Q Estimates.

The display coil was measured at 10 megacycles in the BRC standards laboratory on a Type 260-A Q Meter which was previously calibrated against BRC Q Standards. Ten separate measurements were made, the average Q measurement being 378 and the average capacitance measurement being 58 $\mu\mu f$.

Mr. Caron visited BRC on May 1 to accept the Q Meter from Dr. G. A. Downsbrough, President of the company. During his visit, he stopped at your editor's desk and passed along the following information concerning his career in the engineering field. After he received his B.S. in Radio Engineering from Tri State College at Angola, Indiana, he was successively employed in an engineering capacity by Raytheon Corporation, Hazeltine Corporation, and Radio Condenser Co. Mr. Caron joined Radio Condenser Company in 1948 as Supervisor of TV Engineering. He is presently with the Special Apparatus Division of Radio Condenser Company as Project Engineer.

Along with our special congratulations to Mr. Caron, BRC wishes to thank our many friends who visited with us at the IRE show.

Dr. G. A. Downsbrough, President of BRC, presents a Type 160-A Q Meter to Mr. E. J. Caron of Radio Condenser Co., winner of the Q Meter contest.

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Transistor Measurements With The HF-VHF Bridge

GEORGE P. McCASLAND, *Sales Engineer*

Figure 1. The author is shown measuring a transistor, using the RX Meter, a battery power supply, and a specially made transistor measuring jig.

Introduction

The very high-power efficiency at low-power levels together with the reliability and low-cost potentials of the transistor have led to increased usage of this device by circuit designers in both the commercial and military fields of electronics. Hearing aids, portable radios, phonographs, dictating machines, portable cameras, machine-tool controls, clocks, and watches are but a few examples of the commercial applications of transistors. Along with many other applications, the military have employed the transistor in the "Explorer" and "Vanguard" satellites now circling the earth.

As the usage of the transistor increases it is apparent that there is a need to develop new measuring techniques. The transistor circuit designer can no longer get by with specifications published by transistor manufacturers alone; (Normally, the manufacturers will specify transistor parameters for a given set of bias conditions and a single frequency only.) he is now often faced with the problem of determining parameters for a wide range of bias and frequency conditions. This article describes a transistor measuring technique, using the RX Meter and certain hybrid equations, which will yield information about the parameters of transistors over a wide range of bias and frequency conditions.

The RX Meter is well suited to transistor measurements because its bridge elements will pass, directly, a current of up to approximately 50 milliamperes, and its two-terminal measurements can furnish the parameters for radio-frequency transistor circuit design. Contributing to the ease with which transistor measurements can be made are the self-contained design of the instrument, with the signal generator, bridge, and detector in a single, compact package, and the fact that the "unknown" terminals are located on the flat, top surface of the instrument, where attachment of the measuring jig may be conveniently made.

The idea of using the RX Meter for transistor measurements is not a new one; a number of research, development, and production groups have been engaged in this type measurement for some time. For example, Messrs. Earhart and Brower of Texas Instruments recently used the RX Meter to measure a new VHF silicon transistor.[4] Because of the lack of published information on the theory and practice of transistor

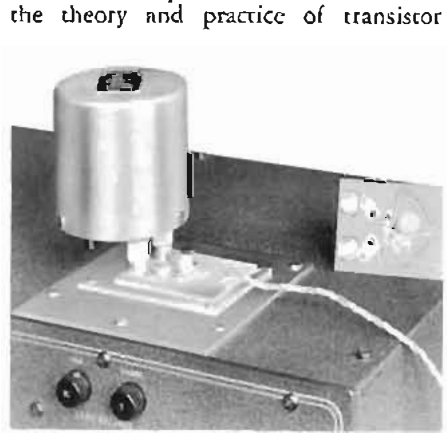

Figure 2. Jig box, shown, includes binding posts to which the 103-A series coil or other components may be connected for extending the RX Meter capacitance, inductance, or resistance range.

YOU WILL FIND . . .

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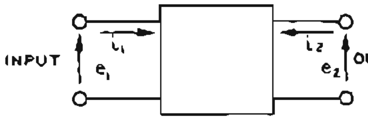
measurements with the RX Meter, however, it has been found that few of these groups are using the RX Meter to the limit of its capabilities. This article is intended to fill the gap in the literature, and at the same time, promote a more complete understanding of the use of the RX Meter for measuring transistors by persons currently using the instrument for this purpose, and by those persons unaware of its transistor-measuring application.

Transistor Parameters

Consider the following network equivalent of the transistor.



(1)

Linear equations can be written using a set of independent variables to relate e_1, e_2, i_1 and i_2; the independent variables being the input, output, and transfer characteristics of the transistor commonly known as the transistor parameters. These transistor parameters are constants for a given set of bias and frequency conditions. One of the most popular and most widely used sets of transistor parameters is the hybrid or h set of parameters. The linear equations for the transistor, represented by network (1), in terms of hybrid parameters are:

$$e_1 = h_{11}i_1 + h_{12}e_2$$
$$i_2 = h_{21}i_1 + h_{22}e_2$$

where the h parameters are h_{11}, h_{22}, h_{12}, and h_{21}. The choice of number subscripts here is based on personal preference. IRE Standards[3] suggest the use of either number or letter subscripts, as convenient. Table I is a cross-reference of number and letter subscripts assuming a common-base transistor configuration.

TABLE I

COMMON-BASE SUBSCRIPTS

Number	Letter
h_{11b}	h_{ib}
h_{12b}	h_{rb}
h_{21b}	h_{fb}
h_{22b}	h_{ob}

Determining the Hybrid (h) Parameters

The h parameters can be determined by solving the hybrid equations. By arbitrarily open- and short-circuiting pairs of terminals in (1), a current or a voltage can be made zero to aid in the solution. Parameters h_{11}, h_{22}, h_{21}, and h_{12}, are numerically evaluated with the help of RX Meter measurements. The method of evaluation is outlined in Table II below.

TABLE II

SOLUTION OF THE HYBRID EQUATION

(Assuming Common-base Configuration)

Parameter	Circuit Condition	Description
$h_{11b} = \dfrac{e_1}{i_1}$	$e_2 = 0$	Input impedance with output short-circuited.
$h_{12b} = \dfrac{e_1}{e_2}$	$i_1 = 0$	Reverse voltage transfer ratio with input open-circuited.
$h_{21b} = \dfrac{i_2}{i_1} = -\alpha$	$e_2 = 0$	Forward short-circuit current transfer ratio with output short-circuited.
$h_{22b} = \dfrac{i_2}{e_2}$	$i_1 = 0$	Output admittance with input open-circuited.

The various circuit conditions for each h parameter refer to the ac circuit only. DC bias voltages are not disturbed when the ac circuit conditions are changed. From Table II it should be obvious that an input measurement on the RX Meter provides h_{11b} directly. By converting an output RX Meter measurement to an admittance, h_{22b} is also obtained directly.

Parameters h_{21b} and h_{12b} relate voltages and currents on both input and output sides of the network providing the network transfer characteristics. The h_{21b} parameter is found numerically from the ratio of two input impedance measurements. Parameter h_{12b} is found from the product of the difference of two output admittances and the ratio of h_{11b} over $-h_{21b}$. The derivation of h_{21b} and h_{12b} is given in the appendix. In summary, the method of obtaining the common-base h parameters from RX Meter measurements is simply and conveniently shown in Table III.

To obtain common-emitter or common-collector h parameters a set of simple conversion formulas can be used. As an example, we can convert h_{22b} to h_{22e} with the formula

$$h_{22e} = \frac{h_{22b}}{(1 + h_{21b})}.$$

The above formula and other conversion formulas are given by Scher[6]. These formulas yield approximate values.

RX Meter Jigs

The RX Meter measurements indicated in configurations A through D in Table III require the use of four jigs which attach the transistor to the RX Meter and supply proper dc bias. These jigs can all be operated from a common power supply. A schematic diagram of the jigs with a transistor in the socket is shown in Figure 3. Each diagram represents one of the configurations correspondingly marked A through D in Table III.

The specific jigs described in this article were designed for a Raytheon Type 2N417 PNP transistor for measurements in the 20-mc range. The 0.01 μf ceramic by-pass capacitors, power supply, and jig socket selections all reflect the transistor to be measured and the approximate frequency range. The by-pass capacitors have good by-passing action in the 20-mc range and the jigs perform well in this range. For frequencies appreciably different from 20-mc, different capacitors may have to be selected. At frequencies approaching 250-mc the jig series inductance may require evaluation. However, no serious problems are anticipated in using similar jigs over the entire 500-kc to 250-mc range of the RX Meter.

Since the power supply (Figure 4) has high resistance in the emitter bias

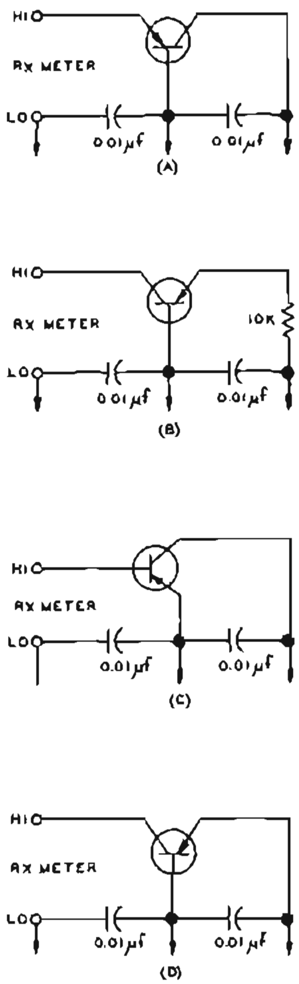


Figure 3. Jig Schematics

circuit, the power supply resistance determines the emitter bias current and bias will be a constant as different transistors are plugged into the jigs. Use of the 90-volt bias battery permits sufficient bias to be developed even with the h_{22b} jig which has a fixed 10k resistor in the emitter circuit.

The socket used in our jigs accommodates the transistor type with four pins mounted on a 0.200-inch diameter circle. However, the jigs can be modified slightly to accommodate sockets for different type transistors, including the new universal transistors sockets.

As a special feature, the h_{11b} and h_{11b} jigs utilize Q Meter binding posts in parallel with the RX Meter terminals for use in extending the RX Meter ranges.

A drawing showing construction details of the jigs is being prepared. Interested persons may obtain a copy by calling or writing Boonton Radio Corporation, or one of our representatives.

Typical Measurements and Calculations

To illustrate transistor measurements the common-base h parameters of a Raytheon Type 2N417 transistor have been determined for the following conditions using the RX Meter and the jigs.

$$V_{CB} = -6 \text{ volts}$$
$$i_e = 5 \text{ ma.}$$
$$f = 20 \text{ mc}$$

The four necessary RX Meter measurements (one measurement in each of the four jigs) and necessary conversions to rectangular and polar impedance and admittance coordinates are shown in Table IV.

RX Meter readings are the parallel R_p and C_p equivalent of the unknown and can be readily converted to rectangular and polar impedance forms. Readings are converted to rectangular admittance by taking the reciprocal of the RX Meter parallel equivalents expressed in ohms. The reciprocal of R_p is the conductance G and the reciprocal of XC_p in ohms is the susceptance B, where G and B are in ohms.

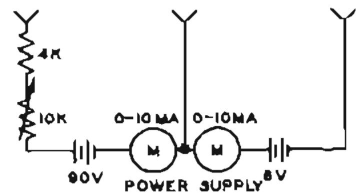


Figure 4. Power Supply Schematic

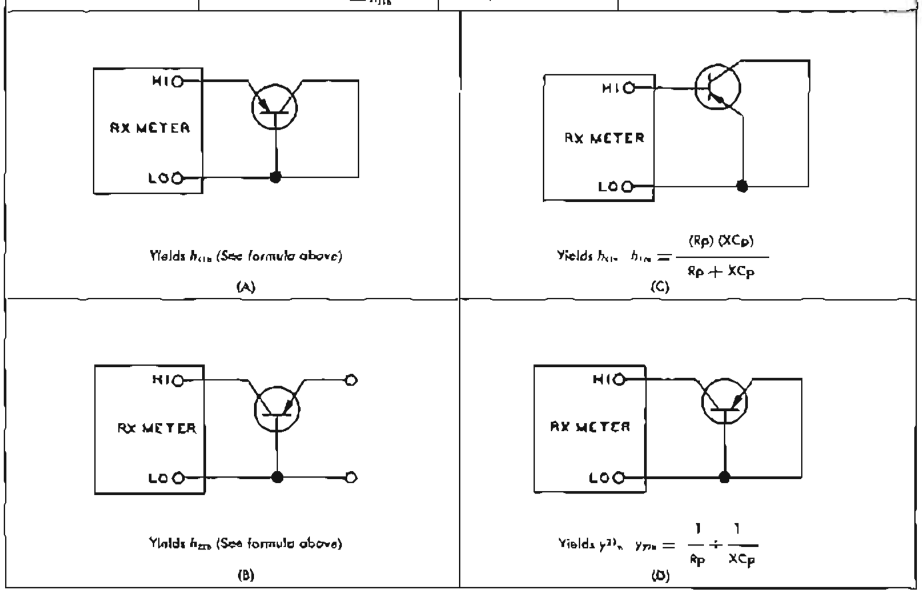
When converting from the parallel equivalents to the rectangular form of impedance, a series-parallel conversion chart such as is found in the RX Meter Instruction Manual is of help. In working with the data in Table IV, it was convenient to divide the parallel equivalents by 20 to coter them into the chart and to multiply the series equivalent answers by 20 after removing the values from the chart. To use the chart properly, a given combination of R_p and C_p must both be divided by and the answers multiplied by the same number. An ordinary reactance chart was used in converting C_p readings to reactance.

Using jig B of Figure 3 to make the RX Meter measurement necessary to obtain h_{22b}, the input circuit of the transistor is effectively open-circuited

TABLE III

METHOD FOR OBTAINING, COMMON-BASE h PARAMETERS FROM RX METER MEASUREMENTS

Parameter	Formula	Units	Required RX Meter Measurements
h_{11b}	$\frac{(Rp)(XCp)}{Rp + XCp}$	Ohms	Configuration A below
h_{22b}	$\frac{1}{Rp} \div \frac{1}{XCp}$	Ohms	Configuration B below
$h_{21b} = -\alpha$	$-\frac{h_{11b}}{h_{11b}} + 1$	Magnitude and phase angle	Configuration A & C below
h_{12b}	$(y_{22b} - h_{22b}) \frac{h_{11b}}{-h_{21b}}$	Magnitude and phase angle	Configurations A, B, C, & D below



Yields h_{11b} (See formula above) (A)

Yields h_{11b}, $h_{11b} = \frac{(Rp)(XCp)}{Rp + XCp}$ (C)

Yields h_{22b} (See formula above) (B)

Yields y_{22b}, $y_{22b} = \frac{1}{Rp} \div \frac{1}{XCp}$ (D) with a 10k resistor. To test the effectiveness of this open circuit, 7.5k and 4.7k resistors were substituted for the 10k resistor, while constant emitter bias was maintained, without materially changing the RX Meter's indication of output admittance.

The target of this example is the four common-base h parameters. From Table IV h_{11b} and h_{22b} are directly available in rectangular impedance form.

$$h_{11b} = 65 - j66 \text{ ohms}$$

$$h_{22b} = (0.505 + j1.80) \times 10^{-3} \text{ mhos}$$

From Table III:

$$h_{21b} = -\alpha = -\left(\frac{h_{11b}}{h_{11c}} + 1\right)$$

Values of polar coordinates from Table IV are substituted in this equation and the following calculations are performed.

$$-\alpha = -\left(\frac{93 \angle -45^\circ}{153 \angle 200^\circ} + 1\right)$$

$$= -(.61 \angle -245^\circ + 1)$$

$$= -(-.26 - j.56 + 1)$$

$$= -(+.74 - j.56)$$

$$= -(0.93 \angle -37^\circ)$$

$$h_{12b} = -\alpha = 0.93 \angle 143^\circ$$

Again from Table III:

$$h_{12b} = (y_{22b} - h_{22b}) \frac{h_{11b}}{-h_{21b}}.$$

Inserting values from Table IV, the following calculations are performed.

$$h_{12b} = \Big[(1.61 + j1.31) - (.505 + j1.80)\Big](10^{-3}) \frac{93 \angle -45^\circ}{93 \angle -37^\circ}$$

$$= (1.105 - j.49)(10^{-3})(1 \angle -8^\circ)$$

$$= (1.21 \times 10^{-3} \angle -24^\circ)(1 \angle -8^\circ)$$

$$h_{12b} = 1.21 \times 10^{-3} \angle -32^\circ$$

A summary of the h parameters measured and calculated for the 2N417 transistor are given in Table V.

RX Meter Operating Techniques

In preparation for a measurement, the RX Meter is balanced as usual with a jig attached to the terminals but without a transistor in the socket. The use of the jigs does not in any way interfere with normal operation of the RX Meter.

With normal bridge operating conditions, the voltage appearing at the RX Meter terminals may be 100 to 500 millivolts or more. This is sufficient, in many cases, to drive a transistor beyond its linear range of operation. The terminal voltage, however, may be reduced to 20 millivolts or, in some cases less than 20 millivolts, by reducing the level of the oscillator output with a series resistor in the oscillator+B lead. (See page 16 of the Instruction Manual and Notebook #6[2].) During the measurements to obtain the data presented in Table V, good operation of the RX Meter was obtained at actual measured terminal voltages of 5 to 25 millivolts. The RX Meter measurements did not vary over this voltage range, indicating linear operation of the transistor for this range of signal level.

The Q Meter binding posts on the h_{11c} and h_{11b} jigs provide for easy connection of Type 103-A Inductors and good commercially available capacitors and resistors to extend the RX Meter capacitance, inductance, and resistance ranges. RX Meter range extension is explained in detail in the Instruction Manual and in a previous Notebook article[2].

TABLE IV

CONVERSION OF RX METER MEASUREMENTS TO RECTANGULAR AND POLAR IMPEDANCE AND ADMITTANCE COORDINATES

Jig	RX Meter Readings		Parallel Equivalent		Rectangular Z	Rectangular Y	Polar Z
	Rp (ohms)	Cp ($\mu\mu f$)	Rp (ohms)	Cp (ohms)	ohms	ohms	ohms
C (h_{11e})	168	+ 17.5	168	$+j454$	$-144 - j52$		$153 \angle +200^\circ$
A (h_{11b})	130	— 60.7	130	$-j131$	$65 - j66$		$93 \angle -45^\circ$
D (y_{22b})	620	+ 9.0	620	$+j885$		$1.61 + j1.13$	
B (h_{22b})	1980	+ 14.3	1980	$+j556$		$0.505 + j1.80$	

TABLE V

MEASURED AND CALCULATED h PARAMETERS

Parameter	Numerical Value
h_{11b}	$65 - j66$ ohms
h_{22b}	$(0.505 \times j1.80) + 10^{-3}$ mhos
h_{21b}	$0.93 \angle 143^\circ$
h_{12b}	$1.21 \times 10^{-3} \angle -32^\circ$

Application of the h Parameters

The maximum power gain for the 2N417 in the common-base configuration can be readily calculated assuming conjugate input and output impedance matching and lossless neutralization[7].

$$P.G. = \frac{(h_{21b})^2}{4h_{11br}\,h_{22br} - 2\,\text{Re}\,(h_{12b}\,h_{21b})}$$

In this equation, h_{11br} and h_{22br} are real parts of h_{11b} and h_{22b} and Re $(h_{12b}h_{21b})$ is the real part of the product of h_{21b} and h_{21b}. The values from Table V are substituted in the formula for power gain and the necessary calculations are performed.

$$P.G. = \frac{(.93 \angle 143^\circ)^2}{4 \times 65 \times .505 \times 10^{-3} - 2\,\text{Re}\,(1.21 \times 10^{-3} \angle -31^\circ \times .93 \angle 143^\circ)}$$

$$= \frac{.864 \angle 286^\circ}{131 \times 10^{-3} - 2\,(.403 \times 10^{-3})}$$

$$= \frac{.864 \angle 286^\circ}{131 \times 10^{-3} - .806 \times 10^{-3}}$$

$$= \frac{.864 \angle 286^\circ}{130.194 \times 10^{-3}}$$

$$\text{Power Gain} = 6.62 \angle 286^\circ = 6.62$$

The author wishes to express his appreciation to Mr. D. E. Thomas of the Bell Telephone Laboratories for his valuable assistance in connection with this work.

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THE AUTHOR

George P. McCasland joined the BRC staff as Sales Engineer in January, 1958. From 1954 to 1957 he was associated with the Esso Standard Oil Co., holding posts with Electrical Engineering and Budget Groups of that company's Baton Rouge Refinery.

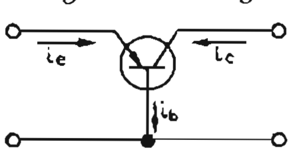
Mr. McCasland received a BEE degree with a Communications Option from the University of Virginia in 1952 and a masters degree in Industrial Management from M.I.T. in 1954.

APPENDIX

Derivation of h_{21} and h_{12} Formulas

h_{21b}

The elementary current relationships of the junction transistor are shown in the following schematic diagram:



where: $i_e = i_c + i_b$ by Kirchoff's law (1)
$i_c = \alpha\, i_e$ by definition. (2)

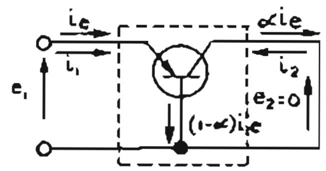
Substituting (2) in (1):

$$i_e = \alpha\, i_e + i_b$$

$$i_e - \alpha\, i_e = i_b$$

$$i_e\,(1 - \alpha) = i_b \qquad (3)$$

The transistor is redrawn substituting αi_e for i_c from equation (1) and $i_e(1-\alpha)$ for i_b from equation (3) and short-circuiting the output circuit in (4). The dotted box symbolizes the network for which network current $i_1 = i_e$ and $i_2 = -\alpha\, i_e$.



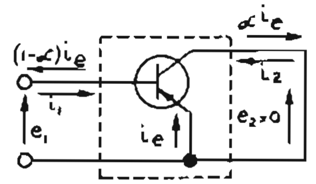
(4)

$$h_{11b} = \frac{e_1}{i_1}, \qquad (5)$$

when $e_2 = 0$ by definition.
Substituting i_e for i_1 in (5), since $i_1 = i_e$ from (4)

$$h_{11b} = \frac{e_1}{i_e} \qquad (6)$$

The transistor is now redrawn schematically in the common-emitter configuration.



(7)

where: $i_1 = -i_b = -(1-\alpha)\, i_e$.

Now, by definition, $h_{11e} = \frac{e_1}{i_1}$, when $e_2 = 0$. (8)

Substituting the equality $i_1 = -(1-\alpha)\, i_e$ shown in (7);

$$h_{11e} = \frac{e_1}{-(1-\alpha)\, i_e} \qquad (9)$$

$$h_{11b} = \frac{e_1}{i_e}, \text{ repeating (6)}$$

By examination of (6) and (9), we see that if h_{11e} of (9) is multiplied by $-(1-\alpha)$, the product will be equal to h_{11b} of (6) as expressed in (10) below.

$$h_{11e}\left[-(1-\alpha)\right] = h_{11b} \qquad (10)$$

$$\text{then: } -(1-\alpha) = \frac{h_{11b}}{h_{11e}}$$

$$\alpha = \frac{h_{11b}}{h_{11e}} + 1$$

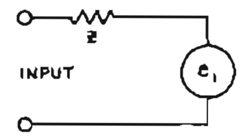
$$h_{21b} = -\alpha = -\left(\frac{h_{11b}}{h_{11e}} + 1\right) \qquad (11)$$

h_{12b}

For the transistor network shown below, representing a transistor in the common-base configuration, the reverse voltage transfer ratio by definition is:

$$h_{12b} = \frac{e_1}{e_2}, \text{ when } i_1 = 0. \qquad (12)$$

By Thevenin's Theorem, the input to the network can be represented as shown below.



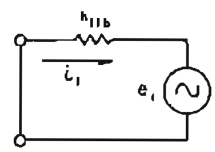
(13)

The open circuit input voltage, e_1, can be assumed to be caused by a voltage generated within the network, which in this case is a voltage transferred back from the output circuit. Z is the short-circuit input impedance, a part of Thevenin's concept. Since h_{11b} is the short-circuit input impedance of (13), it can be substituted for Z. The input can then be short-circuited and the result shown in (15), where:

$$e_1 = h_{11b} i_1$$

and

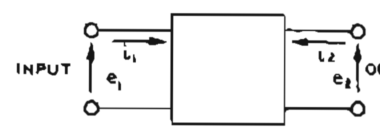
$$\frac{e_1}{h_{11b}} = i_1. \qquad (14)$$



(15)

Now the transistor is drawn as a network, (17), and by definition:

$$h_{22b} = -\frac{i_2}{e_2}, \text{ where } i_1 = 0 \qquad (16)$$



(17)

The negative sign in (16) stems from network convention, where i_2 is shown flowing toward the network but actually flows in the opposite direction (outward from the network) for the common-base transistor circuit configuration.

Referring to (17), the input to the network can be shorted reducing e_1 to zero. By definition:

$$y_{22b} = -\frac{i_2}{e_2}, \text{ when } e_1 = 0. \qquad (18)$$

In equation (18), y_{22b} is one of the admittance family of parameters. If the open-circuited output admittance, h_{22b}, is subtracted from the short-circuited output admittance y_{22b}, the resultant output circuit admittance is that due to the current flowing in the short-circuited input circuit, or in other words, to the short-circuit input admittance. This relationship is shown in (19), where:

$$y_{22b} - h_{22b} = -\frac{i_2}{e_2} =$$

short-circuited input admittance appearing in the output (19)

$$\text{Now: } h_{21b} = -\alpha = \frac{i_2}{i_1}$$

and

$$h_{21b}\, i_1 = i_2. \qquad (20)$$

Substituting (20) in (19):

$$y_{22b} - h_{22b} = \frac{-h_{21b} i_1}{e_2} \qquad (21)$$

$$i_1 = \frac{e_1}{h_{11b}} \text{ as shown in} \qquad (15)$$

Substituting (15) in (21):

$$y_{22b} - h_{22b} = -\frac{h_{21b} e_1}{h_{11b} e_2} \qquad (22)$$

and

$$(y_{22b} - h_{22b}) = \frac{h_{11b}}{-h_{21b}} = \frac{e_1}{e_2} \qquad (23)$$

Equation (23) is true for the case when i_1 flows and e_1 is the generated feedback voltage which does not appear at the input terminals for the short-circuited case.

In measuring h parameters it is assumed that all measurements are performed at signal levels for which the transistor is a linear network.

If the short-circuit of (14) is removed, e_1 will appear across the input terminals and at the same time i_1 will go to zero. Therefore, equation (23) holds when i_1 is zero as well as when i_1 is flowing.

$$(y_{22b} - h_{22b}) \frac{h_{11b}}{-h_{21b}} = \frac{e_1}{e_2} = h_{12b},$$

when i_1 is zero.

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HARRY J. LANG, *Sales Manager*

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Radio and electronics are not new to Earl Lipscomb, the company's President; he has been active in the field since 1937 when he began as a consulting engineer. During World War II, he served in the training, production, and procurement phases of the Navy's electronic program. His five-year tour of duty included training at the Navy Radar School at M.I.T., and service with Navy Bureau of Personnel and the Navy Materiel Division. Upon completion of Navy service in 1947, he founded Earl Lipscomb Associates. In May of that year, Boonton Radio Corporation, recognizing the growing need for electronic instrumentation by customers throughout the Southwest, was one of the first companies to engage the services of the newly formed organization.

Earl Lipscomb

As the needs of the electronics industry expanded, so did the facilities and services of Earl Lipscomb Associates. The modern building, which houses the Dallas headquarters, comprises over 10,000 square feet of office and shop space and includes a complete clerical staff engaged in the distribution of technical information and processing of customer orders. Modern, well-equipped service laboratories at the Dallas and Houston locations provide calibration and repair service for all of the products manufactured by thirteen leading producers of precision electronic equipment. These facilities are staffed by six factory-trained service engineers.

A special group of five sales engineers, with specialized training in electronic measuring techniques, and two engineering trainees devote their entire effort to the solution of customer problems. This group is equipped to provide demonstrations of all equipment in the customer's laboratory; having at its disposal a unique "laboratory on wheels" known as the "Travelab". This mobile laboratory, which travels regularly throughout the Southwest, is completely equipped to provide operating demonstrations and serves to keep the customers abreast of the latest in instrument design and application.

It is the objective of Earl Lipscomb Associates to not only provide the customer with the proper instrumentation for his specific needs, but also to insure that these products continue to provide reliable service. For their constant effort toward this end, and for their record of efficient and faithful service to our customers, BRC extends a vote of thanks to Earl Lipscomb Associates.

The "Travelab", carries a full line of instruments to the customers' laboratory or plant.

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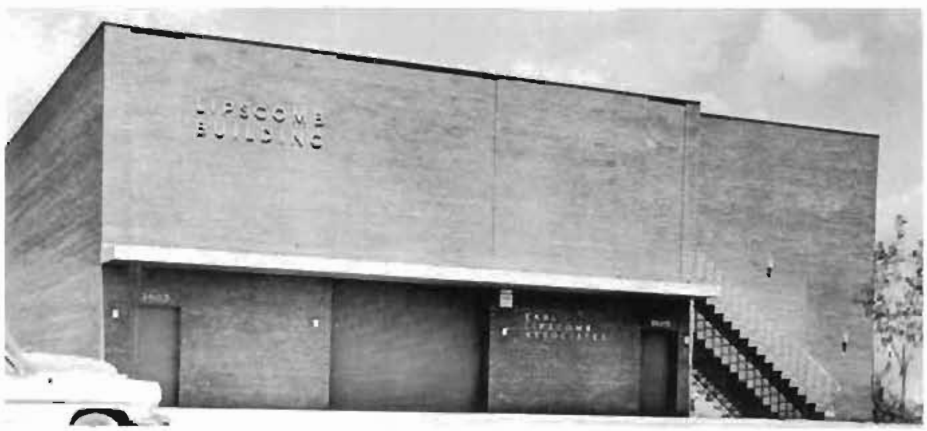

Dallas headquarters of Earl Lipscomb Associates

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APPOINTED SALES REPRESENTATIVE

We are pleased to announce that, effective July 1, the Gene French Company has been appointed BRC sales representative in the New Mexico, Utah, and Colorado area. The company maintains offices in Albuquerque and Denver and is fully equipped to handle sales, application engineering, and service for all BRC products.

"Gene", who has previously handled BRC instruments in New Mexico, is handling that area. "Hugh" Hilleary is heading the new Denver office. Please do not hesitate to call upon them for information or demonstrations.

Gene French

Hugh Hilleary

EDITOR'S NOTE

Amateur radio has given thrills and pleasures to countless thousands of persons the world over. Few people realize, however, that this favorite pastime is almost as old as the art itself. There were radio amateurs before the beginning of the present century; not too long, in fact, after Marconi astounded the world with his invention of wireless telegraphy. But amateur radio came into its own when private citizens discovered this means of personal communication with others and set about learning enough about "wireless" to build homemade stations. Its progress since those early days has been remarkable. In the first years, amateurs were stuck with 200 meters and could barely get out of their backyards. Today, with years of experimentation under the amateur belt, international DX is a reality and QSOs with countries all over the world are commonplace.

Personal communications between HAMS is only part of the amateur radio story. These "home stations" have posted a brilliant record of public service. Amateur cooperation has played an important part in the success of many an expedition and, in many cases, has been the only means of outside communication during several hundred storm, flood, and earthquake emergencies in this country. These public service endeavors were so successful in fact, that in 1938 the American Radio Relay League (ARRL) inaugurated a new emergency-preparedness program, registering personnel and equipment in its Emergency Corps and putting into effect a comprehensive program of cooperation with the Red Cross.

The HAM and amateur radio is constantly in the forefront of technical progress too. Amateur radio developments have come to represent valuable contributions to the art. During World War II, thousands of skilled amateurs helped to develop secret radio equipment for both Government and private laboratories. In the prewar years, technical progress by amateurs provided the keystone for the development of modern military communications equipment.

Modern radio owes a lot to these indefatigable amateurs for their contributions to the art. We are proud to number among their lot eight of BRC's employees.

L. O. Cook	ex - 8BRU
V. E. Hopler	K2LTH
G. A. Sanford	K2ARW
N. L. Riemenschneider	W2LKO
G. P. McCasland	K2RLK
N. L. Greendyk	ex - W2KSR
D. M. Terp	KN2JTN
J. P. Van Duyne	W2MLX

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Noise Limited Receiver Sensitivity Measurement Technique

John P. VanDuyne, *Engineering Manager*

Types of Noise

The fundamental limitation of noise on the useful amplification of a network has long been known. For the first several decades of the radio receiver design art, the noise which supplied this limitation was out of the control of the receiver engineer. Prior to 1940, most receivers in use operated below 30 mc, a region in which atmospheric and man-made electrical disturbances supplied the noise which limited practical receiver sensitivities.[1] Laboratory determination of the sensitivity of such receivers then consisted of measuring the input level for a predetermined output. Since the external noise levels mentioned above were usually much higher than the internal noise in well designed receivers, most sensitivity specifications were written in terms of microvolts input, behind a stated dummy antenna, for a given power output.

During the 1930's, as the useful high frequency communication spectrum pushed above 10 mc, the receiving tubes then in use were found wanting, in that they generated amounts of noise which exceeded the noise from external sources. As a result, new tube designs were developed to reduce tube shot noise and induced grid noise. Although tube designers were successful in keeping tube noise lower than the external noise at frequencies below 30 mc until 1940, the problem became acute enough that the I. R. E. took cognizance of internal receiver noise in its "1938 Standard for Measurement of Radio Receivers".[2]

As the communication industry pushed above 30 mc, especially as stimulated by the pressures of World War II, it was found that the external noise sources had dropped to negligible proportions, compared to internal receiver noise, above 100 mc. The need for greater and greater radar sensitivity, VHF communication range, and less "snowy" TV pictures rapidly pushed internal receiver noise down to the level where cosmic radio background radiation (popularly called cosmic "noise", due to its randomly fluctuating character) limits useful receiver sensitivity from 30 to 100 mc, or down to 10 mc in the absence of man-made and atmospheric noise. Recent advances in low-noise receivers and high-resolution antennas used by radio astronomers are resulting in distribution maps of this radiation.[3]

In spite of their widely different origins, cosmic noise and internal receiver thermal agitation voltages are similar in character and pose similar measurement problems. Man-made and atmospheric noise are very different, in that they have a discrete impulse nature and require different measurement and system evaluation technique. The reader is referred to A. S. A. Standards C63.2 and C63.3 for a discussion of their measurement. The term "noise" used in the following discussion refers to thermal agitation phenomena unless otherwise qualified.

Measurement Objectives

A general word should be said with regard to the effect of the objective of the measurement on the choice of measurement technique. These objectives fall in several categories, the most common of which may be stated as follows:

1. The comparision, on a uniform basis, of equipments of the same design, or from the same process.
2. The comparision of equipment of basically different design with similar desired preformance, for purposes of selecting the superior design.
3. The study of equipment performance with the intent of improving it by redesign.
4. The study of equipment performance for the purpose of learning more about the physical principles on which it operates, or to evaluate the extent to which the measured performance approaches the theoretical limit.

The techniques to be discussed require increasing degrees of skill and precision as the objective changes from (1) to (4) above.

Noise Limited Sensitivity Criteria

As previously mentioned, the reception of weak radio signals below 30 mc was limited by external noise. Hence, when measured under laboratory conditions, such receivers seemed "noiseless", since they had only sufficient amplification to produce rated output on signals supplied by a much noisier source than the laboratory signal generator. Therefore, the early concept of sensitivity was a specification of the input required, behind some specified network (dummy antenna) to produce a prescribed output. Long range communications operators found such receivers inadequate when used in quiet locations on well designed directional antenna systems. Consequently, amplifications were in-

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creased, and it soon became possible to provide full rated output with no input supplied, due to the receiver's internal noise sources. An early solution to this problem was to state the sensitivity, as previously noted, but qualified by a statement of the minimum signal-to-noise ratio tolerable at rated sensitivity. A more standardized form of this was introduced in the I. R. E. 1938 Standard for Measurements of Radio Receivers.[2]

Equivalent Noise Sideband Input

This standard introduced the concept of "Equivalent Noise Sideband Input" or "ensi" as it is usually abbreviated, for the measurement of broadcast receivers. Ensi is measured by supplying an unmodulated carrier, of a specified level (E_x), through an appropriate source impedance to the receiver under test and noting the output noise power (P_n). 30% 400 cps A. M. is then applied and the output 400 cps power (P_x) is measured with the aid of a bandpass filter to eliminate the noise power. Then,

$$(1)\ \text{ensi} = 0.3\ E_x \sqrt{P_n/P_x}$$

There are several possible sources of error in this measurement which must be eliminated or corrected if similar results are to be repeated at different locations (objectives 1 and 2) with different equipment, or if anything approaching an absolute performance measurement is desirable (objectives 3 and 4). First, the meter used to read P_n and P_x must be a true rms reading device, such as a thermocouple milliammeter. This meter must have reasonably constant sensitivity over the output pass band of the receiver being tested. This is necessary for a proper summation of the noise power spectrum to permit its proper comparison with the 400-cps demodulated sideband power. A second potential source of error lies in the linearity of the receiver being tested.

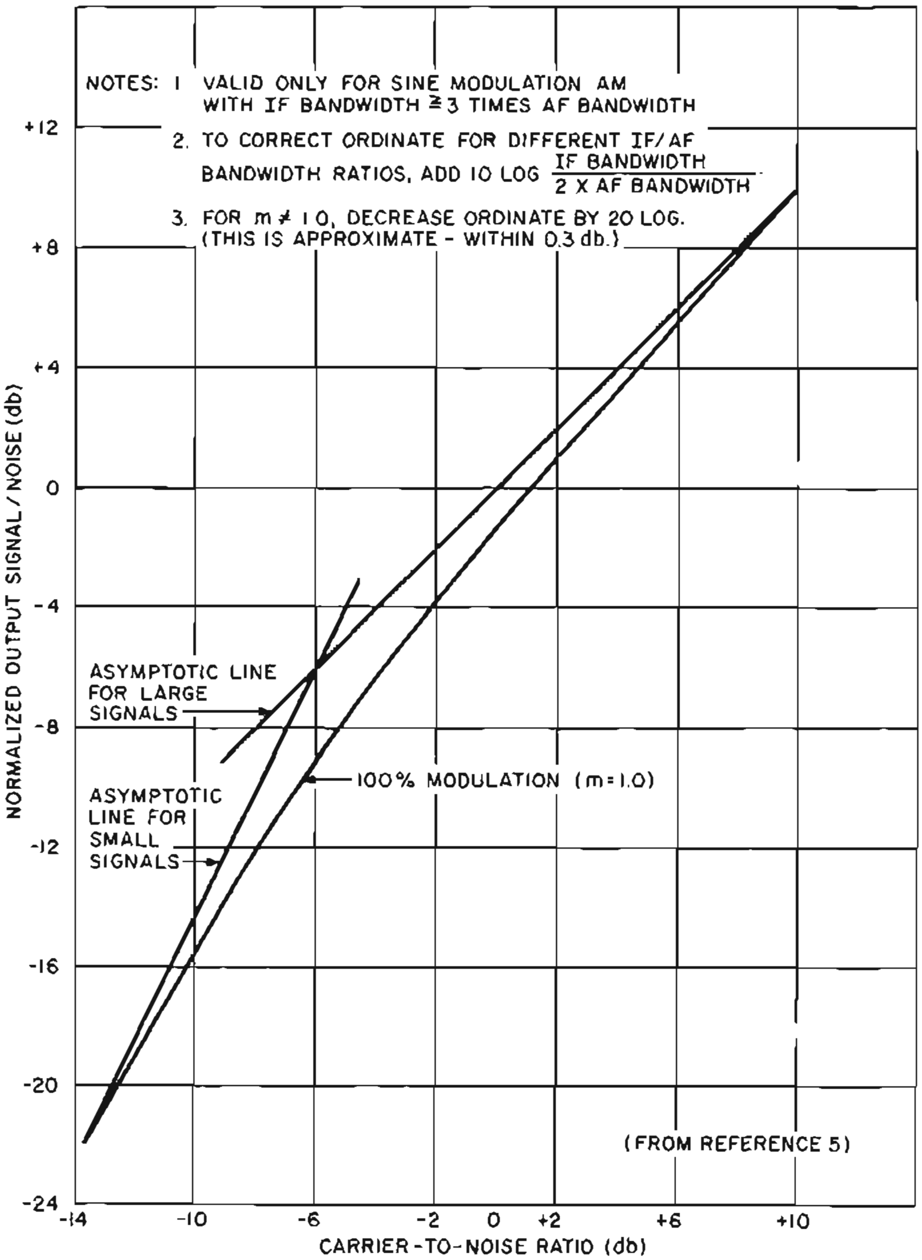


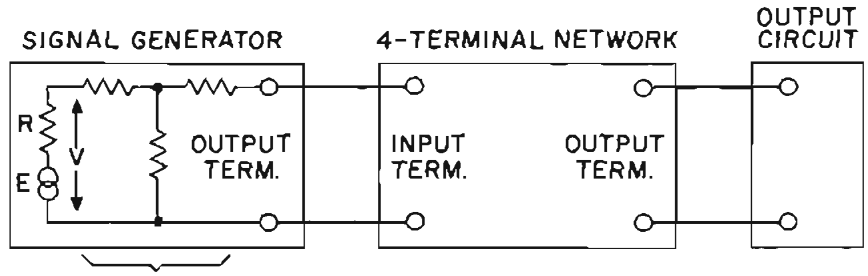
Figure 1. Universal Curve for Output Signal-to-Noise Ratio

Since thermal noise has a peak-to-rms ratio of about 13 db[1], the receiver (and the output meter) must not overload at voltage levels up to 4.5 times that of the demodulated sideband. The receiver detector is another element, the transfer linearity of which must be considered if the ensi measurement is to be of maximum value. The stated definition of ensi ignores detector non-linearities. This is justified if the usual high-level diode peak detector (so called "linear" detector) is used.

In the description of the ensi measurement, the carrier value E_x was mentioned as "specified". As a general rule, this value should be from 3 to 10 times the resulting ensi value. The 1938 I. R. E. Standard states that the measurement is to be made at a level of 5 μv if the "absolute sensitivity" of the receiver is 5 μv or less and at 50 μv if the sensitivity is between 5 and 50 μv. These precautions are necessary, due to the fact that the output signal-to-noise ratio of an A. M. detector is a non-linear function of the carrier-to-noise ratio. It is also a function of the I. F.-to-A. F. bandwidth ratio and it varies with the shape of the amplitude-vs-frequency response curves of these two portions of the receiver. In general, however, if the carrier-to-noise ratio into the detector exceeds 10 db, the errors in the output signal-to-noise ratio becomes negligible. Figure 1, taken from reference 4, illustrates this fact. For further detail on this matter, the reader should consult references 4, 5, and 6.

It should be noted, especially, in comparing different receiver designs for weak signal performance, that the effects of the second detector on output signal-to-noise ratio will not show up with a standard ensi measurement. For this reason the ensi measurement has given way in many specifications, to the measurement of sensitivity as the input, behind a specified source impedance, required to produce a stated output with a specified minimum signal-to-noise ratio. In evaluating equipments under such a specification, the second detector effects will usually be negligible if the output signal-to-noise ratio is 10 db or more and if the noise-free transfer characteristic of the detector is linear.

An additional proposal for a sensitivity figure which combines the concept of maximum gain with the limitations imposed by the receiver internal noise level has been suggested by J. M. Pettit.[7] In this proposal, the concept of "standard gain setting" is introduced. This is defined as the setting of the gain control which permits the delivery of a previously decided upon standard noise output to a specified load. This standard noise output must be specified for a given class of service and for specific equipments. For example, it might be specified as 0.5 milliwatts in 600 ohms as is typical for some communications equipment. The procedure for measurement is to connect the test equipment to the receiver, but with the signal generator adjusted for zero output. The receiver gain control is then adjusted for standard noise output. The output of the signal generator is then increased until "standard output" is obtained on the output indicator. The level from the generator is then noted as the receiver sensitivity. If standard noise output is not achieved, then a "maximum gain" or a "maximum sensitivity" criteria is used as previously discussed. This procedure is a formalization of



Attenuator of characteristic impedance R and attenuation >>1

$S_g \equiv$ Power available at Sig. Gen. Output terminals.

$$S_g = \frac{V^2}{R4}$$

$S \equiv$ Signal power available at the output terminals of the 4-terminal network.

$N \equiv$ Noise power available at the output terminals of the network.

$G \equiv S/S_g$ which depends on generator impedance but not on load connected to network. This is defined for midband frequency of network.

Figure 2a. Equivalent Circuit for Noise Figure Definition

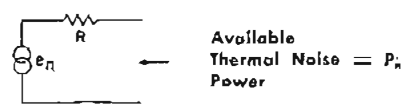
the previously mentioned general class of sensitivity specifications which state an input signal level for a specified signal-to-noise ratio.

The measurement precautions mentioned under the discussion of ensi apply to any sensitivity measurement involving the ratio of single-frequency power to the power in a noise spectrum.

Noise Figure

In an effort to work out a more basic measure of receiver performance when limited by random noise, several workers proposed rating the noise characteristics of a receiver, independent of its amplification. The specific proposal which has come into general use is that by D.O. North[8]. A later paper by Friis[9] developed a more rigorous general definition of Noise Figure. This allowed the concept to be applied to networks generally. In addition, Friis developed techniques for handling the noise performance of networks in cascade. This work permits calculation of a system noise figure from that of its components or vice versa.

Figure 2a shows the general circuit analyzed. The concept of available power is used in this analysis to avoid loss of generality due to dependence on the receiver input impedance or the load connected to the receiver output. Friis defines Noise Figure of a network (sometimes called Noise Factor and Excess Noise Ratio) as "the ratio of the available signal-to-noise ratio at the signal generator terminals to the available signal-to-noise ratio at its output terminals". From Figure 2b, we have the available signal-to-noise ratio of the generator as S_g/KTB and that for the network as S/N.



$$P_n = \frac{4KTRB}{4R} = KTB \text{ watts.}$$

e_n = equivalent Thermal noise voltage.
$e_n^2 = 4KTB$ where: $K = 1.38 \times 10^{-23}$.
Thus, the available signal-to-noise ratio for the above generator is S_g/KTB, and the network output available signal-to-noise ratio is S/N.
(From Fig. 1, p. 419 of Reference 9)

Figure 2b. Available Thermal Noise Power from a Resistor

If $F \equiv$ Noise Figure, we have by our statement above

$$(2) \qquad F = \frac{\frac{S_g}{KTB}}{\frac{S}{N}} = \frac{S_g}{S} \times \frac{N}{KTB};$$

$$(3) \qquad \text{but } G \equiv \frac{S}{S_g};$$

$$(4) \qquad \text{so } F = \frac{N}{GKTB};$$

(5) or $N = FGKTB$, which includes the amplified signal generator thermal noise power GKTB so the available output noise, due only to sources in the network, is (F-1) GKTB.

For simplicity, the usual measurement method is to adjust the attenuator A (Figure 2a) such that the output noise power is doubled by the generator signal. Under this condition, $S = N$ and from equation (2):

$$(6) \qquad F = \frac{S_g}{KTB} = \frac{V^2}{RAKTB},$$

$$\text{if } e^2_s = \frac{V^2}{A}$$

where e_s is in microvolts delivered to a load from an R ohm source — (implicit due to our use of the available power concept).

$K = 1.38 \times 10^{-23}$ (Boltzman's constant)

$B =$ bandwidth in kc

$T = 290°$ Kelvin (17°C) (for arithmetic convenience)

$R =$ generator source in ohms

$$F = \frac{e^2_s}{4RB} \times 10^6 \quad \text{(as a power ratio)}$$

$$(7) \qquad \text{or } F_{db} = 10 \log \frac{(e^2_s \times 10^6)}{4RB}.$$

It is important to note that the bandwidth B introduced in Figure 2b represents the bandwidth of an equivalent rectangular power pass band of gain G and an area equal to that under the actual power gain vs frequency curve of the device being tested. In mathematical notation,

$$(8) \qquad B = \frac{1}{G}\int_0^{\infty} G_f df$$

where $G_f =$ available power gain at frequency f.

$G =$ available power gain at the frequency of the CW measurement.

In a practical case, B may be determined by plotting the squared ordinates of the voltage gain vs frequency curve as a function of frequency, calculating the area of the resulting curve graphically, and dividing by the value of G at the specific frequency, within the passband, to which the generator in Figure 2a is tuned.

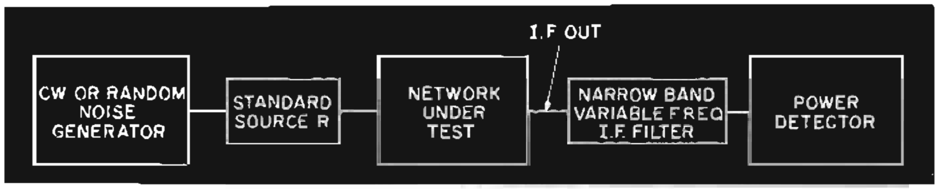


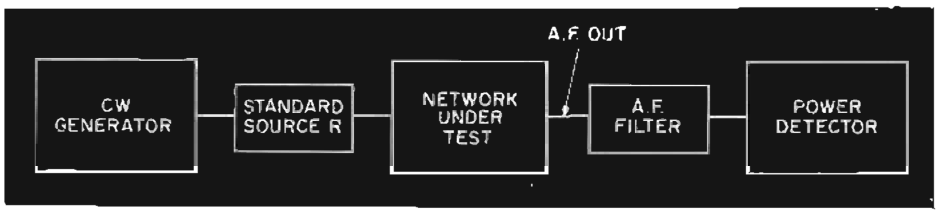
Figure 3. Single Frequency Noise Figure — I. F. Filter Method

Alternative to the use of a CW signal generator is the use of a random noise generator such as a temperature limited diode. This permits a simpler measurement to be made, without the need to determine B, but yields an answer which is actually the mean value of F. For purposes of receiver "front-end" evaluation, this is often sufficient, but for the evaluation of the noise figure of networks involving modulation or demodulation, or cascaded networks of different bandwidths, or networks in which the source resistance is a variable with frequency, the measurement of "single frequency" noise figure is often necessary if proper evaluation is to be made.

I. F. noise spectrum is much wider than ½ of the I. F. response (Reference 6), thus when using the I. F. filter method of Figure 3, a much narrower than expected I. F. filter is required. By the use of a CW generator and A. F. filter (Figure 4), useful data on the variation of single frequency noise figure through a network pass band can be derived rather simply. In solving equation 8 it should be noted that the value for G is that which corresponds to the frequency of measurement in the pass band. The resulting value of B is used in equation (7) to solve for the single frequency noise figure.

There are several types of measurement for which the CW noise figure measurement is necessary. A typical one is the case of a receiver with a selective filter ahead of the first noise source which is comparable to the I. F. selectivity. In this case, a temperature limited diode measurement with a wide-band power detector may indicate a poor noise figure (mean noise figure). Measurement with a single frequency method may disclose a good band center noise figure, but with rapid deterioration toward the band edges, which indicates an incorrect choice of R. F. selectivity, impedance match compromise, or too wide a coupling circuit between the first and second stages of amplification if the second stage contributes appreciable noise.

Some interesting data results from solving equation (7) for e_s.



Note: the standardization of the output power detector for a 3db output power increase should be accomplished by introducing a known 3db change in IF gain.

Figure 4. Single Frequency Noise Figure — A. F. Filter Method

The single frequency noise figure concept as distinct from the mean noise figure (which is what has been described above) is thoroughly discussed mathematically in reference (10). The classical method of measurement is to insert a variable center frequency narrow-band filter between the network being measured and the power detector. The single frequency noise figures which result from measurement through each center frequency of the narrow filter are then weighted by their relative gains and averaged. See Figure 3 for this set-up. Since this technique is often inconvenient, a frequently useable approximation results from the use of a CW generator, a narrow-band A. F. filter, and a power detector as in Figure 4. The A. F. noise spectrum resulting from wide-band demodulation of an

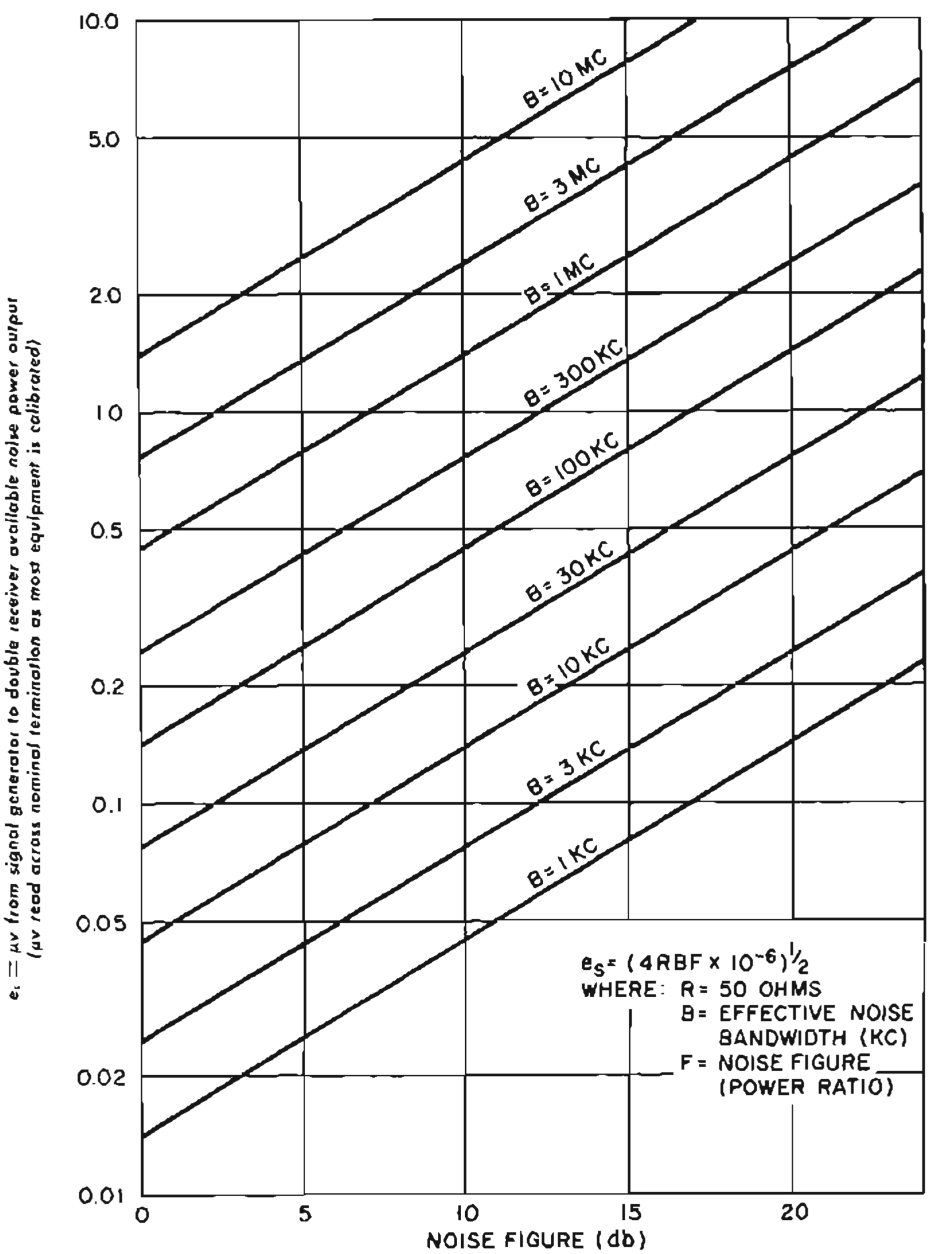


Figure 5. Values of e_s for Various Noise Figures and Bandwidths

(9) $e_s = \sqrt{4RB \times 10^{-6} \text{ antilog } F_{db}}$.

Figure 5 plots these results for several bandwidths and for R = 50 ohms. Note, for example, that with a noise figure of 3 db and a bandwidth of 3 kc, the input signal, e_s, required to double the available noise power output is 0.035 microvolts. This is approximately true for a voice frequency VHF A. M. communications receiver of good design. Obviously, a generator of extremely low leakage and good low-level calibration is required to make this measurement. An external 40-db pad is usually required to reduce the usual 1.0 —μv minimum level to the order of 0.01 μv.

There are several sources of error in noise figure measurements. Typical are those associated with the super-heterodyne selectivity of the receiver under test. In general, the I. F. and image frequency rejections must exceed 10 db if significant error is to be avoided. It is interesting to note that such spurious responses will give a pessimistic (high) noise figure if a CW generator is used for the measurement, but an optimistic reading if a broad-noise spectrum generator is used. This is due to the fact that in the former case, the receiver is exposed to unwanted noise generators (I. F. or image) which have no signal counterpart.

Another source of error may occur in the determination of the doubled output noise power. This is not as critical as in the case of the ensi measurement, since the detector output with a CW signal at a 3-db signal-to-noise ratio is largely composed of noise sidebands. The best method is to calibrate the I. F. amplifier for a 3-db gain differential, so that the detector operating point stays the same. If the device permits, insertion of a 3-db attenuator is best, if it can be done without an accompanying change in bandpass. Alternatively, the gain control can be calibrated for a 3-db gain change. Either of these two methods can be used to calibrate the detector characteristic to answer the question of its power response. It should be noted that any attempt to use a modulated signal for noise figure measurement is beset with all the errors of the ensi method and should be avoided.

A precaution which is important to all sensitivity measurements, but especially so in the case of noise figure, is the need to accurately control the generator source impedance and noise temperature. In the absence of contrary system requirements, a resistance should be used equal to the nominal transmission line impedance for which the receiver is designed and corrected to 290° K from the actual temperature.

Conclusion

Sensitivity measurements made on receivers which have sufficiently low internal noise to detect thermal noise in the source are among the most exacting which can be made on a radio receiver. Good results require good equipment, careful set-up, and careful experimental technique. Most important, however, is a thorough understanding of the theoretical basis of the measurement and the use to which the results are to be put. The precautions and suggestions in the preceding discussion have all been thoroughly proven by extensive use in the laboratory and are offered to the reader as a guide to better experimental procedure.

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More About Transistor Measurements With The HF-VHF Bridge

GEORGE P. McCASLAND, *Sales Engineer*

The response to the transistor measurement article, which appeared in Notebook No. 19, was gratifying indeed. Many helpful suggestions and criticisms were received from Notebook readers. Your author hopes that the information presented here will serve to correct and clarify several points in question.

h_{21b} and h_{12b} Formulas

The derivation of h_{21b} as explained in the appendix to the original article is not entirely correct. Diagram 4 on page 5 of Notebook No. 19 shows e_1 to be positive toward the emitter terminal of the transistor. In diagram 7, on the other hand, e_1 is shown positive toward the base terminal. Actually, current flow shown in both cases requires that e_1 be positive toward the emitter terminal. Therefore, the derivation of the h_{21b} formula should have been given as:

$$h_{11e} = \frac{-e_1}{i_1} \text{ from diagram (7)}$$

$$i_1 = -(1-\alpha)\, i_c \text{ from diagram (7)}$$

$$h_{11e} = \frac{-e_1}{-(1-\alpha)\, i_c} = \frac{e_1}{(1-\alpha)\, i_c}$$

$$\text{Now: } h_{11b} = \frac{e_1}{i_c} \text{ from diagram (4)}$$

$$h_{11e} = \frac{h_{11b}}{1-\alpha}$$

$$-\alpha = \frac{h_{11b}}{h_{11e}} - 1 = h_{21b}$$

In deriving the formulas for h_{21b} and h_{12b}, certain approximations were made which were not specifically mentioned in the original article. Formal network theory shows these approximations. For example, from the table of Matrixes of 3-Terminal Networks on page 506 of *Reference Data for Radio Engineers*, by I.T.T. we find that:

$$h_{11e} = \frac{h_{11b}}{d}$$

From page 503 of the same book:

$$d = h_{11}\, h_{22} - h_{12}\, h_{21} - h_{12} + h_{21} + 1$$
$$\Delta^h = h_{11}\, h_{22} - h_{12}\, h_{21},$$

and

for junction transistors:

$$\Delta^h << h_{21}$$
$$h_{12} << 1$$
$$d \approx 1 + h_{21}.$$

Therefore:

$$h_{11e} = \frac{h_{11b}}{1 + h_{21b}}.$$

$$h_{21b} = \frac{h_{11b}}{h_{11e}} - 1 = -\alpha.$$

Converting RX Meter Readings to Admittance and Impedance

The RX Meter directly reads out resistance in parallel with a capacitance ($+C_p$) or an inductance ($-C_p$). If RX Meter readings are converted to admittance, $+C_p$ converts to a positive susceptance and $-C_p$ converts to a negative susceptance. When RX Meter readings are converted to impedance, $+C_p$ converts to a negative reactance and $-C_p$ converts to a positive reactance. These changes of sign must be remembered when using the series-parallel conversion chart. Considering these changes in sign then, correct rectangular Z's in Table IV on page 4 of Notebook No. 19 are 144—j52 and 65+j66 while correct polar Z's are $153\angle{-20^\circ}$ and $93\angle{45^\circ}$. The signs and j's preceding C_p (ohms) in the same table are somewhat misleading and should therefore be disregarded.

Calculation of h_{21b} and α

Using the corrected h_{21b} formula and corrected values from Table IV, the new sample calculation of h_{21b} should read as follows:

$$h_{21b} = -\alpha = \frac{h_{11b}}{h_{11e}} - 1$$

$$= \frac{93\angle{45^\circ}}{153\angle{-20^\circ}} - 1 = .61\angle{65^\circ} - 1$$

$$= .26 + j\,.55 - 1$$

$$= -.73 + j\,.55$$

$$h_{21b} = -\alpha = .93\angle{143^\circ}$$

$$\alpha = .93\angle{-37^\circ}$$

MEET OUR REPRESENTATIVES

BIVINS AND CALDWELL

HARRY J. LANG, *Sales Manager*

Bivins and Caldwell, Inc. was founded by John F. Bivins and David J. Caldwell shortly after their return to High Point, North Carolina at the close of World War II, in the belief that electronics and radio communications would play an increasingly important role in the future industrial development of the South. The specific need for a technical group to provide local customers with sales and application engineering services on complex electronic equipment was apparent and the partners formed their organization to represent leading manufacturers of communications and electronic laboratory test equipment.

Both partners, by virtue of their backgrounds, brought a wealth of specialized experience to the new company. John Bivins majored in physics and engineering at Duke University and had been in the radio and communications business in High Point for nearly ten years prior to World War II. Dave Caldwell majored in physics and engineering at Davidson, taught physics at that school for two years after graduation, and later held posts in the production and cost accounting fields. During the War, Dave Caldwell served in the Planning Section (G3) of the Army, and John Bivins was employed as a special engineer with the Navy Department, dealing with planning and supervision of Radar installations on Naval vessels.

Industrial expansion throughout the South during the past eight years directly confirmed the early beliefs of Bivins and Caldwell and also brought about a decisive change in the Bivins and Caldwell organization. While broadcasting equipment accounted for the major portion of their business up until approximately 1949, the rapid growth of electronic manufacturers and related industries, created increasing demands for specialized services in the application of electronic instrumentation and the organization now handles precision electronic laboratory equipment exclusively. Bivins and Caldwell joined BRC in 1952 and has continuously handled our products since that time.

With increased business activity came the requirement for expansion of the company's personnel and facilities. C. M. Smith, Jr. joined the expanding company in 1950. "Smitty" is an engineering graduate of North Carolina State College with extensive communications engineering experience. During World War II he served as an Electronics Officer with the U. S. Navy and was an instructor at the famed M.I.T. Radar School in Boston, Mass. Later additions to the staff have increased total personnel to over 21; 9 of which are field engineers who concentrate exclusively on customer problems.

Headquarters of Bivins and Caldwell, Inc. in High Point, N. C.

The company's facilities have also been increased, with branch offices in Atlanta, Georgia, and Orlando, Florida. The new headquarters building in High Point is one of the most modern and best arranged office facilities in the area. The Service Department here, under the direction of Mr. Robert L. Moore, has complete repair and recalibration facilities and is well equipped to service all products that are currently handled. All offices are equipped

Shown in conference are, left-to-right,
John Bivins, Douglas Severance, David Caldwell, and C. M. Smith, Jr.

with TWX service for efficient communication with all factories. The company also maintains a fleet of eight vehicles for use by field engineers. The Bivins and Caldwell Fall Road Show, which runs for about four weeks, covers some 4,000 miles and is an annual event attended by several thousand engineers in the local area.

Bivins and Caldwell believes that their organization must operate on the premise of fundamental engineering integrity. Their basic objective is to assist customers in solving engineering problems and to recommend the best available test equipment and techniques for the particular job to be done. We at BRC proudly salute Bivins and Caldwell for their faithful service to our many valued customers throughout the South.

EDITOR'S NOTE

John P. Van Duyne Appointed Engineering Manager at BRC

The appointment of John P. VanDuyne as Engineering Manager, effective August 18, 1958, has been announced by Dr. George A. Downsbrough, President. Mr. Van Duyne comes to BRC with 15 years of experience in the fields of engineering and electronics. The major part of his experience has been in the development and production of electronic instruments.

After receiving his Bachelor of Science degree in Electrical Engineering from Rensselaer Polytechnic Institute of Troy, New York in December 1943, Mr. Van Duyne served with the U. S. Signal Corps where he instructed in radar and radio relay techniques. During his last year of service, he was engaged in the design of Radio Countermeasures epuipment in the Coles Laboratory at Fort Monmouth, New Jersey.

Following his discharge from the Signal Corps in 1946, Mr. Van Duyne joined the Measurements Corporation of Boonton, N. J. as Project Engineer and was engaged in the development of signal generators.

In August 1948, he became associated with the Allen B. DuMont Laboratories, East Paterson, New Jersey, serving successively as Senior Engineer, Section Head of the Advanced Development Section, and Section Head of the Circuit Design and Development Section.

From March 1953 until he joined BRC in August, 1958, Mr. Van Duyne held posts with the Westinghouse Electric Corporation in Metuchen, New Jersey. He served as Engineering Section Manager and was engaged in the design of color television receivers until January 1956 when he was appointed Manager of TV Engineering. He served in the latter capacity until he joined BRC.

An active radio "ham", Mr. Duyne is also a member of Tau Beta Pi, Eta Kappa Nu, Sigma Chi and the Institute of Radio Engineers.

JOHN P. VanDUYNE

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A General Purpose Precision Signal Generator

CHARLES G. GORSS, *Development Engineer*

Design Considerations

The primary purpose of a signal generator is to simulate, acurately, some part or parts of a transmission system which are not conveniently operated at a test area or in a laboratory. Most important among the design considerations of such a device are its size, the precision with which it simulates test signals, and the stability of the simulated signals. Because test and laboratory space is usually limited, the signal generator is required to cover, in one small package, a band of frequencies wide enough to test an entire system, usually many times the size of the signal generator itself. As electronic systems have become more precise, the precision requirements placed upon the signal generator designed for use with these systems have become more stringent. A natural companion to precision is stability: the generated frequency, in particular, must not vary under the influence of the power line or amplitude modulation. In simulating weak signals into a high-sensitivity receiver, it must be possible to set the output of the signal generator to provide signals as low as 0.1μv with the knowledge that the results are not being clouded by leakage from the generator enclosure. Therefore, the design of the enclosure is very important along with all of the circuit design considerations.

Oscillator

The range chosen for this design is 10 to 500mc, the area of most intensive use in equipment development. Covering such a range with a single oscillator, implies that some changes in the parameters of the frequency determining elements will have to be made at periodic intervals through the range. It is impractical to imagine that varying one element, or varying both L and C for that matter, could possibly permit coverage of the 10 to 500-mc range in one band. At lower frequencies one can often choose from several different coils with a selector switch, but in the UHF range, leads and connections must be short, precluding the use of simple switching.

Figure 1. Type 225-A Signal Generator

In common use today is the concept of the turret, a device which actually removes the inductive element of the resonant circuit and replaces it with another element. This would seem to suit our purpose, because in such a device, lead length can be controlled. However, positioning must be very accurate and very stable. Providing a means for contacting these coils as they come into position requires careful consideration. The contacts must have low inductance and stray capacitance as well as a stable

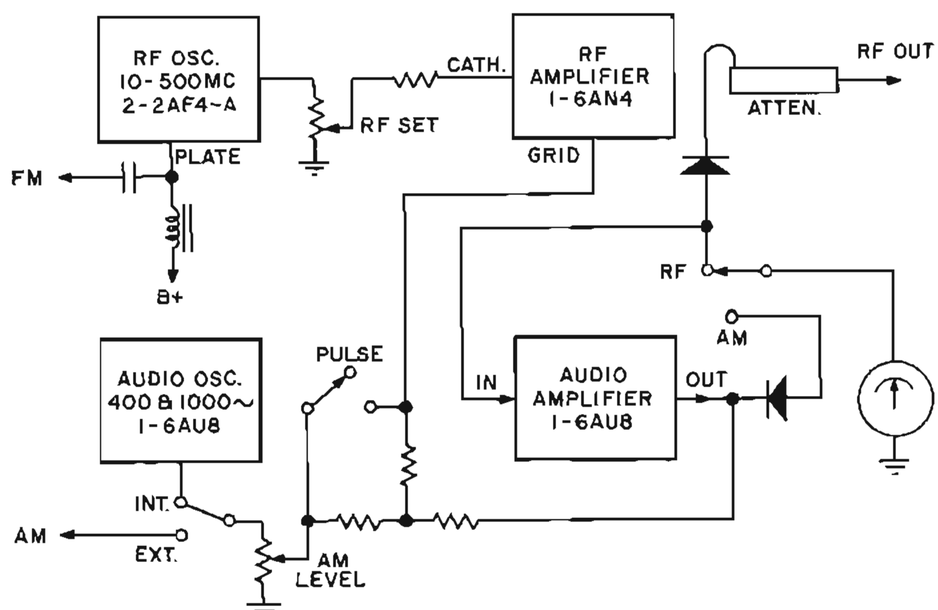

Figure 2. Functional Block Diagram

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low resistance. The physical structure of the turret is very important also, as any tendency of the basic structure to change dimensions or warp would result in frequency instability.

Use of the turret imposes certain limitations on the actual oscillator circuitry. The preferred circuit is one which employs a minimum number of moving contacts. It would be desirable then to design an oscillator with a feedback network which could be fixed for all frequencies so that additional switch contacts would not be required. A circuit which fits the turret requirements quite well, with a minimum number of contacts in the frequency determining network, is shown in Figure 3. The feedback is accomplished with a capacitive divider from one plate to the opposite grid, using the grid capacitance of the tube to cathode together with a fixed mounted capacitor from the other plate. This two-tube oscillator is particularly adaptable to our requirements for several reasons. First, it gives more power than a single tube of the 2AF4-A class; an important factor because good isolation from the modulator will be a requirement of the overall system and the more power there is to dissipate in isolation, the less reaction there will be. Second, the feedback is simple and fixed. Third, the two-tube oscillator works very well with a split-stator capacitor which requires no wiping contacts. This is important because wiping contacts on an oscillator capacitor would introduce noise and instability. In this oscillator, the center of the tank is at ground potential and therefore the rotor of the capacitor is also at ground potential for RF frequencies. With this arrangement the capacitance to ground of the capacitor drive is noncritical. Since the center of the oscillator coil is also roughly at the neutral or ground plane; plate power can be injected at this point from a common supply ring on the turret. This ring may be a simple slip ring rather than a switchable contact. Actually the oscillator turret is so constructed that the center of each coil is permanently tied back to this common slip ring through individual 100-ohm resistors. These resistors serve to break up undesirable RF paths, but do not introduce any appreciable plate voltage or radio frequency loss.

Coupling from the oscillator is accomplished by a pickup coil wrapped on the same form as the oscillator tank. Its output is picked up by two wiping contacts (similar to the contacts in the tank circuit) on the side of the turret.

Mechanical considerations in this part of the circuit have been very carefully thought out. The contact buttons on the turret are of coin silver and the mating spring fingers are of beryllium copper with a rolled-on coin silver overlay of 0.0025-inch thickness. (Silver plating can not be depended upon to withstand wear.) The turret itself is cast in an unmodified Epoxy Resin made by CIBA known as Araldite 6060 casting resin. This material has a reasonably low coefficient of expansion (50 ppm/°C and contains no filler material. The result is a very stable casting with no internal stresses and good machinability. Araldite resin is used to cement the silver button contacts into the casting. The circuit itself is mounted on a silver-plated brass chassis in a way which minimizes lead lengths and maintains the fundamental circuit symmetry. The basic enclosure tying the entire assembly together is a heavy aluminum casting mounted on a 1/4" silver-plated aluminum base plate. Positioning of the turret is accomplished by means of a stainless steel shaft which mounts a heavy hardened steel detent plate. The detent plate is restrained by an arm and roller assembly which is substantially spring loaded for positive positioning.

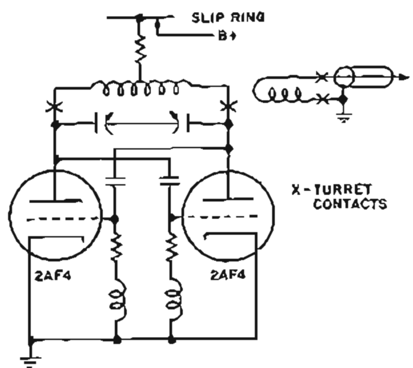

Figure 3. Oscillator Circuit

Modulator

There are many ways to modulate an RF signal once it is produced and it was necessary to evaluate these various methods in order to make the wisest choice. It was thought, at first, that it would be best if the modulator did not require tuning. This would immediately simplify the job by eliminating one tuning capacitor. It developed however, that this approach would create a problem in the output system. The piston attenuator in the output system operates with a rather large insertion loss; around 30 db over our frequency range. What is more, at least 20 db of attenuation is required between the modulator and the oscillator to prevent spurious frequency modulation as a result of amplitude modulation. An untuned modulator would impose an additional insertion loss. The output requirements of 0.1 volts is db below a milliwatt. With 20 db required for isolation, 10 db for modulation, and 30 db for the attenuator, there is a total loss of 60 db between the oscillator and the output. This means that a ridiculously high figure of roughly 500 watts would be required to provide the desired output. The oscillator discussed previously, puts out about 500 milliwatts lightly loaded.

Diode Modulator

Another approach worthy of consideration is some form of diode modulator following the piston attenuator. There seem to be two objections to this approach. First, the noise generated in the diodes would be objectionable at low levels such as 0.1μv. Second, the high levels of voltage necessarily applied in order to produce a 0.1-volt output would necessitate operation of the diodes in a more linear operating range and thereby destroy the modulation capabilities. Typical low-level diode modulators of this type have about 50K μv maximum input and about 10 db insertion loss, meaning that the maximum modulated RF output would be around 15K μv. An advantage of this type modulator would be that the piston attenuator could couple directly to the oscillator and eliminate the second tuned stage. However, the poor low-output capabilities make it unsuitable for our purpose.

Tuned Grounded-Grid Trode Amplifier

If the modulator could be designed to provide a 20-db gain, instead of the 10-db loss introduced by an untuned modulator, there would be only a 30-db loss beyond the oscillator and the oscillator output requirement would need to be only 500 milliwatts. A tuned grounded-grid amplifier, utilizing a 6AN4 UHF triode, will provide this 20-db gain. The gain drops as 500 mc is approached, but improved attenuator coupling at the higher frequencies compensates for this effect.

There are several ways to modulate a tuned grounded-grid triode amplifier; by means of the plate, grid, or cathode. Both plate and cathode modulation require power. Grid modulation does not require power, but requires careful selection of operating point for reasonable linearity. Since this stage would not be operating as a class C amplifier, linear plate modulation would not be possible. Therefore, in the interest of simple low-powered amplitude modulation, a grid-modulation system was chosen.

In such a system, the grid must be well grounded for the RF signals but not for audio signals. This requires that a suitable capacitor be placed from grid to ground. A common failing of a high-frequency, grounded-grid stage is instability caused by the existance of inductance in the grid circuit. This results in positive feedback and possibly oscillation. Therefore, the grid capacitor selected must be a very low-inductance device.

The RF ground of the stage is established by a large silver-plated brass shield closely contoured to the tube socket and passing directly through the center of the tube socket and the grid pins which are located 180° apart. This shield not only establishes ground but shields the plate from the cathode. The grid leads are soldered to a sheet of silver-plated copper which covers the entire surface of the shield. A thin sheet of reconstituted mica separates the copper sheet from the shield. The copper sheet, covers both sides of the shield and acts as a very low-inductance bypass capacitor of 1500 $\mu\mu f$. This arrangement imposes a maximum RF reactance of 10 ohms at 10 mc; the lowest generated frequency. The maximum audio frequency to be passed by the modulator is 20 kc. At this frequency, the bypass capacitor is not less than 5000 ohms. To enable good pulse modulation, this capacitance would require 15 ma in order to permit the grid voltage to rise from 0 to 10 volts in 1 μs. The 15-ma current is based on the fact that a 10-volt pulse will cut off the amplifier. This is easily accomplished in the average pulse generator as long as there is no large series impedance between the generator and the grid.

Maximum isolation between the oscillator and the amplifier was given previously as a criterion for minimizing spurious frequency modulation. For this reason, as well as in the interest of maintaining the operating point of the modulation at an optimum level, the output level control was incorporated in the coupling between the oscillator and the modulator. This is merely a variable resistive voltage divider into which the oscillator output is fed. Only enough voltage to drive the modulated amplifier to the level which produces 0.1-volt output is taken from the divider. If a fixed attenuator were used it would have to be made small enough to insure full output at all frequencies under the limits imposed by tube parameter variations. The amplifier gain would then have to be reduced under these conditions to match this fixed attenuator. The result would be that, under many conditions, the stage gain would be reduced unnecessarily to match the fixed attenuator and, although the resulting spurious FM would be within the advertised specifications, it would not be as low as the tubes were capable of making it if they were allowed to perform at their optimum capabilities.

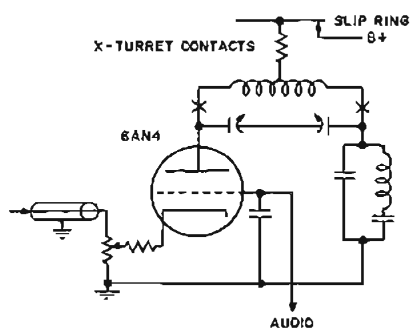

Figure 4. Modulator Circuit

Another reason for not changing the anode potentials in some uncontrolled manner is that, as previously stated, control-grid modulation requires careful selection of the stage operating point. The bias level is quite critical if low distortion is desired. Because the attenuator coupling decreases with decreasing frequency, the average level of drive to the final stage increases as the frequency decreases. The impedance of the modulator's tuned load also is higher for each lower band because the same tuning capacitor is used on all six bands. These factors result in a different optimum grid bias being required for each range. This is accomplished by means of a switch, coupled to the range knob, which selects the proper bias for each range. The distortion is further reduced by overall inverse feedback which will be described further on in this paper.

To maintain maximum isolation between the amplifier and oscillator, the amplifier is housed in a separately shielded casting, very similar to the oscillator casting. The energy passing between the two castings is fed through a coaxial cable which is enclosed in copper tubing to prevent leakage. This coaxial cable couples directly into the variable resistive attenuator which in turn couples to the cathode of the amplifier. This construction prevents stray fields or circulating currents which might cause sudden unpredictable increases in spurious FM at discrete frequencies.

The grounded-grid amplifier circuit tank (Figure 4) is similar to a push-pull tank except that one tube has been replaced with a reactive network. A true push-pull stage utilizes a complex driver transformer and would furnish more power than is required. This one tube arrangement was used in order that the amplifier circuitry would be similar in design to the oscillator, with the same design of turret and tuning capacitor, so that the two circuits would track naturally. The turret is the same as the turret used in the oscillator except that there is no pickup winding. (The attenuator pickup coil couples directly to the tank coil.)

The tuning capacitor is the same capacitor used in the oscillator except that several plate sections have been omitted. The oscillator and amplifier tanks are necessarily of slightly different design because, in the oscillator, the feedback capacitors add to the minimum tank capacitance. In the amplifier, the connections from the turret to the tuning capacitor are slightly longer to accomodate the attenuator. Therefore, this tank has a somewhat higher inductance and a lower capacitance than the oscillator tank. In order to assure reasonable tracking with these unequal tank parameters, fewer plate sections were used in the amplifier tuning capacitor. The reduced $\triangle C$ in the amplifier, in combination with the lower residual capacitance, results in a frequency range which closely tracks the oscillator.

The turret is cast with one flat side

so that it can be positioned closely to the flat attenuator pickup coil. Between the attenuator tube and output tank, on either side of the plane occupied by the attenuator pickup coil, are two parallel monitor wires about $\frac{3}{16}$ inch apart. One end of each of these is grounded and the other ends are tied to a IN82A diode monitor. This monitor system, therefore, intercepts the same field that enters the piston attenuator tube and performs the monitor function for all ranges. In a narrow-range, one-band generator, output monitoring is accomplished by feeding some energy from the tank through a small coupling capacitor. This system provides more usable energy, but would not be satisfactory in a wide-range, multi-banded unit. The two pickup conductors also serve as a Faraday shield, allowing only the TE_{11} mode to propagate into the tube. Any mode which is propagated into a wave guide significantly below its cutoff frequency decays at a rate which is exactly logrithmic with distance along the waveguide. This rate is exactly related to guide diameter in a circular waveguide. However, there are various modes which can be introduced into a waveguide which have differing rates of decay. The presence of more than one mode would tend to distort the ideal attenuation law. The attenuator is based on the TE_{11} mode because this mode decays at a rate which is less than any of the other propagation modes. The TM_{01} mode is shorted out by conductors arranged in the same manner as the monitor loop, because the attenuation rate of the TM_{01} mode is only 4.9 db per radius more than the TE_{01} mode and could cause errors in the attenuator output. The TE_{02} mode, at an additional 17.3 db per radius, is also an annoyance. This problem is solved by arranging the pickup and tank coils symetrically around the axis of the tube.

The attenuator pickup coil itself is a single loop of wire in the same plane as the center turns of the tank coil. A 50-ohm carbon film resistor in series with the loop provides a 50-ohm source impedance. At the point where the loop connects to the output coaxial cable, there is an impedance compensating circuit composed of a 50-ohm resistor and a capacitor in series to ground. This tends to draw current of a leading phase when the loop phase is lagging, resulting in the maintenance of a good low VSWR source impedance from the attenuator output. This results in a VSWR of less than 1.2.

In order to accurately measure the amplitude modulation percentage, as well as the RF output, the IN82A monitor diode is by-passed for RF only. Amplitude modulation at an audio rate remains as an audio voltage imposed on the diode dc output. To monitor RF, diode output is fed directly to a 20 μa meter on the panel through suitable calibrating resistors. For AM monitoring, the diode output is connected to an ac-coupled amplifier which builds up the audio envelope, then feeds it to a cathode follower. The cathode follower in turn drives a diode voltmeter which is fully bypassed for frequencies as low as 20 cps. This dc output is then switched to the same 20 μa meter through suitable calibrating resistors. The meter is marked both in % AM and an RF Calibrate position. It normally reads RF but by means of a momentary contact switch can be made to read % AM.

The output of this AM metering amplifier is used in another related manner. A certain percentage of the voltage from the cathode follower outputs returned out of phase with the incoming modulation voltage. This tends to further reduce the AM distortion and provdes a high degree of stability for the entire modulation system. Since this places a resistive network between the input terminals and the RF amplifier grid which would tend to slow down pulses fed into this point, a switch at the full-clockwise position of the AM LEVEL control removes the inverse feedback, providing direct connection from the input terminals to the RF amplifier control grid for pulse modulation.

In the pulse position, the instrument continues to operate as described. A 10-volt negative pulse will turn the amplifier off. The amplitude modulation terminals are dc coupled to the grid of the RF amplifier and thus it is not desirable to swing this point in a positive direction. All that lies between the AM posts and the grid is an RF filter which prevents RF leakage. This had to be appropriately damped to prevent ringing or extreme overshoot and therefore limits the minimum rise time to 2 μs.

Frequency Modulation

Frequency modulation has been included in this instrument in its most elementary form. Means have been provided for amplitude modulating the plate of the oscillator from an external post. This provides low deviation frequency modulation which, though uncalibrated, is somewhat predictable in magnitude and will be useful in the range above 100 mc where sufficient deviation for the narrow-band FM communications channels is present.

Shielding

The instrument has been thoroughly shielded against RF leakage. Where flat cover plates engage the RF shield castings, a mating tongue and groove joint lined with silver-plated brass mesh assures perfect sealing. The aluminum cover plates are silver plated and join the casting in a similar manner. Where shafts protrude from the enclosure, double circular wiper fingers are used, one over the other. Every joint is carefully sealed by some resilient, highly conductive device which will retain high pressure and good electrical contact. The RF filters employ very low-inductance discoidal ceramic capacitors which have a resonant frequency well above the range at which they are used in the generator. The series elements in the filter are toroidal coils wound on ferrite cores, providing the maximum series loss in the smallest package possible.

Power Requirements

The power requirements of the instrument have been kept low; in the order of 70 watts. This gives the instrument a good degree of stability due to the freedom from excessive heating. What is more, it has permitted the use of a very effective, but simple, power supply. The power transformer is a resonant circuit type, regulating transformer which provides excellent stabilization of plate and heater voltages. In addition, the dc output to the oscillator plate is gas-discharge tube regulated and the filaments are regulated by a hot-wire, series-regulating ballast. This regulation results in a frequency stability-vs-line voltage of much better than 0.001% total frequency change for a 5-volt line shift.

Internal Modulation

The instrument has an internal modulating oscillator operating at 400 or 100 cycles, as well as provision for external amplitude modulation.

Frequency Variation Controls

It has been previously stated that the oscillator and amplifier were designed in such a way as to track together as the frequency is changed. However, some adjustment is necessary to peak the amplifier output for optimum performance at a specific frequency. This is accomplished by means of a small trimmer knob which is coaxial with the large coarse frequency control knob. This control provides differential motion between the oscillator and the amplifier tuning capacitors. Friction in the system is such that the oscillator shaft does not turn when this knob is operated, permitting tuning of the amplifier through a 6 to 1 reduction with negligible frequency change. The shafts are coupled together with a spring-loaded, phosphor bronze drive cable so that, when the large frequency knob is turned, both oscillator and amplifier tuning shafts operate in unison.

The main frequency dial, which is over 6 inches in diameter, is directly attached to the oscillator capacitor drive shaft. Gear teeth on the perimeter of the oscillator dial are driven by a vernier dial which is divided into 100 parts. The vernier dial turns 10 times throughout the full oscillator dial range. This provides a logging scale which divides any range into 1000 parts. Backlash is negligible because the main dial is a spring-loaded, split-gear assembly.

Rack Mountable

As a package the instrument is on a standard 19-inch rack mountable panel. End bells provided with the instrument cover the protruding panel ends in case rack mounting is not desired.

Conclusion

The BRC Type 225-A Signal Generator is a compact, highly stable, useful signal generator free from the spurious effects which can make life unpleasant for the engineer or technician. The following published specifications speak for themselves in demonstrating the degree of success achieved.

Specifications

RADIO FREQUENCY CHARACTERISTICS

RF Range
Total Range: 10 to 500 mc.
No. Bands: 6
RF Accuracy: ±0.5% (after two hour warmup)
RF Settability: ±0.05%
RF Calibration
Main Dial: Increments of approximately 1%
Vernier: 1000 divisions through each range.
RF Stability (after 2 hour warmup)
Short Term: ±0.001% (5 minutes)
Long Term: ±0.01% (1 hour)
Line Voltage: ±0.001% (5 volts)
RF Output
Range: 0.1 μv to 0.1 volts (across external 50 ohm load.)
Accuracy: ±10% 0.1 to 50 k μv, 10 to 250 mc.
±15% 0.1 to 50 k μv, 250 to 500 mc.
±20% 0.05 to 0.1 v, 10 to 500 mc.
Impedance: 50 ohms (25 ohms at terminals of Type 501-B Output Cable)
VSWR: 1.2
RF Leakage: Sufficiently low to permit measurements at 0.1 μv.

AMPLITUDE MODULATION CHARACTERISTICS

AM Range
Internal: 0 to 30%
External: 0 to 30%
AM Accuracy: ±10% at 30% AM, 10 to 250 mc.
±15% at 30% AM, 250 to 500 mc.
AM Calibration: 10, 20, 30%
AM Distortion: 5% 10 to 250 mc.
7% 250 to 500 mc.
AM Fidelity: ±1 db 40 cps to 20 kc.
Incidental FM: 0.001% or 1000 cps, whichever is greater, at 30% AM
External AM Requirements: 10 volts RMS into 4000 ohms for 30% AM

FREQUENCY MODULATION CHARACTERISTICS

FM Range: (External)
0 to between 5 kc and 60 kc, depending upon frequency in the range 130 to 500 mc.
FM Calibration: Deviation sensitivity vs. frequency nomograph
Incidental AM. 10%
External FM Requirements: 10 volts RMS into 1000 ohms

PULSE MODULATION CHARACTERISTICS

PM Source: External
PM Rise Time: 5 μsec 10 to 40 mc.
3 μsec 40 to 80 mc.
2 μsec 80 to 500 mc.
PM Overshoot: 10% 10 to 100 mc.
25% 100 to 500 mc.
External PM Requirements: 10 volts peak negative pulse, 20 ma. peak short-circuit capability.

MODULATING OSCILLATOR CHARACTERISTICS

MO Frequency: 400 and 1000 cps.
MO Accuracy: ±10%

POWER REQUIREMENTS

225-A: 105-125 volts, 60 cps, 80 watts.
225-AP. 105-125 volts, 50 cps, 80 watts.

A Signal Generator Calibrator for RF Level and Per Cent AM

ROBERT POIRIER, *Development Engineer*

In response to consumer requests for slightly higher output voltages the type 245-B RF Voltage Standard which initially was designed to provide accurately calibrated RF output voltages (supplied by an external source) of 0.5, 1.0, and 2.0 microvolts for the purpose of checking receiver sensitivity and low-level calibration of signal generators has been modified to greatly enhance its usefulness. The modified instrument, which supersedes the 245-B RF Voltage Standard, is known as the Signal Generator Calibrator. It is available in two models, types 245-C and 245-D. Added features include: 1) a choice of calibrated RF output voltages; vis., 5, 10, and 20 microvolts or 0.5, 1.0 and 2.0 microvolts; 2) direct reading of three calibrated, unmodulated RF input voltages; viz., 0.025, 0.05, and 0.1 volt, and 3) direct reading of the per cent amplitude modulation of the RF input voltage to 100%. With the exception of attenuation, the types 245-C and 245-D are identical. The Type 245-C is the high output instrument providing 5, 10, and 20 microvolts calibrated output voltage and the Type 245-D is the low-output instrument providing 0.5, 1.0, and 2.0 microvolts calibrated output voltage.

Figure 1. Type 245-C/D Signal Generator Calibrator

Principles of Operation

In order to provide for slightly higher output voltages of the same accuracy as obtained from the Type 245-B RF voltage standard, it was considered expedient to change the shunt resistance element of the micropotentiometer[1] from 0.0024 ohms, as used in the 245-B, to 0.024 ohms for the 245-C. Output voltages of 5, 10, and 20 microvolts are thereby obtained for the same nominal input voltages of 0.025, 0.05, and 0.1 volt respectively, without changing the physical structure of the attenuator. All other methods of increasing the output of the Type 245-B micropotentiometer, such as decreasing the input impedance to 6 ohms (in lieu of changing the shunt resistance element) or increasing the input voltage requirement are inappropriate to the desired result. The problem of changing the resistance of the shunt element without appreciably affecting its physical dimensions is largely a matter of finding a suitable material from which to make the element; noting that the thickness of the resistance element is closely restricted at one end by the size of the individual molecules which influences the mechanical stability of the resistance, and at the other end by the depth of penetration of the RF current which influences the electrical fidelity of the attenuator. The effort which has recently been given to the search for new materials and new methods of fabricating the shunt resistance element has yielded an improved element for both the 245-C and 245-D Signal Generator Calibrator. The RF Voltage monitor and attenuator system, described in previous Notebook articles [2,3,4] is shown in Figure 2. The value of the disc resistor used is different for the Type 245-C and 245-D as indicated.

The measurement of unmodulated RF input voltages of 0.025, 0.05, and 0.1 volt will be more fully exploited in the Types 245-C and 245-D than it was in the 245-B. Two innovations bearing directly on this function are: 1) a meter function switch to provide correct calibration of either the RF input voltage or the RF output voltage, and 2) frequency compensation of the RF monitor to eliminate the need of a frequency correction curve. Both the meter function switch and the frequency compensation permit greater accuracy of input voltage measurement than is attainable with the Type 245-B. The meter function switch precludes the ±6% maximum error of input voltage measurement which in the Type 245-B results from the tolerances of the nominal low frequency values of the resistors in the micropotentiometer. Frequency compensation is accomplished by means of a low-Q inductive network in series with the input to the RF voltmeter which reduces the input VSWR due to capacitance of the RF monitor circuit.

A completely new function of the Signal Generator Calibrator is the direct reading of per cent amplitude modulation of a modulated RF input signal around the 0.1-volt input level. Operation of the % AM feature requires an initial calibration of the unmodulated carrier at the 0.1-volt input level using the input voltage measuring function of the instrument. The meter function switch is then set to % AM and reads the full wave average of the detected modulation envelope. The additional gain required is developed in a two-stage transistor amplifier with inverse feedback to stabilize the gain and improve the linearity of the %AM scale. Although the RF detector operates

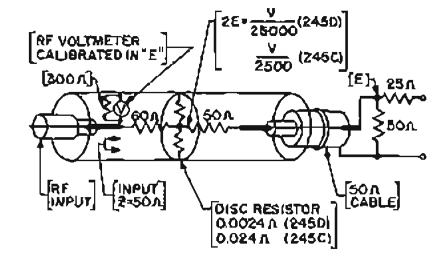

Figure 2. Attenuator and Voltmeter

nearly square law, the % AM indication is nearly linear because the AM detector is biased to be always forward conducting. [2,4] The waveform of the detected modulation versus the modulation envelope is represented in Figure 3. The output polarity of the diode is negative. Referring to Figure 3, any peak-to-peak limits of the detected modulation are represented as $(y_1 - y_2)$ and the corresponding peaks of the modulation envelope are represented as $(x_2 - x_1)$.

From

$y = -x^2$ (for the square law diode)

$$y_1 - y_2 = x_2{}^2 - x_1{}^2 =$$

$$(x_2 - x_1)(x_2 + x_1)$$

For any given operating point x, y, $(x_2 + x_1)$ is a constant, 2x since $x + \Delta x + x - \Delta x = 2x$

$$\therefore (y_1 - y_2) \propto (x_2 - x_1)$$

for either distortionless or odd harmonic modulation distortion. Some even harmonic distortion of the modulation envelope is produced in the instrument as the AM detector law changes near the crests of 100% modulation. Experimentally, this has been found negligible and is anticipated in the calibration of the %AM meter scale.

Since the % AM readout is in terms

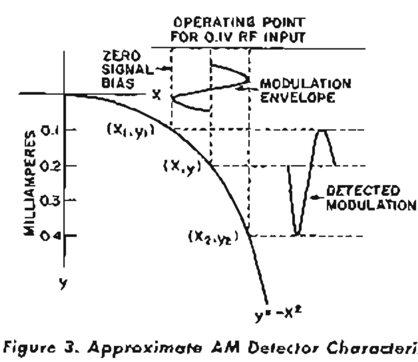

Figure 3. Approximate AM Detector Characteristic

of the full-wave average of the detected modulation, the % AM indication is subject to the usual errors resulting from interpreting peak-to-peak information from an average readout. The instrument is provided accurately calibrated for undistorted sinusoidal modulation. Users of this instrument should be cognizant of a source of error which is not necessarily related to distortion of the modulation envelope; viz., RF carrier shift of the initial unmodulated calibrated level which accompanies the modulation. Nonlinear modulation is, of course, a common cause of this carrier shift; an uncommon cause (which may become increasingly common) is inverse feedback regulation of the output power level of a signal generator. The Signal Generator Calibrator will be offered with the % AM indication calibrated for no shift of the carrier power level with modulation.

Uses of the 245-C and 245-D

The Signal Generator Calibrator retains the originally conceived function of providing accurate, low-level RF output voltages. Uses for this and the new features are enumerated as follows:

1. Accurate spot checks can be made of receiver sensitivity over the range of 500kc to 1000 mc and 0.5 microvolt to 20 microvolts.
2. By associating the precision fixed attenuator (74 db in the 245-C or 94 db in the 245-D) with a precision piston attenuator the range of calibrated low-level output voltages can be extended well below 0.5 microvolt, for receiver sensitivity and noise figure measurements, subject only to limitations in shielding the receiver from the signal source.[5]
3. Accurate spot checks can be made of unmodulated signal generator output voltages from 500kc to 1000mc at high-output levels between 0.025 and 0.1 volt inclusive, at low levels (with a suitable receiver) between 0.5 microvolt and 20 microvolts inclusive, or at a level less than 0.5 microvolt with precision fixed attenuators.
4. Fast, accurate measurement can be made of % amplitude modulation to 100% for modulating frequences from 20 cps to 20 kc.

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2. Gorss, C. G., "An RF Voltage Standard Supplies A Standard Signal At A Level of One Microvolt," BRC Notebook No. 5, Spring, 1955.
3. Moore, W. C., "Use of the RF Voltage Standard Type 245-A," BRC Notebook No. 7, Fall, 1955.
4. Gorss, C. G., "Calibration of An Instrument for Measuring Low-Level RF Voltages," BRC Notebook No. 14, Summer, 1957.
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NEW BRC INSTRUMENTS TO BE SHOWN AT IRE SHOW

Booths 3101 — 3102

Boonton Radio orporation will offer three new instruments to the electronic industry in 1959. The instruments, which will be on display in the BRC exhibit at this year's IRE Show in New York, include: the new Type 225-A Signal Generator, the new Types 245-C and D Signal Generator calibrators, and the new Type 202-G Telemetering Signal Generator.

Designed for operation in the 10 to 100-megacycle range, the Type 225-A Signal Generator embodies circuit and structural design innovations which provide a new standard of precision and stability.

The Types 245-C and D Signal Generator Calibrators, for the first time, provide a convenient portable instrument for measuring and calibrating the RF level and percentage AM on Signal Generators in the range from 500KC to 1000MC. They may also be used to provide a calibrated source of low-level RF voltage for the precision testing of receiver sensitivity.

The Type 202-G FM-AM Signal Generator is an improved generator for the testing and calibration of FM telemetering systems offering RDB modulating frequencies and complete coverage of the recently extended 215 to 260 MC telemetering band.

Visit booths 3101 and 3102 at the IRE Show during March 23 to 26 where BRC personnel will be on hand to give you more facts about these and other BRC instruments.

A Telemetering FM-AM Signal Generator

FOR COVERAGE OF THE RECENTLY EXTENDED TELEMETERING BAND

HARRY J. LANG, *Sales Manager*

Type 202-G Telemetering FM-AM Signal Generator

With the rapid development of FM telemetering systems in the post-war period, BRC, as a pioneer manufacturer of FM Signal Generators, developed the Type 202-D Signal Generator in 1949 to provide FM-AM coverage of the then assigned 215-235 mc telemetering band. The 202-D, which was subsequently assigned military nomenclature SG-59/U, offered continuous RF coverage from 175 to 250 mc at output levels from 0.1 μv to 0.2 volt and was designed for both internal and external FM and AM. Frequency modulation was provided by a reactance-tube circuit, specially designed to maintain constant deviation, with carrier frequency. Controls provided continuously variable deviation from 0 to 240 kc Amplitude modulation, from the internal audio oscillator, was continuously variable from 0 to 50% and this range could be entended to 100%, employing an external modulating oscillator. Pulse modulation, from an external source, was also provided. The internal audio modulating oscillator provided a choice of eight fixed frequencies with nominal values between 50 cps and 15 kc.

Further development and expansion of FM telemetering, increased the applications of this type precision signal generator. Both military and commercial system requirements called for the incorporation of such a unit into the complete telemetering system. Recognizing this trend, BRC redesigned the 202-D and made available the Type 202-F Signal Generator in 1957. The 202-F was mounted in a new type of cabinet (which has now become standard on many of our instruments) that would provide for both bench and rack mounting. The instrument, as furnished, includes a complete cabinet and dust cover. By simple removal of the cabinet end bells, the instrument will mount in a standard 19" relay rack, making it ideal for system applications. Several other circuit innovations were also incorporated which improved stability and modulation fidelity.

When the telemetering band was recently extended up to 260 mc, it became immediately apparent that a further redesign of the 202-F was necessary in order to provide complete RF coverage of the new band. The new Type 202-G Signal Generator offers continuous RF coverage from 195 to 270 mc completely blanketing the new 215 to 260 mc band. As a further aid to the convenient checkout of telemetering systems, the nominal audio modulating frequencies, provided in the earlier 202-D and F, were replaced with the following standard RDB values:

50, 400, 730 cps; 1.7, 3.9, 10.5, 30.0, and 70.0 kc.

All other mechanical and electrical characteristics are identical to the obsolete 202-F.

EDITOR'S NOTE

25th Anniversary for BRC

1959 represents a major milestone for BRC marking the completion of 25 years as a designer and manufacturer of precision electronic laboratory instruments. Befittingly, this anniversary closely follows the purchase of a new 70-acre plant site in the picturesque Rockaway valley and the announcement that plans have been formulated to erect a new, enlarged factory in the near future. Much progress and expanison has taken place in the electronics industry since the inception of BRC back in 1934 and as we reflect back over the years, we feel a certain amount of pride in the knowledge that we have made a direct contribution to this growth.

Founded in 1934 by William D. Loughlin, a pioneer in the industry, BRC, from its earliest years, concentrated its engineering skills toward the development of general-purpose precision laboratory tools for the electronic designer. The first product was the now famous Q Meter which still represents a basic tool and has been accepted as a standard throughout the industry. Improved versions of the Q Meter have since been designed and continue to form a major portion of BRC's product line.

With the development of frequency modulation techniques in the late thirties, the company's interests were focused in this area and resulted in the first commercial FM Signal Generator which made its debut at the Boston IRE meeting in 1940. BRC has since expanded its line of FM Signal Generators to cover all commercial and military applications in this field.

During World War II, BRC, like all U. S. industry, was directly engaged in the manufacture of critically needed products for the Armed Services. In addition to continuing the production of several versions of the Q Meter and Signal Generators, BRC, in cooperation with the M.I.T. Radiation Laboratory, developed microwave Signal Generators for the calibration of radar systems. These efforts won the company numerous awards and commendations and served to encourage a broadening of activities in the post-war period.

Boonton Radio Corporation in 1939

Beginning in 1946, an entirely new line of FM Signal Generators was introduced to serve the then embryonic television industry and several new instruments for general purpose RF impedance measurements followed shortly thereafter. With the introduction of the VOR aircraft navigation system and the ILS aircraft landing system, specialized Signal Generators for these systems were designed and introduced in the period from 1948 to 1953.

Boonton Radio Corporation as it appears today

Beginning in 1953, the basic Q Meters were redesigned and a Sweep Signal Generator, a self-contained VHF impedance bridge, Q Standards, and an RF Voltage Standard, as well as a Film Gauge for industrial applications, were added to the line. Last year marked the development of a unique Q Comparator which is now in production.

In partial celebration of its 25 years in the electronic instrument industry, BRC is introducing four new instruments at the New York IRE Show this month: the Type 225-A Signal Generator, the Types 245-C and D Signal Generator Calibrators, and the Type 202-G FM-AM Signal Generator. These new instruments are described in this issue.

WIN A Q METER

Visit the BRC exhibit (Booths 3101-3102) at the IRE show and enter the Q Meter Contest. A factory reconditioned Type 160-A Q Meter will be awarded again this year to the person whose Q estimate is closest to the actual measured Q of a special coil which will be on display in the BRC exhibit. Complete information will be furnished by BRC representatives in attendance at the exhibit.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Applications of the Signal Generator Calibrator

JAMES E. WACHTER, *Project Engineer*

Figure 1. Type 245-C Signal Generator Calibrator.

The Type 245 Signal Generator Calibrator, available in two models, is a portable, self-powered, precision instrument which may be used as a calibrated high-level rf voltmeter; a source of calibrated microvolt-level voltage of known impedance; and a calibrated percent AM meter. Being portable and furnishing its own power, the instrument is especially adaptable to correlating signal generators and receivers located considerable distances from one another, such as might be found in airline communications installations and general service depots. With its calibrated output voltages, the instrument is ideal for calibrating receivers, since it permits the use of almost any shielded signal source, regardless of output accuracy, obviating the need for elaborate test equipment. Used for calibrating signal generator output systems and for calibrating percent amplitude modulation, the Calibrator not only saves inspection time, but frees other test equipment for more profitable use elsewhere.

Description

A simplified block diagram of the Signal Generator Calibrator is shown in Figure 2. The two basic parts of the instrument, a non-frequency-sensitive rf voltmeter and a precision fixed attenuator, are combined in a rigid mechanical unit of coaxial construction, a cross section of which is shown in Figure 3. The coaxial unit, together with suitable switching, amplifying, and metering circuits are housed in a small, sloping-panel cabinet measuring 9 inches wide, 5 inches deep, and 5 inches high, and weighing only 5 pounds. All of the operating controls and a very sensitive meter are located symmetrically on the front panel. Power is supplied from internal mercury batteries.

Used as a high-level input voltmeter, the Calibrator operates as a 50-ohm monitor of the input voltage at the voltmeter diode, which, within the accuracy specifications of the instrument, is essentially the same as the voltage applied to the input cable. The instrument will read, directly, input rf voltages of 0.1, 0.05, and 0.025 volt.

When operated as a source of low-level voltage, the Signal Generator Calibrator must be supplied from an external source. The voltage applied to the instrument is monitored at the input to the coaxial attenuator and the low-level output from the attenuator appears in series with a 50-ohm impedance-matching resistor. The rf voltmeter is calibrated to indicate the output voltages of 20, 10, and 5 μv (245-C) or 2, 1, 0.5 μv (245-D) across a 50-ohm termination connected directly to the output jack of the Calibrator. The accessory Type 517-B Output Cable supplied with the Calibrator provides a 50-ohm terminating resistor followed by a 25-ohm impedance-matching resistor which raises the equivalent source impedance at the end of the cable to 50 ohms.

When using the Calibrator as a percent AM meter, it is necessary to perform an initial setup, using an unmodulated rf input signal to establish the % AM meter reference. This setup is performed with the Meter Function switch in the RF IN position. Once the reference is established, the instrument is switched to % AM and the voltmeter detects the same signal with modulation applied. The ac component of the voltmeter is amplified, detected, and indicated on the calibrated % AM meter scale.

A more complete description of the theory and design of the Signal Generator Calibrator is given in references 1, 2, and 3. Complete performance specifications are given in this article under "Specifications".

Calibrating Signal Generator Output

Calibrating the output of a signal generator is generally performed using

YOU WILL FIND . . .

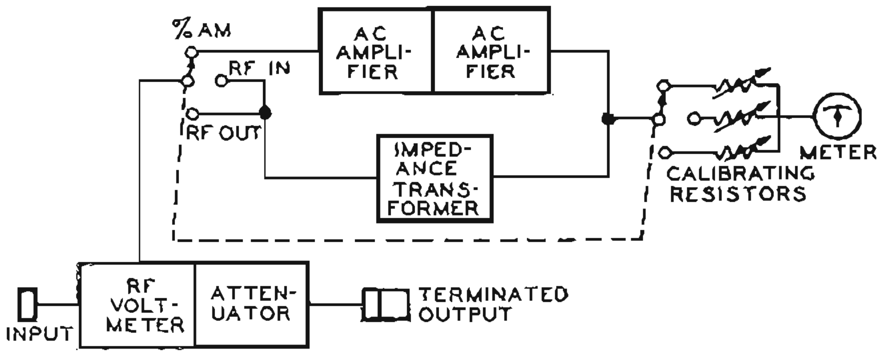

Figure 2. Signal Generator Calibrator System Block Diagram.

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a bolometer bridge. Most bolometer bridges are essentially high-level devices (of the order of 0.1 volt) and therefore can only be used to check the upper end of a piston type attenuator, meaning that the mechanical law of the attenuator must be relied upon for lower voltage levels.

Using the Calibrator, high-level voltages of 0.1, 0.05, and 0.05 volt can be read, directly, merely by changing ranges and adjusting the generator output to the appropriate calibration mark on the meter. If it is desired to calibrate voltages higher than 0.1 volt, it is only necessary to insert a precision pad of the appropriate attenuation between the generator output and the Calibrator input. For example, to calibrate a generator output of 0.2 volt, a pad of 6 db attenuation would be used, and the generator output level would be adjusted until the Calibrator indicated 0.1 volt.

Signal generator low-level output voltages can be calibrated using the Calibrator as a transfer device. The 245-C will calibrate, directly, the levels of 20, 10, and 5 microvolts and the 245-D will calibrate the levels of 2, 1, and 0.5 microvolts. Used in this manner, the Calibrator acts as a precision fixed attenuator of 80 db attenuation in the case of the 245-C and 100 db in the case of the 245-D.

The signal generator is connected to the Calibrator, with the Calibrator set to indicate output voltage, and the Calibrator output is connected to a receiver having a suitable signal-to-noise ratio and a means of indicating relative signal level. The generator output is then adjusted until the Calibrator indicates the desired output; for example, 2 μv. The receiver is tuned and peaked to this signal and the receiver output noted. The Calibrator is removed from the setup, and the signal generator, with its output set at minimum, is connected to the receiver. The generator output is increased until the receiver indicates the same signal level as previously noted. The generator is now delivering 2 μv (within specification tolerance) and its attenuator calibration can be checked accordingly.

Should it be desired to check the attenuator of a generator at lower levels than the Calibrator will provide (either 5 or 0.5 μv); this may be done using precision fixed attenuators between the Calibrator output and the receiver.

Exploring Law of Attenuation

A point of interest is the fact that the law of attenuation of a piston attenuator can be reasonably well explored using the Signal Generator Calibrator. The generator output is first checked at the maximum level and at levels of 6 db and 12 db below maximum using the Calibrator input voltmeter calibrations of 0.1, 0.05, and 0.025 volt. Then, the minimum calibrated output from the generator is checked using the Calibrator and a receiver. The checks at maximum and minimum output fix the overall calibration of the attenuator. If the measured attenuation at very low levels is less than is indicated by the attenuator dial calibration, it is an indication of rf leakage internal to the attenuator system; i.e., the attenuator is receiving power from other than the desired source, a variation in the attenuator bore diameter, or errors in the dial drive. The three high-level checks determine the linearity of the attenuator calibration, since it is in this high-level region, where the attenuator loop is near the mouth of the attenuator tube, that the law of the attenuator is generally violated by spurious modes of propagation.

Measuring Receiver Sensitivity

The Signal Generator Calibrator provides signal levels which are most often required for sensitivity measurements of both narrow and broad-band receivers: 2, 1, and 0.5 μv from the 245-D and 20, 10, and 5 μv from the 245-C. However, if levels lower than 0.5 μv are required for more sensitive receivers or for noise figure measurements, these can be obtained by associating the Calibrator with either a precision piston attenuator (for complete versatility) or with precision fixed attenuators. The lowest level of measurement attainable then is subject only to the limitation in shielding the receiver from the signal source.

Used in conjunction with a swept signal source, the Calibrator permits the dynamic checking of receiver sensitivity. The receiver, previously aligned with a high-level signal, is supplied with a calibrated low-level, swept signal from the Calibrator and the receiver sensitivity curve is observed as a visual display on an oscilloscope. This test requires the use of a well shielded sweep signal generator such as the BRC Type 240-A.

It should be noted here, that in receiver sensitivity measurements where the highest degree of accuracy is required it may be necessary to correct the Calibrator voltmeter indications for the output cable attenuation. If the output cable supplied with the Calibrator is used, it is advisable to apply this correction at frequencies above 500 mc. For any other cable used to connect the Calibrator to the receiver, the correction will depend upon the length and type of cable employed.

Calibrating Signal Generator Percent AM

Calibration of signal generator percent AM is generally performed using an equipment to heterodyne the signal generator carrier frequency down to some frequency acceptable to an oscilloscope and then sweeping the oscilloscope with the signal generator modulating frequency. The trapaziodal pattern thus obtained on the oscilloscope is a measure of the percent AM, subject, of course, to the interpretation of the observer.

With the Calibrator only two steps are required for percent AM calibration. First, a reference level is established on the Calibrator input range, using the unmodulated output of the signal. Then, with the Calibrator set on the % AM range, modulation is applied to the carrier and percent AM is read, directly, on the Calibrator % AM scale.

Distortion

It should be noted that the instrument is accurately calibrated for undistorted, sinusoidal modulation. The presence of distortion, therefore, introduces a corresponding error in the meter reading as indicated in Figure 4. It follows that with distortion present and the most accurate results required, it is necessary to know the kind and amount of distortion so that the meter reading may be corrected accordingly.

Carrier Shift

Another possible cause for error in the % AM indication would be a shift in carrier level due to the application of modulation. If percent AM calibrations are to be made under conditions of carrier shift, it may be desirable to recalibrate the % AM meter. For any reasonable linear carrier shift with modulation, this recalibration is easily accomplished using the internal % AM meter sensitivity control to adjust the meter to indicate correctly a known percent AM.

In this connection, it is interesting to note that the Calibrator may be used to detect gross shifts in carrier level caused by the application of modulation. When used in the RF IN position as an rf voltmeter, the meter indication for a modulated carrier will be somewhat greater than for the same level carrier without modulation. For a carrier level of 0.1 volt, for example, the meter indication is increased approximately $\frac{1}{16}$ inch for 50% modulation. If, when modulation is applied, the meter reading does not increase about the expected amount, carrier shift is indicated. Should the meter read less with modulation applied, a considerable shift in carrier level would be indicated, indicating the presence of so-called "downward modulation" (positive peak clipping of the envelope).

Monitoring Transmitter Output

An antenna connected to the Calibrator input and placed in proximity to a transmitter antenna, or at some other point radiating power, would enable the Calibrator to indicate relative field strength when used in the RF IN position. If the input to the Calibrator were adjusted to the 0.1-volt reference level, the instrument could be used in the % AM position to indicate, directly, the percent AM present.

An improvement of this application would be the connection of a well shielded, sensitive receiver to the low-level output of the Calibrator to directly monitor the transmitter while in reasonable proximity to it.

Impedance Matching

It should be noted that for the best accuracy in the measurements described herein, as with most measurements involving the interconnection of instruments, the problem of matching impedances must be taken into account. Reference 4, covers this subject in some detail.

have an output impedance of 50 ohms. If the output impedance is not 50 ohms, an impedance matching pad should be used. The signal level should be monitored at the input to the Calibrator with an accurate ac voltmeter of high input impedance. Internal meter sensitivity controls are provided in the Calibrator to adjust the meter for correct indication of the three input voltages of 0.1, 0.05, and 0.025 volt

A second method for calibrating the input voltmeter is to compare the Calibrator voltmeter reading directly to the calibrated output of a known accurate signal generator having a piston type attenuator, such as the BRC Type 202-E or Type 225-A Signal Generators. These generators have a 50-ohm output impedance. If a generator with other than 50 ohms output impedance is used, a matching pad must be employed, and

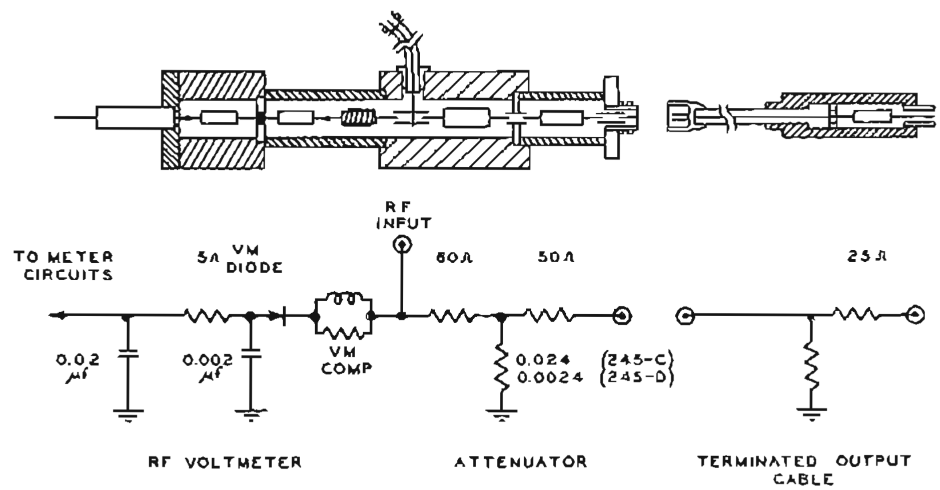


Figure 3. Coaxial Voltmeter-Attenuator Assembly and Output Cable.

Recalibration

The general design of the Signal Generator Calibrator is such that recalibration is seldom necessary. However, if recalibration becomes necessary, it is recommended that the instrument be returned to the factory for the most accurate calibration. If this is not convenient, field calibration of a lesser degree of accuracy may be performed.

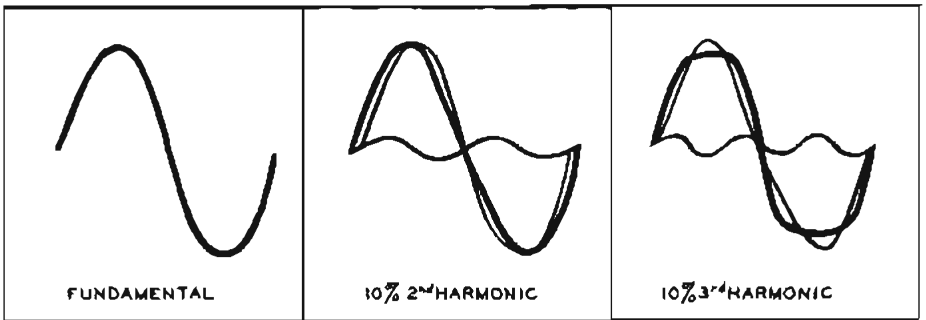
Input Voltmeter

Field recalibration of the input voltmeter may be accomplished using either of two methods. The first method involves the use of a 1000 cps source and a 60 μf additional bypassing capacitor. The bypass capacitor is connected at the voltmeter output at the end of the coaxial block. The 1000 cps should be of low distortion and the source should

the attenuation of the pad must be taken into account when determining the input to the Calibrator as indicated by the signal generator attenuator calibration. The input signal is adjusted for 0.1, 0.05, and 0.025 volt and the internal sensitivity controls in the Calibrator are adjusted so that the Calibrator indicates these levels correctly. If a generator of dubious accuracy is used, the three levels can be established by calibrating the generator at the three levels, using a bolometer bridge.

% AM Meter

The % AM Meter can be recalibrated in much the same way as the rf voltmeter; i.e., by comparison with a known percent AM. As mentioned previously, it is desireable that the modulation be of low distortion and that there be no



	10% 2nd HARMONIC	10% 3rd HARMONIC
P.P.V. ERROR	+1.5%	-10
AVG. V. ERROR	0	+3.3 %
METER READING ERROR	-1.5%	+13.3%

Figure 4. Example of Error Caused by 10% Second and Third In-phase Harmonic Modulation Distortion.

carrier shift with modulation. A suitable generator for this purpose is one such as the BRC Type 232-A. If only an uncalibrated AM source is available, it may be calibrated using the trapazoidal method. A convenient modulating frequency is 1000 cps and a good calibration point is 50%. Several points may be checked, if desired, and a compromise setting may be made for any slight error noted. With the modulated signal applied to the Calibrator, the internal % AM meter sensitivity control is adjusted so that the meter indicates the correct % AM.

Low-Level Output

When the Signal Generator is received from the factory, it is accurately calibrated. It would be advisable at this time to calibrate the high and low-level outputs of several signal generators and to record the information. This data would then serve as a convenient means for checking the calibration of the signal Generator Calibrator at some later date when there is reason to believe that its accuracy has deteriorated.

Field recalibration of the Calibrator low-level output (attenuation) can be performed easily by direct comparison to a signal generator having an accurately calibrated precision piston attenuator which has been standardized with a bolometer bridge. A sensitive receiver of good signal-to-noise ratio, with a means for indicating relative signal level, is required to monitor the low-level signals. The generator is connected to the receiver and a low-level signal of the order of 2 μv is applied. The receiver is tuned and peaked to this signal and the relative signal level indication noted. The generator is then disconnected from the receiver and the Calibrator (set for a 2 μv output) is inserted; the input cable connected to the generator and the terminated output cable connected to the receiver. The generator output is increased until the receiver indicates the same signal level as was previously indicated (2 μv), and the internal sensitivity control in the Calibrator is adjusted so that the Calibrator meter indicates the correct calibration. The same procedure would be used for the two lower levels of 1 μv and 0.5 μv utilizing the sensitivity controls provided for each level. This method is of greatest value when the generator has been previously checked by the Signal Generator Calibrator.

As mentioned previously, it is necessary to properly match impedances if the best accuracy is to be obtained. Also, it is advisable to use identical type and length cables when making comparison measurements.

Specifications

RADIO FREQUENCY MEASUREMENT CHARACTERISTICS

RF Range: 500 Kc. to 1000 Mc.
RF Voltage Measurement Levels:
Input: 0.025, 0.05, 0.1 Volts.
Output: 5, 10, 20 μv. (245-C).
0.5, 1, 2 μv. (245-D).
RF Voltage Accuracy:
Input: ±10% 500 Kc. to 500 Mc.*
±15% 500 Mc. to 1000 Mc.*
*When supplied from a 50 ohm nominal source, with a VSWR <2.
Output: ±10% 500 Kc to 500 Mc.
±20% 500 Mc. to 1000 Mc.
RF Impedance:
Input: 50 ohms.
Output: 50 ohms.*
*At output jack on Instrument and at output connector of Type 517-B Output Cable.
RF VSWR:
Input: <1.3 500 Kc. to 500 Mc.
<1.6 500 Mc. to 1000 Mc.
Output: <1.05 500 Kc. to 100 Mc.*
<1.07 100 Mc. to 500 Mc.*
<1.1 500 Mc. to 1000 Mc.*
*At output connector of Type 517-B Output Cable.

AMPLITUDE MODULATION MEASUREMENT CHARACTERISTICS

AM Range: 10 to 100%.
AM Accuracy: ±10% 30 cps. to 15 Kc.*
±15% 20 cps. to 20 Kc.*
*Modulating frequency.
AM Frequency Range: 20 cps. to 20 Kc.
RF Input Requirements: 0.05 volts.

References

1. Gorss, C. G., "An RF Voltage Standard Supplies a Standard Signal at a Level of One Microvolt", BRC Notebook No. 5.
2. Gorss, C. G., "Calibration of an Instrument for Measuring Low-Level RF Voltages", BRC Notebook No. 14.
3. Poirier, R., "A Signal Generator Calibrator for RF Level and Percent AM" BRC Notebook No. 21.
4. Moore, W. C., "Use of the RF Voltage Standard Type 245-A", BRC Notebook No. 7.

Measurement of Voltage Sensitive Capacitors

As evidenced by an ever-increasing rate of inquiries, there is a growing interest today in measuring the dynamic parameters of voltage sensitive diode capacitors at radio frequencies. Much information has appeared in the literature on these devices, so that it will suffice to say that these diodes, under certain biasing conditions, exhibit the characteristics of variable capacitors. Our particular interest in this matter is to indicate a means of measuring the equivalent series resistance, equivalent series capacitance, and of course, Q of these capacitors. The Type 250-A RX Meter, a completely self-contained RF bridge operating over a range from 500 kc to 250 mc, and completely described in issue number 2 of the Notebook, has been found ideally suited for this application and the purpose of this article is to outline a method for carrying out these measurements. Since the Q can be very easily found from the resistance and capacitance, and inasmuch as the equivalent series capacitance and equivalent parallel capacitance are approximately the same where the Q involved is 10 or greater, measurements can be simplified to the determination of the equivalent series resistance and equivalent capacitance.

Basic Technique

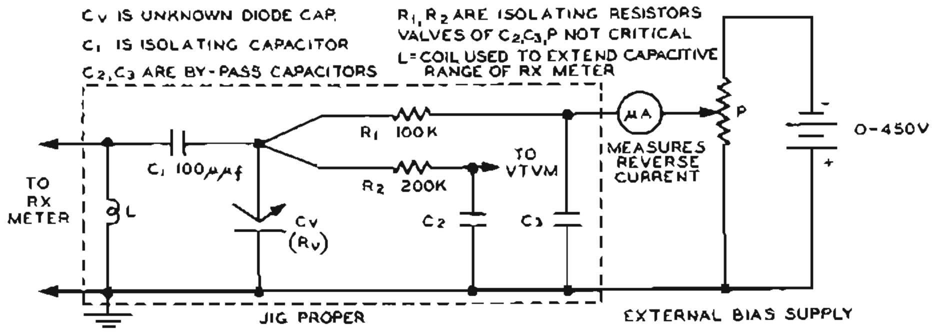
Basically the technique used in measuring these voltage sensitive diode capacitors is very similar to that employed with other circuits requiring a known and controlled amount of bias applied to the component under test. Methods of introducing biasing potentials and typical measuring circuits for the RX Meter have been discussed in issue number 6 of the Notebook. This article also illustrated a method for extending the capacitance range of the RX Meter beyond the direct-reading range of 20 $\mu\mu f$ available on the instrument. Actually, then, the matter of measuring the voltage sensitive capacitive diode merely involves the introduction of the necessary bias to establish the proper operating point and a knowledge of the method used for extending the capacitive range of the RX Meter. This information, when properly applied will yield for frequencies below 20 mc, the equivalent parallel resistance and capacitance, which can then be converted to the Q of the circuit.

High-Frequency Technique

For measurements of these voltage sensitive variable capacitors at frequencies above approximately 20 mc, however, it became apparent that some modification of the low frequency techniques would have to be employed since the coils used at 50 mc or above would have inductances below 0.1 microhenries, and the 0.003 microhenry residual inductance of the bridge would now start to introduce errors and have an increasingly greater influence on the accuracy of the measurements as smaller coils were used for range extension.

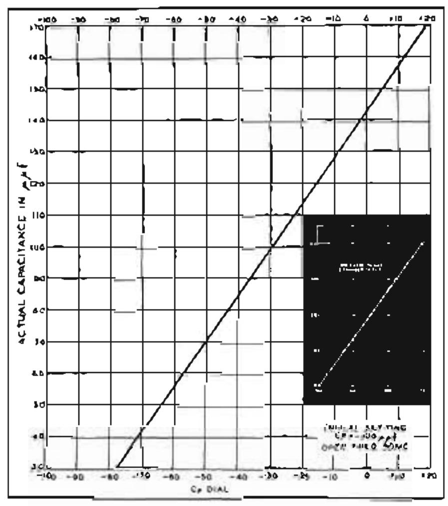
As a matter of background information, it might be interesting to note that a solution to this situation was approached in various ways. For example, an initial attempt was made to circumvent the use of a coil entirely, by first using a series capacitor in the order of 20 $\mu\mu f$, measuring it both for its capacitance and losses (including the jig), and converting these values to their series equivalents. By then adding the actual diode to be tested, similarly reading the entire circuit parameters and converting them to their series equivalent, an evaluation could be made of the diode by subtracting out the losses external to it. However, after some extensive study and evaluation of this method, it was decided that while it would yield a realistic value of capacity in most cases, the conversion and determination of the equivalent series resistance of the diode was not satisfactory.

Some thought was also given to the use of a quarter-wave-length line as a means of converting the capacitance of the diode to an inductive reactance, thereby extending the range automatically to an equivalent 100 $\mu\mu f$. However, this advantage was gained at the expense of various disadvantages; namely, each quarter-wave-length line would be suitable at only a given frequency, the values of resistance that could be meas-



Courtesy Pacific Semiconductors, Inc.

Figure 1. Typical circuit for measuring voltage sensitive capacitors at frequencies above approximately 20 mc.



Courtesy Pacific Semiconductors, Inc.

Figure 2. Calibration curve for C_P dial when measuring capacitances.

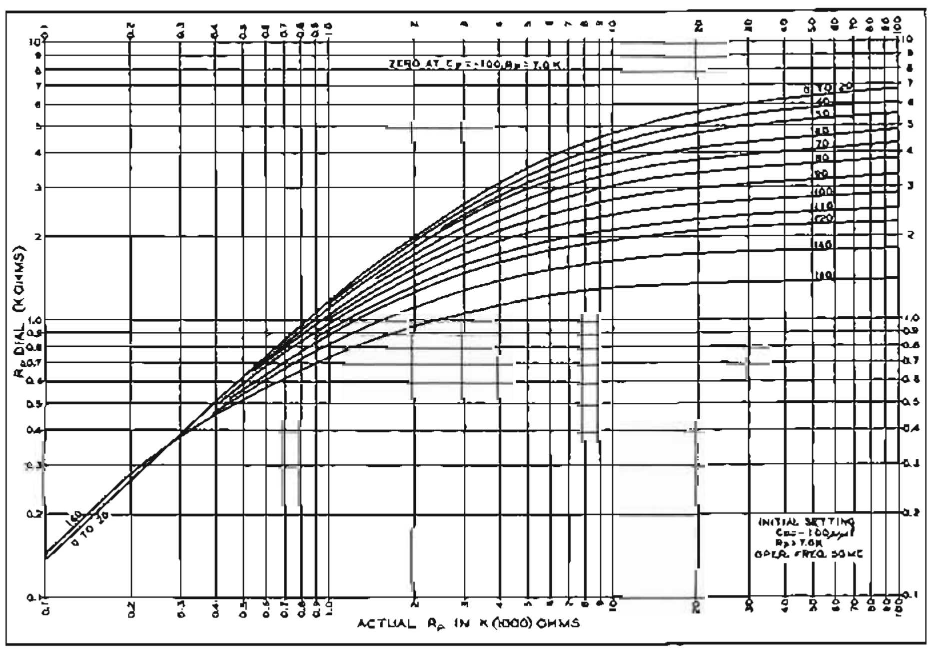
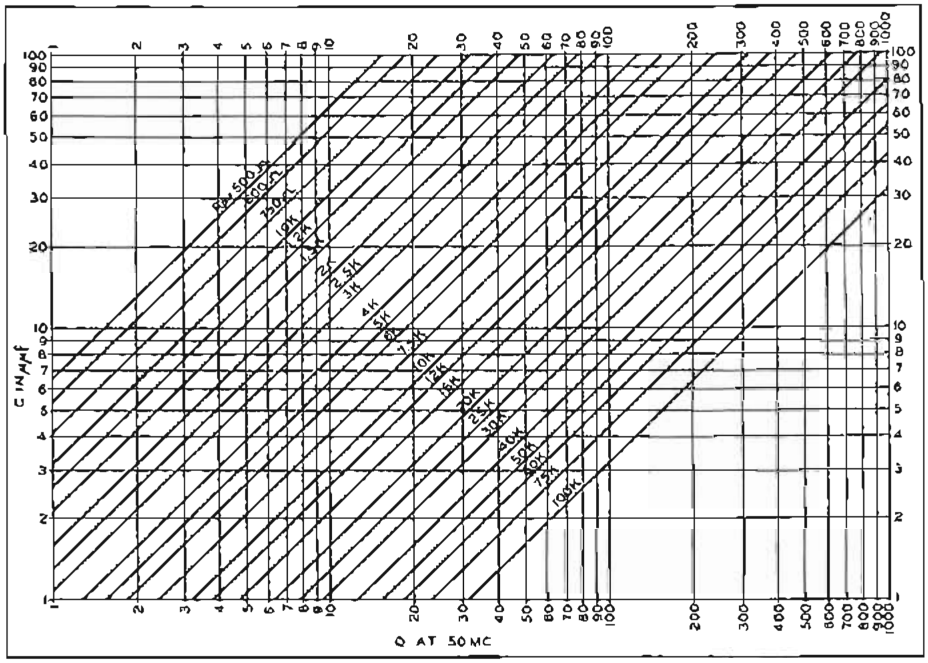


Figure 3. Calibration curves for the R_P dial.



Courtesy Pacific Semiconductors, Inc.

Figure 4. Typical family of curves for converting parallel resistance and capacitance to Q.

ured in this manner were not high enough, and while this quarter-wave-length line, when coupled through a 515-A adapter did eliminate the residual inductance of the bridge, it did not help with the residuals that were added in the jig circuit at the other end of the cable. Some thought was also given to the use of a half-wave-length line but this was also discarded because of the frequency limitation and because it did not eliminate jig residuals.

With the abandonment of the above approaches, it became apparent that an ideal method would be one that would take into account, not only the residual inductance in the 250-A RX Meter, but also the various strays and residuals presented by the fixture or jig used with the diode capacitor. Figure 1 represents a typical circuit used for this measurement.

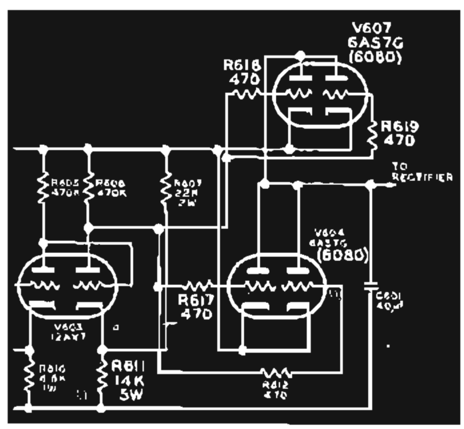
Essentially the method consists of using a coil having the necessary inductance to transpose the balance point of the bridge far enough to yield the desired capacitance, and then actually calibrating the C_p dial using various capacitors of predetermined values to obtain a correction curve to be used with the RX Meter at a given frequency. This curve is shown in Figure 2. Further, by using various values of capacitance in parallel with high quality resistors, whose resistance has been measured previously directly at the terminals of the RX Meter, at the calibration frequency, it is possible also to calibrate the R_p dial of the RX Meter to include the jig residuals. Such a calibration curve is shown in Figure 3. Using these calibration curves at a given frequency, it is now possible to measure the diode capacitor directly in the circuit as shown, and having the necessary calibration curves, the information is conveniently corrected to reflect the two parameters of the diode under test. From this information and using the curve shown as Figure 4, it is now possible to scale off the Q of the device. If so desired, the equivalent series resistance can also be determined from the corrected values of parallel capacitance and parallel resistance.

The author wishes to express his appreciation to Pacific Semiconductors, Inc., Culver City, California, for the invaluable help and data tendered.

Type 240-A Series Tube Modification

Frequent replacement of the 6AS7G regulator series tube on instruments received for repair, together with the fact that the plates of this tube were observed to glow excessively during inspection and testing operations at the factory, led to a quality control investigation of the Type 240-A Sweep Signal Generator power supply circuit. Tests revealed that due to circuit design changes and a change in the 5U4G rectifier tube furnished by our supplier, the 6AS7G series tube was being operated under overload conditions. To correct this condition, an additional regulator series tube with associated dropping resistors, has been incorporated into the 240-A power supply circuit.

Though the simplest circuit modification would be to add another 6AS7G tube, this was not possible in the case of instruments already in production, because the power supply chassis layout would not permit the addition of a 6AS7G tube. For this reason, the 6AS7G series tube in these instruments has been replaced by two smaller 6080 series tubes. In instruments produced in the future, however, the layout of the power supply chassis will be changed to accommodate two 6AS7G series tubes. A schematic diagram of the modified circuit is shown below. Components changed or added are identified in bold print.



Section of Type 240-A Power Supply Schematic

BRC WINS GOVERNOR'S SAFETY AWARD

On June 3, Governor Robert B. Meyner of New Jersey presented a governor's safety award to Boonton Radio Corporation for the remarkable

Governor Meyner Presents Governor's Safety Award to Dr. Downsbrough.

record of accumulating more than 1,820,000 man hours of work without a disabling accident. The last disabling accident at the plant occurred on April 15, 1952.

The governor's award is the highest given in a plant safety program sponsored by the Bureau of Engineering and Safety of the New Jersey Department of Labor and Safety. Since the inception of the safety program nine years ago, only nine companies in the state have received this award.

Presenting the award plaque to Dr. George A. Downsbrough, BRC president, the governor told employees:

"I congratulate all of you on this achievement. Over the course of this safety program, I have been accustomed to seeing this foremost of awards go, for the most part, to the giants among our industries. I am happy at this evidence that safety conciousness is by no means restricted to our bigger corporations."

Accepting the award, Dr. Downsbrough told the employees that the company was proud of this excellent safety record and urged that they not rest on their laurels but continue to be alert and safety conscious. On behalf of the employees, he presented the governor with an inscribed ashtray fashioned from a casting used on the Type 202 Signal Generator.

In a final note, the governor asked the employees to extend their good safety habits to the road and to their homes.

C. W. QUINN JOINS BRC FIELD ENGINEERING STAFF

Charles W. "Chuck" Quinn joined BRC as Sales Engineer in March of this year. Beginning his association with the company at this time afforded "Chuck" the opportunity to serve in the BRC booth at the IRE show where he was able to get first-hand information from our customers regarding their measurement problems. In more recent months, he has visited or otherwise been in touch with many of our customers along the east coast from the Metropolitan New York area south to the Metropolitan Washington area.

A native of New Jersey, "Chuck" served with the U.S. Navy from 1942 until 1946. He was graduated from Purdue University with a B.S. degree in Electrical Engineering in 1947. After graduation he accepted a position with Collins Radio Co., Cedar Rapids, Iowa where he was engaged in the development of receivers, frequency synthesizers, and fm transmitting equipment.

From October 1951 until he became associated with BRC, "Chuck" held engineering posts with Measurements Corporation, Boonton, N. J. During this time he gained additional experience in the development of noise meters, vacuum tube voltmeters, signal generators, and pulse generators.

"Chuck's" interests outside BRC lie in amateur radio (W2MMK) and photography. His more relaxing moments are spent in swimming and ice skating.

Considering his years of experience in the electronic measurement field, "Chuck" promises to be a valuable addition to the BRC engineering family.

Charles W. Quinn

If you are in the local area and have a measurement or applications problem, call or write "Chuck" at any time. Remember too, that BRC has Engineering Representatives throughout the U.S.A., in Canada, and overseas. A telephone call or letter is all that is required to put them at your service.

EDITOR'S NOTE

Q Meter Contest Winner

The Q of the problem coil displayed in the BRC exhibit at the IRE Show is 193. Winner of the Q Meter, with an estimate of 193, is Mr. Arno M. King of the Naval Research Laboratory in Washington, D. C.

In a letter to our General Manager, Mr. King confides that his recent, almost daily, use of the Q Meter prompted him to rule out such extreme estimates as 5 and 5,000 but from there on it was strictly a guessing game. "I have no secret procedures to offer," he writes, "and readily concede that only a very large element of luck could have placed me within 10% of the exact value, let alone the 1% tolerance which has been needed to win in the past."

Anyone who viewed the coil at the show will bear Mr. King out on this point. Our engineers contrived a series of various size coils, connected in a bridge-type configuration, which all but defied anything but educated guessing as to the value of Q. Yet, out of a total of 1200 entries, 11 persons, besides Mr. King, estimated within ±1% of the measured Q. These persons, whom we feel are deserving of honorable mention, are listed below:

Estimate

- 191 M. H. Brown, Rollan Electric Co. Chicago, Ill.
- 192 Art Ward, Livingston Electronics Essex Fells, N. J.
- 192.5 R. Lafferty, The Daven Co. Livingston, N. J.
- 192.7 C. R. Müller, Sperry Gyroscope Great Neck, N. Y.
- 193 Arno M. King, Naval Research Laboratory Washington 25, D. C.
- 195 Seymour S. West, Western Reserve University Cleveland, Ohio
- 195 C. E. Young, Naval Research Laboratory Washington 25, D. C.
- 195 Tom Crystal, 316 St. Paul St. Brookline, Mass.
- 195 Nick Lazar, Corning Glass Works Bradford, Pa.
- 195 S. Krevsky, USAS RDL Belmar, N. J.
- 195 Jerry Vogel, Marine Electric Corp. Brooklyn, N. Y.
- 195 B. Bedoinn, ARMA Corp. Garden City, N. Y.

Mr. Arno M. King winner of the Q Meter contest.

The display coil was measured on a Type 260-A Q Meter in the BRC standards laboratory by our Quality Control Engineer. Several measurements were made resulting in a computed average Q of 193 and a computed average capacitance of 352$\mu\mu$f.

Mr. King, our winner, was born in Cleveland, Ohio. He attended Washburn University in Topeka, Kansas in 1939 then transferred to Bucknell University in Lewisburg, Pa., where he received a B. S. degree in 1943. Following graduation, he was employed by the Naval Research Laboratory, Washington, D. C. where he has spent all of his professional career working on problems associated with tracking radar. Currently, he is serving as Head of the Terminal Equipment Section of the Tracking Branch, Radar Division. In addition to being a member of the IRE, he holds memberships in Tau Beta Pi, Pi Mu Epsilon, Sigma Pi Sigma, and the Scientific Research Society of America.

Our congratulations to Mr. King and sincere thanks to all of our friends who visited with us at the show.

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BRC's 25th Anniversary

The NOTEBOOK

BOONTON RADIO CORPORATION • BOONTON, NEW JERSEY

Technical Contributions of the BRC Notebook

Discussed by Engineering Educator
GEORGE B. HOADLEY, D. Sc.

Electrical Engineering Department Head
North Carolina State College

As the Boonton Radio Corporation enters a new phase of its corporate existence, it is appropriate to pause and reflect — to view in retrospect the contributions of the BRC Notebook. We could just leaf through our file, stopping to read interesting articles, but in so doing, we might lose some basic messages. So we shall try to review the contributions of the Notebook as a whole, drawing liberally on its writings, without many quotations, yet with references to Notebook articles so that the curious reader may expand his horizon.

One of the primary requirements for a quality measurement is the combination of a fine instrument and skillful operator. Instrument companies try to achieve this combination when they sell an instrument, by supplying an instruction manual. But this is usually not enough. Users must repeatedly be urged towards a full understanding of the instrument they are using and so the BRC Notebook was born to distribute[1] to users and to as many other interested persons as possible, information of value on the theory and practice of radio frequency testing and measurement. In addition, as new techniques and applications were developed, they have been brought to the readers, together with material about new instruments.

It is easy to connect equipment together and get meter readings or balances, but to be certain these really mean something, the experimenter must understand the circuit theory of his equipment. As an example of this, consider the measurement of radio receiver characteristics.[2] We plan to use a signal generator to feed the radio receiver. The signal generator has dials and meters which indicate the microvolt output, but we know that impedances must be right for these readings to be correct. But what is "right"? We look into the catalog[3] specifications of our signal generator, and under the heading "RF Output Voltage" we find "the maximum open-circuit output voltage from the

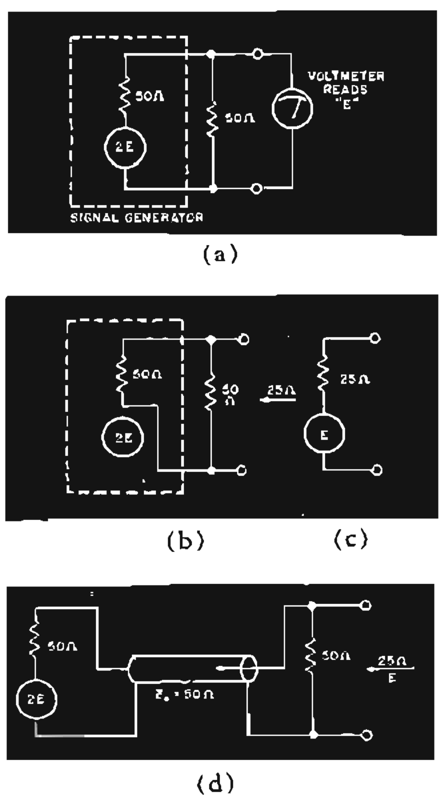

Figure 1. Thevenin's Theorem applied to signal generator use.

To Our Many Friends:

As we pause at the end of the first quarter century of our operations as designer and manufacturer of electronic test equipment, we would first like to express our sincere thanks to our customers, our employees, our suppliers, and our business associates through whose help and cooperation we have been successful. As we look through our catalogs and back issues of The Notebook, we are pleased and proud of the contributions Boonton Radio has made to the electronic industry.

However, as fine as these contributions may be, we are not satisfied and we are starting the next quarter century with a very firm resolve and with plans to improve Boonton Radio's contribution to the electronic industry. We have embarked on a program to enlarge and improve our engineering department in order that we will be able to supply all of our customers with an ever expanding and ever improving line of instruments. We are planning to erect new plant facilities, which will be equipped with the most modern and up to date machinery and tools that we can procure.

I can assure you that we at Boonton Radio are going to give our very best efforts to doing an even better job in the next 25 years and we earnestly solicit the continued cooperation of our personnel, customers, and suppliers in reaching these ambitious goals.

G A Downsbrough

G. A. Downsbrough
President

YOU WILL FIND . . .

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BNC type RF output jack at the front panel is about 0.4 volt. With the standard output cable attached, the maximum calibrated output voltage at the cable terminals is 0.2 volt. When the RF monitor meter is set to the red calibration line and the standard output cable is attached, the RF output attenuator is direct reading in microvolts and continuously adjustable from 0.1 microvolt to 0.2 volt." Under the heading "RF Output Impedances" we find "the RF output impedance of the signal generator as seen looking in at the BNC type front panel connector is 50 ohms resistive. With the standard output cable attached, the RF output impedance as seen looking in at the output cable terminals is 25 ohms resistive."

Now if we know our circuit theory, this is all very clear, and is very logical. But if our knowledge of circuit theory is a knowledge of words rather than an understanding of ideas, we must do some studying, particularly of Thevenin's theorem and transmission line theory. Thevenin's theorem states that a two-terminal network can be replaced by an emf in series with an impedance. Thus in Fig. 1(a), the emf 2E and the 50-ohm resistor inside the dotted rectangle represent a signal generator, as seen at the output jack. In this diagram, a resistor of value 50 ohms has been connected across the output jack, and a very high impedance voltmeter is shown measuring a voltage E, which is just half of the Thevenin emf of the signal generator because the two 50-ohm resistors form a 2:1 voltage divider. The impedance at the terminals, as seen by the voltmeter is 25 ohms, which is the value of the two resistances in parallel, as shown in Fig. 1(b). Consequently, when the signal generator and the 50-ohm load are considered as a unit, the Thevenin circuit at the terminals is as shown in Fig. 1(c). It is an emf E in series with a 25-ohm resistance. The open-circuit voltage is this emf, and is equal to the reading on the signal generator dials.

Sometimes we use a length of coaxial cable to connect the signal generator to the receiver under test. The heart of the problem is that this length of cable can approach and exceed ¼ wavelength, and consequently has voltage and impedance transforming properties which might nullify the calibration of the signal generator. To avoid this, we use a cable with a characteristic impedance equal to the impedance looking back into the output jack of the signal generator, and then we load this cable with a resistor of this same value. The circuit is as shown in Fig. 1(d). When the cable is terminated in 50 ohms (the value of Z_0) the input to the cable seen by the signal generator is 50 ohms so as far as the signal generator is concerned, the situation is just like that in Fig. 1(a), and the voltage input to the cable is E. If the cable is lossless, the voltage at the 50-ohm load is also E. This is the emf of the equivalent Thevenin circuit of the whole combination viewed from the terminated end of the cable. This voltage is equal to the reading of the signal generator dials, as the specification stated. The impedance of the Thevenin circuit is 25 ohms, just as the specifications stated, because from the terminals we see 50 ohms in parallel with the cable, which in itself looks like 50 ohms as it is a 50-ohm cable terminated at the signal generator end with the 50-ohm Thevenin impedance of the signal generator itself.

To truly test a receiver, the receiver must be fed from an emf of a known number of microvolts in series with an impedance equal to the impedance of the antenna to be used with the receiver. The signal generator supplies the known emf and resistance. If this resistance is different from the antenna impedance, an extra series impedance, often called a dummy antenna, must be used.

"To appreciate the logic leading to this choice let us consider the source of energy from which the combined system of the antenna and receiver is driven. Electromagnetic energy flowing in free space encounters a conductor and excites in it a voltage which acts in series with the antenna radiation resistance. Like the open-circuit electromotiveforce of a battery this voltage is available to us only in series with the internal impedance of the power source itself. The

TABLE I

Sweep Method	Advantages	Disadvantages
Mechanical Devices	High Q at all frequencies. High output possible (without buffer stage). Workable over wide range of output frequencies. Wide sweep range.	Microphonism causing frequency jitter. Non-linear sweep. Mechanical maintenance problems.
Reactance Tube	Good stability and accuracy. Non-microphonic. Linear sweep.	Limited to narrow sweep at low frequencies.
Saturable Reactor	Wide sweep range. Good stability and accuracy. Non-microphonic. Linear sweep.	Low Q at high frequencies. Susceptable to AC magnetic fields. Hysteresis effects.
Klystron Beat Method	Wide sweep range. Workable over wide range of output frequencies. Linear sweep. Non-microphonic.	Frequency jitter. Low output. Poor accuracy at low freqs.
Ferroelectric Capacitor	Non-microphonic. Linear sweep.	Excessive temperature Coeff. Low Q. Hysteresis effects.

Figure 2. Advantages and disadvantages of various means for frequency sweeping.

variation of this impedance with frequency may require a series-parallel combination of R, L, and C in the dummy antenna. Part or all of it may be contained in the signal generator output impedance."[2]

So far, we have treated the signal generator in terms of its equivalent Thevenin circuit, without much regard for the design of the equipment itself. From the user's point of view, this is wonderful if he can be sure of it! Just how Boonton Radio Corporation goes about making an instrument of which the user can be sure is an education in itself.

First there is electrical design. Take the matter of choosing a method of generating broad-band frequency deviations. All the known methods are studied and their advantages and disadvantages are determined and listed, as in Fig. 2. Then a method is chosen from this listing. For the type 240-A, the saturable reactor was selected.[4]

Then there is mechanical design, often overlooked by one steeped in circuitry. "From the moment the mechanical design of an instrument begins, a myriad of other considerations arise to confront what might otherwise seem a straightforward piece of electronic equipment. The mechanical designer must consider the electronic requirements of the development and project engineers, the functional and saleable appearance, weight and price insisted upon by Sales, and the mechanical urgencies of simple, rigid designs and drives using the proper materials. In addition the Shop must be allowed reasonable tolerances within the limitations of available processes and equipment, Assembly should have units adapted to smooth work flow, and Inspection (and the user!) needs easy accessible adjustments. Among many other factors are Purchasing's and Accounting's hopes that standard parts will be used, and Shipping's plea for enough unobstructed cabinet area to allow proper bracing in the packaging."[5]

The solution of an electro-mechanical design problem is shown in Fig. 3 This is the internal resonating capacitor which is "the heart of the Q Meter, and is an excellent example of the interdependence of mechanical and electrical design. Taking the 190-A Q unit as an example, the electronic requirements are low minimum capacitance, together with low and constant values of inductance and resistance all of which are difficult to attain in conventional designs. Mechanically the design must be extremely rigid to assure constant and accurate re-setability. The massive structure ordinarily needed to attain the last-named end is in direct opposition to the minimum capacity requirement.

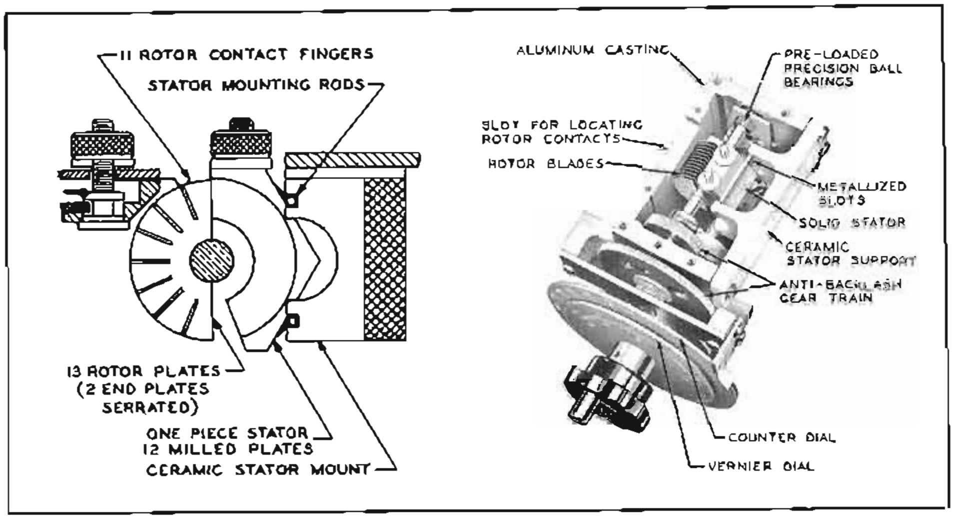


Figure 3. The 190-A capacitor solves the problem of rigidity, low capacitance, and constant inductance.

"The stator is mounted by means of two rods soldered into the metallized slots of a high quality ceramic support. Insofar as possible it floats in air dielectric. In addition, the rotor travel is restricted to less than 180°, with the included angle of the stator reduced by a proportionate amount to result in the largest possible angular gap between the two at the minimum setting. By these means the capacity at minimum was limited to the lowest value compatible with sufficient mechanical strength.

"Low and constant inductance between the stator plates is achieved by milling out a solid bar to leave only the outer shell and the plates, solidly connected with each other along their entire peripheries. A secondary result of this method is the 'built-in' shielding the shell provides from extraneous fields.

"The tandem edge wipers, contacting all the rotor blades in parallel, serve to reduce the associated inductance and resistance to a very low and nearly constant value and are rhodium plated to provide good wearing qualities.

"Where other considerations do not enter, rigidity is attained in the complete unit by mounting all the parts on a rigid cast frame. All shafts are carried on preloaded ball bearings."[5]

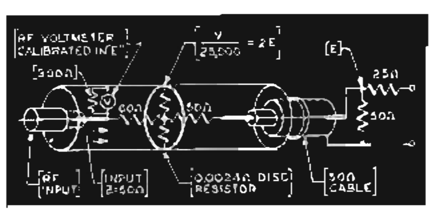


Figure 4. Schematic diagram of the RF attenuator and voltmeter in the Type 245.

Another electro-mechanical design is the attenuator system in the Type 245 RF voltage standard. As shown in Figs. 4 and 5, a disc resistor of very low value, 0.0024 ohm, is mounted across a tube, and two 50-ohm resistors are mounted as central conductors of a coaxial cable, one on each side of the disc. Study[6,7,8] of the circuit theory involved is again based largely on Thevenin's theorem, but the physical arrangement is what assures us that no leakage voltage gets into the output system.

In addition to this attenuator, the Type 245 voltage standard or Signal Generator Calibrator contains a transistorized diode voltmeter calibrated at 0.025, 0.05 and 0.1 volt, rf. This voltmeter is connected across the input of the attenuator, where the impedance is essentially 50 ohms, regardless of the output termination because the attenuation of the unit is so high (2500:1 for type 245-C or 25,000:1 for type 245-D) that the reflection of the output back into the input is negligible. Consequently, this unit can be used to check the output of a signal generator into a 50-ohm load at the three voltages mentioned above. It can also be used to

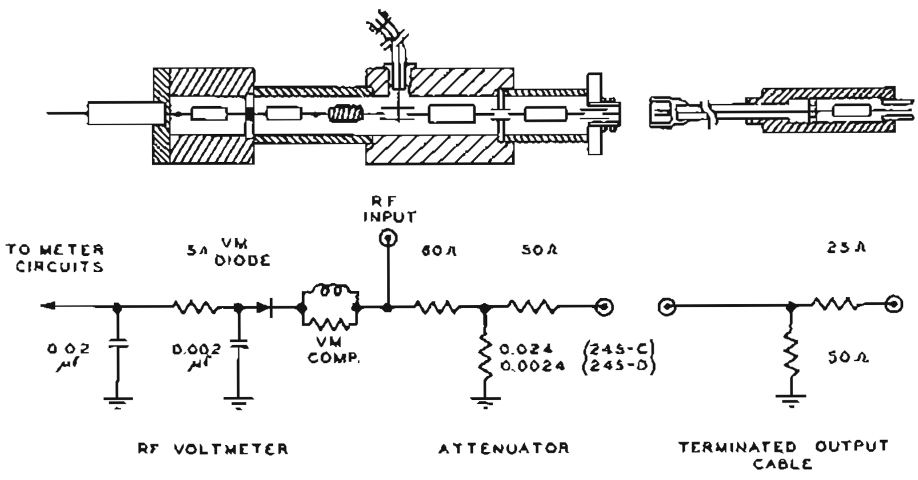


Figure 5. Cutaway view of the actual attenuator in the Type 245.

provide a standard low-level signal in the microvolt range by using the output of the attenuator when the input voltage is on one of the calibration marks. In addition, the input voltmeter is equipped with a demodulating system so it can be used to measure the percent amplitude modulation.

A good understanding of the nature and theory of the quantity to be measured is most important in the making of good measurements. In recent years, measurements of the sensitivity of noise-limited receivers has attracted considerable attention, and a body of knowledge and experimental techniques have grown up.[9] In order to make adequate measurements, one must understand the statistical nature of the noise. For example, a knowledge of the ratio of noise peaks to noise rms value is important so that the equipment can be operated with these peaks still in the linear region of the receiver. Without this information, completely incorrect results are possible. Moreover, results obtained by one type of measurement, while correct in themselves, may not be of much value in a situation not well simulated by the measurement. A thorough understanding of the theoretical basis of the measurement and the use to which the results are to be put is of highest importance.

This idea that the experimenter should understand has been a keynote in many of the theoretical Notebook articles. One "blind spot" of many engineers was tackled in the beginning with a study of the many faces of Q.[10] This quantity may be defined basically as the ratio of the total energy stored in a resonant circuit to the average energy dissipated in the circuit per radian. Such a definition implies that the circuit is in resonance, for if it were not, the stored energy would not be constant. This means that the net reactance of the circuit is zero, and yet many engineers will define Q as X/R. Such a definition can be derived from the basic one only with three assumptions. These are (1) that the circuit is a series circuit; (2) that the X is either the inductive reactance or the capacitive reactance, but not the total; and (3) that the capacitance is lossless unless its losses are included in the R.

The theory of the Q Meter has been described in many places.[10] The circuit (Fig. 6a) is a series RLC circuit coupled to a source of radio frequency current by means of a very small resistance *r* (0.02 ohms in the type 260-A), with a high-impedance vacuum-tube voltmeter connected across the adjustable air capacitor. The inductance is in the coil to be measured. With the frequency and current values held constant, the capacitor setting is changed until the voltmeter indicates the resonant rise. The maximum voltage is $\sqrt{1+Q^2}$ times the voltage obtained by multiplying the coupling resistance by the input current. Consequently, with a known value of current, the voltmeter can be calibrated in terms of Q, and of course if Q is large, the voltage is proportional to Q.

Curiously enough, the mathematical treatment of the Q Meter circuit is most easily handled by converting it to the equivalent parallel circuit shown in Fig. 6(b). This is done by applying Norton's theorem, which is the dual of Thevenin's theorem. The results are:

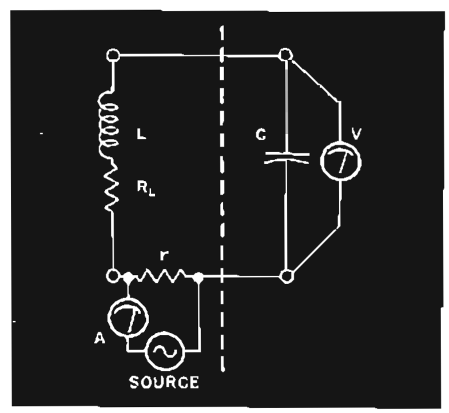
$$G = \frac{R}{R^2+\omega^2L^2} \qquad A_1 = \frac{A_r}{R+j\omega L} \qquad (2)$$

$$jB = \frac{-j\omega L}{R^2+\omega^2L^2} \qquad Q = \frac{|B|}{G} = \frac{\omega L}{R}$$

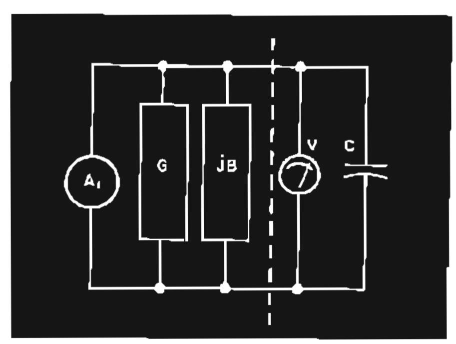
In this, A_1 is a current source, giving the same current regardless of the load.

If we vary the capacitance in the circuit, maximum voltage across the capacitor occurs when the net susceptance is zero. Thus the value of the maximum voltage across the capacitor, is given by dividing the magnitude of A_1 by G_L. When we do this, we get

$$V_c \text{ (Max. Magnitude)} = A_r\sqrt{1+Q^2} \qquad (3)$$



(a)



(b)

Figure 6. The circuit of the Q Meter and an equivalent parallel circuit for analysis.

which is exactly what was obtained in the series circuit analysis.

The relative ease with which we obtained results by using the circuit of Fig. 6(b) leads us to explore the parallel circuit ideas further.[11] Suppose we consider the coil, which should be of high Q, as a "work coil". We adjust the Q Meter and record our readings as Q_0 and C_0. Using the above ideas we get

$$G_0 = \omega C_0/Q_0 \text{ and } B_0 = -\omega C_0. \quad (4)$$

Notice that these include all losses and energy storage in the system, including the vacuum-tube voltmeter!

Now place an unknown in parallel with the capacitance, readjust the Q Meter and record the readings as Q_1 and C_1. Again we calculate

$$G_1 = \omega C_1/Q_1 \text{ and } B_1 = -\omega C_1 \quad (5)$$

But G_1, the new conductance is simply the sum of G_0 and the conductance G_x of the unknown. Also, B_1 is the sum of B_0 and the susceptance B_x of the unknown. So we can write

$$G_x = G_1 - G_0 = \omega\left(\frac{C_1}{Q_1} - \frac{C_0}{Q_0}\right) \quad (6)$$

and

$$B_x = B_1 - B_0 = \omega\,(C_0 - C_1). \quad (7)$$

From these we can get the equivalent parallel resistance and reactance of the unknown as simple reciprocals, thus

$$R_{Px} = \frac{1}{G_x} \text{ and } X_{Px} = \frac{-1}{B_x}. \quad (8)$$

If we want the equivalent series resistance and reactance we write

$$R_{Sx} = \frac{G_x}{G_x^2+B_x^2}$$

and (9)

$$X_{Sx} = \frac{-B_x}{G_x^2+B_x^2}$$

The whole idea[12] that an unknown impedance or admittance can be looked on as either an R_S and X_S in series or as an R_p and X_p in parallel may be new to some readers. Actually no coil is really a series connection or a parallel connection, but is a combination of both, with some capacitance thrown in for good measure.[13] The source of the series resistance is the wire of which the coil is wound and the sources of the parallel resistance are eddy currents in shields, core loss in magnetic materials, and dielectric loss in insulating materials (which is the easiest to reduce to the vanishing point). The parallel resistance in an actual coil becomes predominant if the frequency is carried high enough, so the Q of every coil will reach a maximum at some frequency, (where series and parallel losses are equal), and will decrease as the frequency is raised above this value. At any single frequency, however, we can consider the loss as being all in series resistance or as being all in shunt resistance. We cannot properly divide the loss between the two unless we know the behavior of the coil as a function of frequency.

Parallel equivalences are particularly valuable in many bridge circuits and circuits derived therefrom. One reason is that stray capacitances are most easily treated by such means. Another is that certain circuits just naturally give admittances rather than impedances. Consider the general impedance bridge in Fig. 7(a). The balance equation is

$$Z_1 Z_3 = Z_2 Z_4. \quad (10)$$

If we solve for $1/Z_4$, we get

$$\frac{1}{Z_4} = Y_4 = \frac{Z_2}{Z_1 Z_3} \quad (11)$$

If we use Z_1 and Z_3 as fixed components, the values of the components of Y_4, which are G_4 and B_4, can be found in terms of the series components of Z_2. If we apply this idea to the Schering bridge circuit of Fig. 7(b), we have

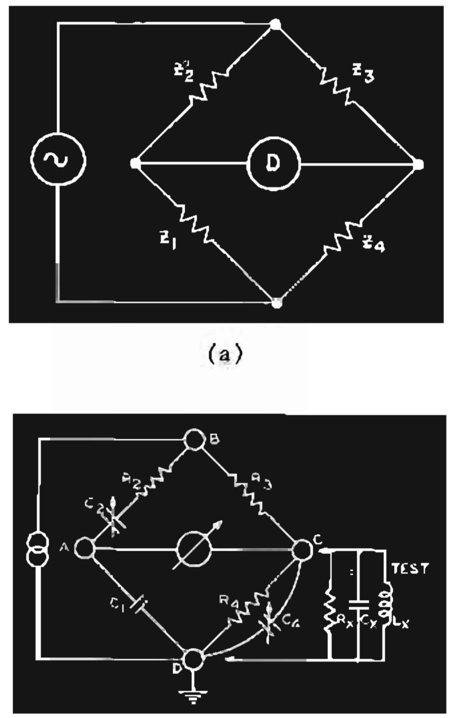
$$G_4 + jB_4 = \frac{j\omega C_1}{R_3}\left(R_2 + \frac{1}{j\omega C_2}\right) \quad (12)$$

If B_4 is a capacitance, then we have

$$G_4 = \frac{C_1}{R_3 C_2} \text{ and } C_4 = C_1 \frac{R_2}{R_3}. \quad (13)$$

The RX Meter uses a modification of this circuit with a special balanced transformer so that the two sides of the bridge can be supplied with equal voltages 180 degrees out of phase with each other, as is shown in Fig. 8. The result[14] of this is that the detector can be operated with one side grounded and with the other side coupled to the usual bridge detector points by a pair of small capacitors. The balance equations are identical with those for the bridge.

This circuit can be used in a manner similar to the way we used the Q Meter for parallel measurements. We balance the circuit with nothing connected across the "Test" terminals. Then we read th evalues of C_2 and C_4. Let us call these C_{20} and C_{40}. We have balance equations as follows



(a)

(b)

Figure 7. The basic impedance bridge circuit and the Schering bridge circuit.

$$G_{40} = \frac{C_1}{R_3 C_{20}}$$

$$\text{and} \quad B_4 = \omega C_{40} = \omega C_1 \frac{R_2}{R_3}. \quad (14)$$

Now we connect an unknown admittance $G_x + jB_x$ in parallel with the elements already in arm 4, and we rebalance the circuit. The new readings of C_2 and C_4 we will call C_{21} and C_{41}. This time the balance equations become

$$G_{41} = \frac{C_1}{R_3 C_{21}} \qquad (15)$$

and $B_4 = \omega C_{41} + B_x = \omega C_1 R_2 / R_3$ (16)

The first of these is a new one, but the second simply indicates that the total B in arm 4 is still the same. Drawing on this fact, we can write

$$B_x = \omega(C_{40} - C_{41}). \qquad (17)$$

Since the value G_{41} is the sum of G_{40} and G_x, we can write

$$G_x = G_{41} - G_{40} = \frac{C_1}{R_3}\left(\frac{1}{C_{21}} - \frac{1}{C_{20}}\right) \qquad (18)$$

The value of G_x is determined by the values of C_2 from two balances and the values of C_1 and R_3. If these latter terms are known, the values of C_{20} can be marked as zero G and the scale of C_2 can be calibrated in mhos of conductance. Alternately, the scale of C_2 can be calibrated in terms of parallel resistance in ohms. Then the zero point of G becomes the infinity point of R_p. This is done in the RX Meter.

The value of B of the unknown involves only the value of angular frequency, ω, and the values of two capacitor settings. The value C_{40} can be marked as zero, and the scale can be calibrated in capacitance values above and below this zero. Then the C_4 dial reads the value of parallel capacitance which has a susceptance equal to the susceptance of the unknown. Thus the positive range of the C_p dial is the region of actual capacitance below the value of C_{40}. The negative range of the C_p dial indicates that the unknown is inductive.

The inductive range of the RX Meter can be extended to higher values of B, corresponding to lower values of L, by adding known capacitors in parallel with the unknown. Other methods of extending the range of measurements are also known,[15] and all of them depend on the experimenter's knowledge of the mathematics of electric circuits for their usefulness.

Transmission line parameters can be found from measurements made on the RX Meter.[16] Application of the circuit theory of transmission lines leads to measurement procedures which enable impedances at a distance from the terminals of the RX Meter to be determined.[17]

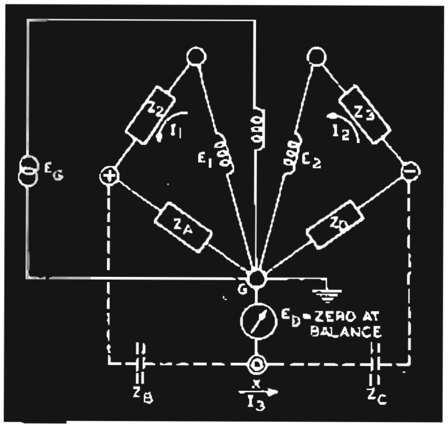


Figure 8. Basic circuit of the RX Meter.

In the area of transistor parameter measurements, circuit theory and measurement technique reach one of their finest meetings. From the circuit theory viewpoint, the problem is to find expressions for transfer parameters in terms of two-terminal impedances or admittances. From the measurement technique side, the problem is to measure the desired impedances or admittances with proper bias currents supplied to the transistor. The RX Meter is ideally suited for this, because it can carry fifty milliamperes direct current applied at the unknown terminals. With appropriate jigs to hold the transistor and supply adequate biasing, the measurements can be quickly made.[18]

The details on all of these things have been brought to the reader of the BRC Notebook. The presentations have been challenging and have encouraged many of us to delve into new theoretical points and experimental procedures. We, the readers of the Notbook are the richer for this, and we look forward to future issues of the Notebook with great expectations. Truly the well designed nistrument and the skillful user produce remarkably useful results.

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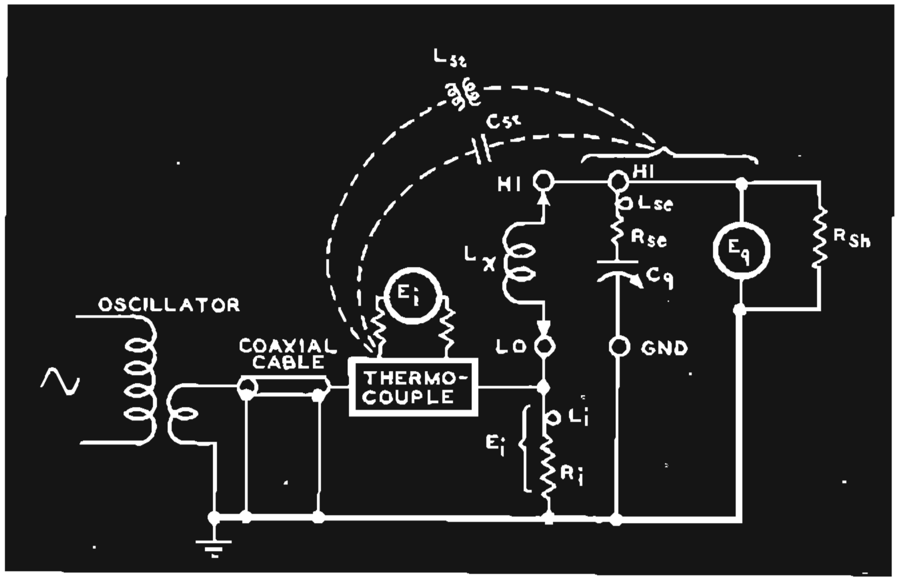
The Evolution of the BRC Q Meter

LAWRENCE O. COOK, *Quality Control Engineer*

Q is defined as 2π times the ratio of energy stored to the energy dissipated per cycle.[1,2] In electronics, the concept of Q is commonly used to designate the ratio of series reactance to series resistance of a coil ($Q = 2\pi fL/R$) or of a capacitor ($Q = 1/2\pi fCR$). While these and other relationships involving Q have been used in radio and electrical engineering for a great many years, the expression Q and its numerical value did not come into popular usage until early in the 1930's, during the time when the broadcast receiver industry was growing at a fast pace and a rapid means for measuring Q was sorely needed. Seeking to fulfill this need, the founders of Boonton Radio Corporation demonstrated the first Q Meter at the IRE Meeting in Rochester, N. Y. late in 1934.



R_i Q circuit injection resistor (100-A and 160-A, 0.04 ohm; 260-A, 0.02 ohm)
L_i Self inductance of R_i
E_i Injection voltage and meter for same
C_{st} Stray capacitive coupling
L_{st} Stray inductive coupling
L_x Coil under test
C_q Calibrated internal resonating capacitor
L_{se} Q circuit residual inductance (100-A, 0.08 μh, 160-A and 260-A, 0.015 μh)
R_{se} Q circuit residual series resistance
R_{sh} Q circuit residual shunt resistance
E_q Q vacuum tube voltmeter
HI — LO External coil terminals
HI — GND External capacitor terminals

Figure 1. Q Meter Fundamental Circuit — Including Residuals

Fundamental Q Meter Circuit

The early model Q Meter employed the "voltage step-up" (also known as "resonance rise") method of Q measurement still used in current models. A simplified schematic of the fundamental circuit is shown in Figure 1. The Q of a resonant circuit, comprising a capacitor (C_q) contained in the Q Meter and an external coil (L_x), is measured by impressing a known voltage (E_i) in series in the circuit and measuring the voltage (E_q) across the capacitor when the circuit is resonated to the frequency of the impressed voltage. Q of the circuit is the ratio E_q/E_i. With E_i known, the voltmeter (E_q) may be calibrated directly in Q and, because the circuit losses occur mostly in the coil, the Q indication obtained closely represents the Q of the coil. By inserting low impedances in series with the coil or high impedances in parallel with the capacitor, the constants of unknown circuits or components may be measured in terms of their effect on the original circuit Q and tuning capacitance.

Basic Design Problems

Though the fundamental Q measurement method just described is extremely simple, the achievement of accurate results over a wide range of frequencies requires the solution of several basic problems.

1. The injection voltage system must be frequency insensitive.
2. Stray coupling occurring between the oscillator (including the injection system) and the Q measuring circuit must be reduced to a negligible value.
3. The Q measuring circuit residual inductance and series and shunt resistive losses must be minimized. Included are input circuit losses in the VTVM which measures the voltage across the resonating capacitor.
4. The oscillator waveform must be relatively free of harmonics.

These factors have been strenuously dealt with in Q Meter design and, over a period of many years, much progress has been made which benefits the user in terms of improved accuracy. Some of the results of this progress, in the LF and lower VHF range of frequencies, will be shown in the remaining paragraphs which trace the development of the Q Meter from the first model marketed, the Type 100-A, to a model currently in production, the Type 260-A.

Type 100-A Q Meter

The Type 100-A Q Meter was the first model to be sold (in early 1935) and is readily recognized because of its 45° panel slope.

A Type 45 tube operated in a tuned-grid oscillator circuit having tickler feedback. Turret selection of 7 calibrated frequency ranges provided a total range of 50 kc to 50 mc, the entire oscillator assembly being shielded to provide isolation from the Q measuring circuit.

The oscillator output current, controlled by adjustment of the dc plate voltage, was fed through a coaxial cable to a thermocouple and then through a 0.04-ohm "voltage injection resistor". This resistor, a closely shielded resistance strip, provided a low value of self inductance so that the voltage drop developed across the resistor was rela-

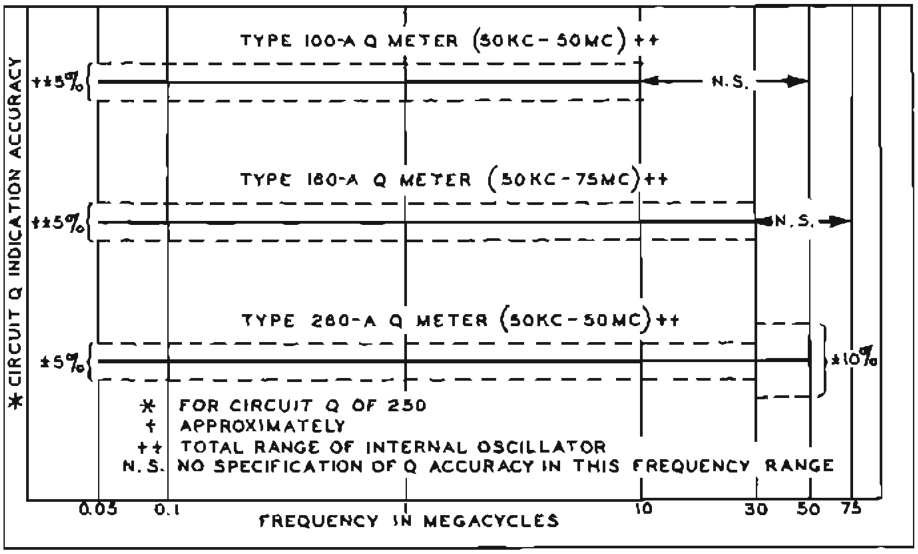


Figure 2. Specification of Circuit Q Indication Accuracy vs Frequency

tively independent of frequency. The thermocouple operated a 3-inch dc meter which was calibrated at two Q range settings in terms of the voltage developed across the resistor at dc and at low frequencies.

The Q measuring circuit included a single-section, receiver-type capacitor having aluminum plates which provided a calibrated capacitance range of 37 to 460 $\mu\mu$f. Impregnated mica insulation was employed in the capacitor for low loss purposes, each mica insulator being tested under conditions of 90% relative humidity. The vernier capacitor was in a separate frame and employed similar insulation.

External terminals for connection of the coils and capacitors to be tested were of the commercial, nickel-plated type mounted on impregnated mica insulators.

The Q voltmeter circuit employed a triode tube operating as a form of "plate rectifier" with provision for zero balance of the cathode current meter. The Type 2A6 tubes were individually selected in the operating circuit for high input resistance at rf, normal input capacitance, low direct grid current, and normal rectified dc output versus ac signal voltage. The 3-inch meter was calibrated in two ranges of circuit Q (0 to 250 and 0 to 500) in addition to signal volts. The VTVM grid return resistor (100 megohms at dc) was of a design chosen for high effective resistance at rf.

The Q Meter power supply was of the unregulated type commonly used at that time.

Performance of this instrument, for the Q measurement of inductors and capacitors, was generally satisfactory at frequencies up to 10 mc, as shown in Figure 2. For increasing frequencies (i.e., above 10 mc) the accuracy gradually worsened because of the effects of injection resistor inductance, stray coupling between the thermocouple system and the Q measuring circuit, and Q measuring circuit residual inductance and residual resistance.

Type 160-A Q Meter

Increased use of higher frequencies in the communications field created a need for improved Q Meter accuracy at these higher frequencies. To meet this need, a new model, the Type 160-A Q Meter, was developed and introduced in 1939, superseding the Type 100-A. In addition to greatly improved accuracy, this model had a 15° panel slope and a considerably different appearance.

The oscillator was essentially the same as used in the Type 100-A instrument except that an eighth frequency range (50 to 75 mc) was added. Mechanical reliability of the shielding was also improved.

The injection system provided a completely shielded thermocouple with the injection resistor being included in the same shielded assembly. Stray coupling to the Q measuring circuit was thus greatly reduced. Additional division lines on the "Multiply Q By" meter scale plate provided a wider (20 to 625) range of circuit Q measurements and improved accuracy.

The Q measuring circuit resonating capacitor, calibrated range 30 to 460 $\mu\mu$f, was of a design especially developed to provide low residual inductance and resistance for this purpose. Main and vernier capacitor sections were included in a single frame to avoid the inductance of a connecting lead. The main rotor and stator were split into two equal sections, the rotor being "center fed"; i.e., to provide a shortened current path, the rotor was grounded by fingers contacting a disk located on the shaft midway between the two sections. Rotor and stator plates fabricated of copper

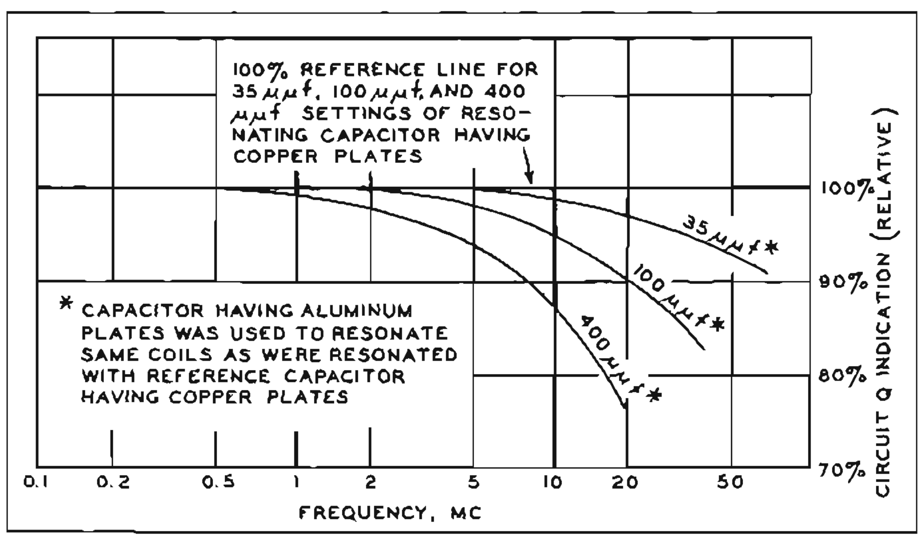


Figure 3. Circuit Q Indication of Q Meter vs Resonating Capacitor Plate Material

provided lowered rf resistance as compared to the aluminum material formerly used. (See Figure 3.) The stator insulators of this capacitor were at first of impregnated mica, but a subsequent design modification substituted pyrex glass balls for improved electrical reliability.

External terminals (Figure 1) were of gold-plated copper to provide high conductivity. To permit shortened internal leads the panel slope was changed from 45° to 15° and the external terminals were mounted integrally with the capacitor. The residual inductance of this unit, measured at the COIL terminals, was 0.015 μh, a considerable reduction from the Type 100-A inductance of 0.08 μh.

For improved readability, a 4-inch meter was used in the Q-VTVM. The meter was critically damped to eliminate the pointer over-swings found in the Type 100-A. The power supply was of the conventional unregulated type.

While the Type 160-A instrument achieved a wide usage in the electronic field and offered greatly improved accuracy at the higher frequencies over its predecessor the Type 100-A (Figure 2), its accuracy at frequencies above 30 mc was limited and the thermocouple factor of safety was low.

Type 260-A Q Meter

Progress in the electronic and instrument art indicated that a revised Q Meter of refined design and improved accuracy was needed. To meet this need, the Type 260-A Q Meter, superseding the Type 160-A, was developed in 1953 and is still being produced. This model is similar in shape and size to the Type 160-A but is recognizable by its recessed dials.

An oscillator of complete redesign employs a modern tube and modern components. The circuit is designed for low harmonic content. Output current control is in the low wattage screen grid circuit. Turret selection of eight calibrated frequency ranges provides a total coverage of 50 kc to 50 mc.

The thermocouple and "Multiply Q By" meter circuit have been redesigned for a lower thermocouple operating temperature and consequent greater safety factor. Thorough shielding is employed and a 4-inch meter with mirror scale provides greater accuracy of setting the injection voltage.

The injection voltage resistor is a 0.02-ohm annular type providing essentially noninductive performance at frequencies as high as 50 mc, a welcome change from the inductive voltage rise experienced with the shielded resistance strip type of resistor used in Q Meters Type 100-A and Type 160-A. Figure 4 plots the error in the Type 160-A largely attributable to this cause. The error in the 260-A is negligible. The lowered resistance value of 0.02 ohms in the Type 260-A versus 0.04 ohms in the Types 100-A and 160-A (this resistor being in series with the Q measuring circuit) raises the measured circuit Q by as much as 15% at the higher frequencies[3]. Thus the circuit Q and the coil Q are brought into closer agreement.

The resonating capacitor (calibrated for a range of 30 to 460 μμf) is of the same design as was employed in the later 160-A's except that the external terminals are supported on a teflon insulator for improved uniformity, strength, and reliability. The direct reading capacitance scale is supplemented by a direct reading inductance scale for use at specified frequencies.

In addition to the usual main Q scale (40 to 250), the Q indicating meter provides a low Q scale (10 to 60) and a ΔQ scale (0 to 50). These direct reading scales, when used in conjunction with the "Multiply Q By" meter (range X1.0 to X2.5) provide a circuit Q measurement range of 10 to 625. Each meter employs a mirror scale for the elimination of parallax error.

The power supply voltages are regulated by a voltage stabilizing transformer and "glow tubes", thus provid-

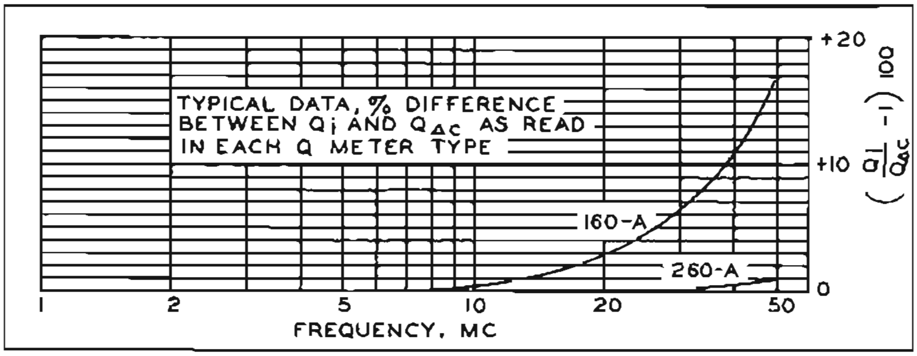


Figure 4. Q Indicated (Q_i) vs Q by Delta-C ($Q_{\Delta C}$) Method

ITEM	100-A Q METER	260-A Q METER
Oscillator Harmonic Content	High at some frequencies, causing Q indication error.	Low at all frequencies, negligible Q error.
Oscillator Output Thermocouple (A) Overload factor (B) Shielding	(A) Small, susceptible to burnout. (B) Poor, causing Q indication error.	(A) Large, burnout rare. (B) Good, negligible Q indication error.
Injection Voltage Resistor	Inductance causes Q indication error at higher frequencies.	Inductive effect negligible; lowered resistance value improves circuit Q.
Resonating Capacitor (Q Measuring Circuit)	Receiver type with aluminum plates, vernier separate, external COIL and COND terminals separately mounted, impregnated mica insulation.	Specially designed, silver plated copper plates, rotor current center-fed, vernier in same frame, external COIL and CAP terminals integrally mounted, teflon and pyrex insulation, residual inductance and resistance greatly reduced.
Circuit Q Measurement Range	10-500	10-625; includes low Q Range and ΔQ range for better accuracy.
Meters	3 inch	4 inch, mirror scale.
Power Supply	Unregulated	Regulated; meter indications stabilized against line voltage fluctuations.

Figure 5. Highlights of Q Meter Design Differences

ing stability of meter indications in the presence of power line voltage fluctuations.

Figure 5 offers a quick review of the design highlights which contribute to the improved performance of the Type 260-A Q Meter. Note that the Q indication accuracy specification now extends upward to include the full frequency range of 50 kc to 50 mc (Figure 2).

Accessory Inductors

The Type 103-A Inductor has long been available as a "work coil" for use in Q Meter measurement of capacitors and other components. The more recently introduced Types 513-A and 518-A Q Standards provide a ready means for the user to check the accuracy of his Q Meter, thus assuring instrument accuracy at the time of Q measurement.

Conclusion

Twenty-five years of electronic engineering effort has brought forth many advancements in the electronic field. We believe that Q Meter design has kept pace in terms of improved accuracy of measurement, improved reliability, and improved stability of operation.

References

1. Moore, W. C., "The Nature of Q", BRC Notebook No. 1, Spring 1954.
2. Stewart, John L., "Circuit Theory and Design", John Wiley and Sons, Inc., 1956, p. 344.
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Boonton Radio Corp. Merges With Hewlett-Packard Co.

Boonton Radio Corporation recently became the newest member of the Hewlett-Packard Co. family of Palo Alto, California as a wholly-owned subsidiary and joined other companies operating under similar status including: F. L. Mosely Co., Pasadena, California, makers of strip chart and X-Y recorders; Palo Alto Engineering Co. of Palo Alto, manufacturers of transformers, potentiometers, and other components; and Dymec, Inc. of Palo Alto, now a division of H-P, manufacturers of precision electronic measuring equipment and systems.

Announcing the arrangement between the two companies, Dr. George A. Downsbrough, President of BRC, emphasized that BRC would continue to operate as a separate company with no changes in either management or personnel contemplated. He stated that plans for a new plant, which were underway prior to the merger, would be accelerated. These remarks were echoed by Messrs. Hewlett and Packard who visited BRC to personally welcome BRC into the H-P fold.

BRC is looking forward to expanding its line and development activities through the use of H-P development of components and close liaison with their development activities.

From a humble beginning in 1939, the Hewlett-Packard Co. has grown to be one of the largest manufacturers of electronic test equipment in the world. The company produces more than 300 different instrument types, including oscillators, voltmeters, signal generators, waveform analyzers, microwave and waveguide test instruments, and oscilloscopes. These products are sold to more than 3000 business organizations throughout the world, with the government, through its various agencies, constituting one of the largest single users of H-P equipment.

In addition to its expansion in this country, Hewlett-Packard has established subsidiaries in Germany and Switzerland. The Stuttgart, Germany plant is due to begin production of H-P instruments for the Continental market early next year. This plant will be the manufacturing outlet of the company's wholly-owned Swiss sales organization which was set up last January.

Boonton Radio Corporation is proud to be a member of the progressive Hewlett-Packard family and is looking forward to expanding with a fast growing industry.

NEW PLANT SOON FOR BRC

The purchase, in December of 1958, of a 70-acre tract in Rockaway Township, a few miles Northwest of the present plant, was the first step towards a long-range expansion program in effect at BRC. The tract is located less than a mile from the recently completed interchange on the newly-aligned Route Route 80 interchange, and is easily accessible to BRC's 150 employees who mostly reside in the area. Ample room is available on the new site for enlarged plant construction and recreation facilities.

More recently, BRC has engaged an Architectural firm to draw up plans for the new building. These plans should be completed in the near future and plans are that ground will be broken early in 1960.

The new building will be a modern, single-story structure providing at least 50,000 square feet of space, or more than double the area of the present plant. The Engineering, Production, and Administration Departments are being layed out with ample room allowed for future expansion. The plant will be fully equipped with the most modern machinery and tools obtainable. Facilities are expected to include air conditioning and a cafeteria.

Details on the new plant will be the topic of a future Notebook article which will be published as soon as the building plans are firm.

An aerial view of BRC's new plant site.

Looking Back 25 Years With BRC

Boonton Radio Corporation was established in 1934, but the scene was set before that time, just after the end of the First World War. Many of the concepts that made wireless communication possible were discovered before the War, but it was during the War that new ideas were evolved and that considerable practical experience was gained in the use of these new ideas. When the War was ended, the public was beginning to appreciate the usefulness of transmitting intelligence over distances without wire and was taking a keen interest in its development.

Manufacturers, recognizing this intense interest, began devoting time and money to the development of improved radios and radio devices. They found it necessary to obtain component parts which were new to most of them, and for which they were sadly lacking in testing methods.

These were the conditions under which the Boonton Hard Rubber Company in Boonton, New Jersey established a group for dealing with the new problems. The staff of the newly formed group, including at first just one physicist, produced coil forms and other radio components using insulating material. Later, additional technical people were employed, and the work of general engineering consultation was undertaken. This type of work naturally led to an understanding of basic test equipment requirements.

Late in 1934, Mr. William D. Loughlin, who had been President of Radio Frequency Laboratories and who was one of the industry's pioneers, together with several associates, purchased one of the buildings which had been used by the Boonton Hard Rubber Company for its radio activities, and formed the Boonton Radio Corporation. The new company concentrated its engineering skill toward the development of new measuring equipment sorely needed by the radio industry at that time. For example, there was at that time a specific need for a quick and accurate method for measuring Q. Q measurements were being made indirectly by means of bridges which measured the effective reactance and resistance concerned. These measurements were too often subject to error because of the involved techniques required, and were time consuming.

Larry Cook, with BRC since 1935, is shown with a preproduction model of BRC's first Q Meter and the current Type 260-A Q Meter.

It was this Q measuring problem, in fact, which led to the development of the first Q Meter. Manufacturers were confronted with the costly annoyance of producing coils that would meet all of the requirements when tested at their own plant, only to be rejected because they did not pass inspection at their customers plant. A need for approved standards was evident and this was among the first assignments of the BRC engineers.

First Q Meter Introduced

In November of 1934, Boonton Radio Corporation presented at the Institute of Radio Engineers' Fall meeting in Rochester, New York, a model of the Company's first Q Meter. This instrument covered the frequency range of 50 kc to 50 mc and was known as the Type 100-A. With this instrument, Q measurements were made simple and rapid. It was also capable of many other valuable laboratory measurements on basic components and circuits. The Q Meter was immediately accepted as a standard by the radio industry and research laboratories. Over the years improved models (the Types 160-A and 260-A) of Q Meters in this frequency range have been introduced.

In 1941, a high frequency model (30 to 200 mc) of the Q Meter, known as the Type 170-A ,was introduced. This instrument was followed by the QX Checker (Type 110-A), similar to the Q Meter but designed specifically for rapid production testing of components with laboratory accuracy. This instrument was very easy to operate and could be handled by unskilled personnel.

Today, a faster, more versatile instrument, the Q Comparator Type 265-A has replaced the QX Checker and the low frequency Q Meter (Type 260-A) and high frequency Q Meter (Type 190-A) are in very broad use.

FM and HF Test Equipment

Just before the Second World War, BRC began development work on a frequency-modulated signal generator to meet the demand for test equipment for the new fm communication equipment. A model of this generator was first presented in 1940 at the Institute of Radio Engineers' meeting in Boston. Several models of these FM Signal Generators, developed by BRC, were used during the war by military and commercial customers.

During the Second World War, BRC provided large quantities of the standard commercial equipment, which had been previously designed for its commercial customers, to the Military Services for use in the War effort. In fulfilling its patriotic duty, the Company prepared a microwave pulse modulated rf signal generator for manufacture. A large number of these instruments were produced for the Military Services for use in testing radar systems. This Signal Generator is still used by the Military.

At the end of the War the FM Signal Generator was redesigned to permit coverage of a wider frequency range, to include AM as well as fm, and to obtain deviations in frequency which did not vary with carrier frequency. This instrument had very low leakage and a wide selection of accurately calibrated output voltages. It soon became the standard in its field.

Aircraft Navigation Test Equipment

In the 1940's, the aircraft transportation field was developing more accurate methods of navigation and better methods of landing in bad weather. A system for solving these problems was approved by the Civil Aeronautic Administration and put into use both commercially and by the military services. During the development phases of this

Aircraft Navigation and Landing System, BRC was asked to develop test equipment of unusual accuracy for testing several of the receivers involved. A signal generator for navigation equipment was produced in 1947, to be followed a short time later by an equipment for testing receivers used in landing aircraft. A more advanced model of the "Glide Path" testing equipment for the landing of aircraft was produced in 1952.

The Productive 1950's

In the last decade, the Company's efforts have been directed toward the development of other self-contained, broad-band, flexible instruments. The RX Meter, introduced in 1953, measures parallel resistance and parallel reactance of two-terminal networks over the LF and VHF ranges. This instrument has contributed to the development of the diffusion-based transistor, which in turn is playing a large part in America's missile and satellite program. More recently, the instrument has ventured from the world of missiles and electronics to become one of the medical world's latest weapons against disease. The University of Pennsylvania School of Medicine put the instrument to work measuring the electrical properties of human and animal tissue.

In addition to the RX Meter, the 1950's brought about the redesign of both the low-frequency and high-frequency Q Meters, increasing the usefulness and accuracy of these instruments, and the development of a Film Gauge for measuring film thicknesses.

In the last 2 years, BRC has offered three new instruments to the electronic industry, all of which were announced at the 1959 IRE Show in New York. These instruments include a Q Comparator designed to give instantaneous and simultaneous readout of Q, inductance, and capacitance on a cathode-ray tube, for production testing of components; a Signal Generator Calibrator which provides accurately calibrated RF output voltages for testing signal generators and receivers and measures percent AM; and a new Signal Generator which provides improved frequency stability over a wide range.

Expansion of Personnel and Facilities

As the BRC instrument line increased, naturally the plant had to be enlarged and the number of employees increased. The original RFL building has been lost in a series of major additions, the last of which was completed about eight years ago. The plant now has about 23,000 square feet of working area and houses all of the Company's operations. Company personnel has grown from as few as six persons to a payroll which now includes more than one hundred and fifty employees.

During the past year, the Company has taken two major steps which figure to play an important part in future expansion. The first was the purchase of a 70-acre plant site on which a new plant will be erected in the very near future, and the second was joining forces with the Hewlett-Packard Company to become a wholly-owned subsidiary of that firm.

Quality Key to Success

Over the years, BRC has built electronic tools which have come to be recognized throughout the world for their superior quality. We attribute this success to the fact that our instruments receive expert care from the drafting boards to the final test department, and to our policy of building only those instruments which have been pioneered in our own laboratories.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

A Transistor Test Set

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The Boonton Radio Corporation Transistor Test Set Type 275-A is an instrument for measuring small signal parameters of a transistor. The common base short circuit current gain Alpha (h_{fb}), the common emitter short circuit current gain, Beta (h_{fe}), and the transistor input impedance common base with the output short circuited (h_{ib}) are measured. The instrument differs from the conventional transistor test set in that the parameters measured are determined by the position of a linear potentiometer required to produce a minimum reading on a sensitive detector. It does not therefore require the calibration of, nor depend upon the constancy of level of the ac signal applied for measurement purposes.

Reference to the photograph (Figure 1) reveals the unit as a self contained, line powered, bench type instrument with well placed, easily operated controls.

The Block Diagram (in Figure 2) shows the basic components of the 275-A. It contains, beside the basic measuring circuit, which is the heart of the instrument, a simple 1-kc oscillator, a sensitive detector of wide dynamic range, and three power supplies.

Theory of the Measurement

The Null techniques used in measuring Alpha and Beta in the 275-A Test Set were devised by D. E. Thomas of the Bell Telephone Laboratories(1). The equivalent circuit of the basic measurement is illustrated in Figure 3. When the potentiometer, which is linear, is moved to a position where the signal voltage between the base of the transistor and ground is a minimum, the small signal Alpha of the transistor is a linear function of the angular position of the potentiometer.

(1) These null techniques are incorporated in the Type 275-A Transistor Test Set under Western Electric License.

Figure 1. Transistor Test Set Type 275-A.

If the circuit of Figure 3 is modified to the circuit of Figure 4 in which R_2 is a fixed resistance and the potentiometer (R_T) is now used as a variable resistance, the small signal β of the transistor is a linear function of the angular position of the potentiometer. Finally, h_{ib} is measured in the variable ratio arm bridge shown schematically in Figure 5. The potentiometer ($R_1 + R_2$) of the measuring circuit provides the two variable ratio arms, and the remaining arms are the standard resistance R_1 and the unknown resistance h_{ib}. The potentiometer position required to make V a minimum is then calibrated in terms of the resistance value of h_{ib}.

Specifications

Now that the basic principle of operation has been explained, let us consider the range of values over which the parameters α, β, and h_{ib} may be measured. Note particularly that these parameters may be measured down to 0.01 ma. emitter current.

YOU WILL FIND . . .

ESTIMATE THE Q
WIN A Q METER

Yes, that is all that is necessary to win the factory reconditioned Type 160-A Q Meter which will be on display in the BRC exhibit at the IRE show to be held in the New York Coliseum from March 21st through March 24th. The Q Meter will be awarded to the person whose estimate is closest to the actual measured Q of the coil to be displayed with the Q Meter. Complete information will be furnished by engineering personnel on duty in BRC Booths 3101 and 3102.

THE BRC NOTEBOOK is published four times a year by the Boonton Radio Corporation. It is mailed free of charge to scientists, engineers and other interested persons in the communications and electronics fields. The contents may be reprinted only with written permission from the editor. Your comments and suggestions are welcome, and should be addressed to: Editor, THE BRC NOTEBOOK, Boonton Radio Corporation, Boonton, N. J.

Alpha (h_{fb}):
RANGE: 0.001 to 0.990 or 0.9000 to 0.9999
ACCURACY: for $f_\alpha \geq 500$ kc. (*)
Better than ±1% for α from 0.100 to 0.990
Better than ±0.5% for α from 0.9000 to 0.9999
for any α % error $= \pm\left(0.1 + \frac{.09}{\alpha}\right)$

Beta (h_{fe}):
RANGE: 1 to 200
ACCURACY: ±2% from 7 to 200 for $f_\alpha \geq 500$ kc. (*)

h_{ib}:
RANGE: At x0.1; 0.30 to 30 ohms
At x1.0; 3.0 to 300 ohms
At x10.0; 30.0 to 3000 ohms
ACCURACY (stated for linear resistors): ±3%

Internal Power Supply:
EMITTER CURRENT (I_E): 0.01 to 100 ma. in 10 overlapping ranges
COLLECTOR VOLTAGE (V_{CB}): 0 to 100 volts in 6 overlapping ranges

External Power Supply Capability:
I_E = 5 amperes max. (for Alpha only)
Note: The base current should not exceed 100 ma. in any case.
V_{CB}: not to exceed 100 volts dc

Meter Accuracy: ±1½% full scale

*f_α ≡ the frequency for which
$$|\alpha| = \frac{1}{\sqrt{2}}\alpha_o$$
where α_o = forward short circuit current gain, grounded base at very low frequency.

It can be seen from this brief outline that the instrument provides maximum accuracy in the range of maximum interest, namely values of α from 0.9 to 0.9999, and sufficient precision (4 significant figures) in Alpha to reliably detect small but important Alpha variations with changes in bias and device characteristics.

Applications

An instrument with the above measuring capabilities will prove equally useful to engineers engaged in circuit design, device development and production, and active network analysis and synthesis.

Fast, simple and accurate measurements of transistor small signal parameters over the operating range required for a particular application are important to the successful design of a large percentage of transistor circuits. Furthermore, these parameter measurements and their relationship to circuit performance are essential in formulating sound device parameter requirements. When device requirements are formulated on a sound basis, the chances of having equipment give satisfactory performance in mass production with transistors produced on the basis of these requirements are greatly improved. Also, in the case of failure to meet prototype design expectations, the search for the cause of the failure is facilitated.

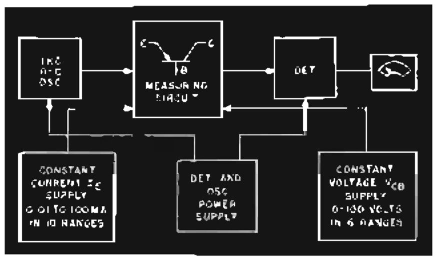
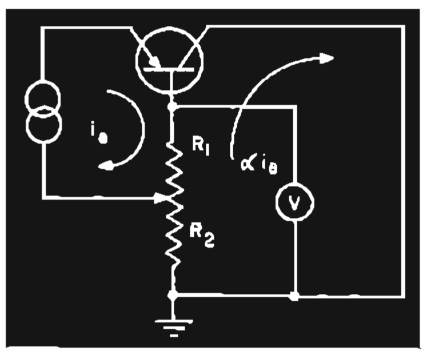


Figure 2. Block Diagram — Transistor Test Set Type 275-A.



When $V = 0$, $i_e R_1 = \alpha\, i_e (R_1 + R_2)$
$$\text{so } \alpha = \frac{R_1}{R_1 + R_2} = \frac{R_1}{R_T}$$
if $R_1 + R_2 = R_T$

Figure 3. AC Equivalent Circuit — Alpha Measurement.

Current Gain and Amplifier Linearity

One of the simplest examples of the use of the BRC Transistor Test Set Type 275-A is in connection with the design of a fully loaded common emitter transistor audio amplifier stage. Since the small signal common emitter current gain, i_{out}/i_{in} is given by

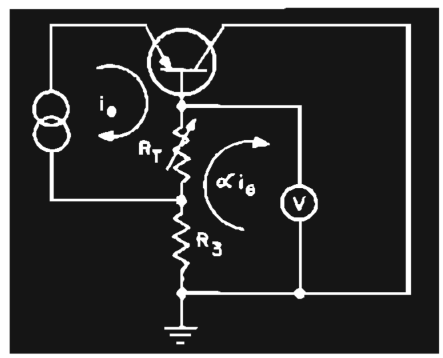
$$\frac{i_{out}}{i_{in}} = \beta = \frac{\alpha}{1-\alpha} \qquad (1)$$

then any change in β over the range of bias covered by the transistor load line will result in nonlinearity of amplification. The magnitude of this nonlinearity will be quantitatively related to the degree of departure of β from constancy. Now if we look at the variation of β as a function of the corresponding variation in α with operating bias, we find that

$$\frac{d\beta}{\beta} = \frac{d\,\frac{i_{out}}{i_{in}}}{\frac{i_{out}}{i_{in}}} = \frac{d\alpha}{\alpha}\left[\frac{1}{1-\alpha}\right]$$

$$\text{or } \frac{d\beta}{\beta} = \beta\,\frac{d\alpha}{\alpha} \qquad (2)$$

Equation (2) shows that the common emitter current gain nonlinearity will be β times the common base current gain nonlinearity.



When $V = 0$, $i_e R_T = \alpha\, i_e (R_T + R_3)$
so $\alpha = R_T / R_T + R_3$
$$\text{but } \beta = \frac{\alpha}{1-\alpha} = \frac{R_T}{R_3}$$

Figure 4. AC Equivalent Circuit — Beta Measurement.

Figure 6 which shows a comparison of β change as compared to α change in a particular transistor over a wide range of operating bias, graphically illustrates the considerably greater change in β than in α in a high α transistor. By precise measurements of α or β with the 275-A, across the desired operating range of bias, limits on α or β variation can be set to meet the required linearity of the application.

Next consider the case in which we wish to improve the common emitter gain linearity, or increase the common emitter gain-bandwidth, by the use of collector-to-base feedback. The necessary reduction in gain to obtain the desired improvement in linearity or increase in band width and the required collector-to-base feedback resistance can be easily determined if the small signal values of α or β are known (Reference 1).

These values can be rapidly determined on the 275-A Test Set.

Current Gain — Switching Transistors

In switching applications it is desirable that transistors maintain a high value of α as close as possible to cut-off; i.e., low emitter current and maximum collector voltage and also, into the saturation region, i.e., low collector voltage and high collector current. The Transistor Test Set, Type 275-A because of its ability to measure at extremely low emitter currents, (0.01 ma.) can give quick answers to α or β variation in the cutoff region. Likewise the saturation region variation can be studied. In the event of serious falloff in α in the saturation region, the source can be traced to either internal generator current gain magnitude fall off at high current densities, or robbing of collector junction voltage by high internal series collector resistance. Thus many of the uncertainties in tracing switching transistor troubles on a large signal basis only can be eliminated.

Input Resistance

One of the most useful functions of the Transistor Test Set Type275-A is to study the effects of the often neglected parameter h_{ib}. The 275-A measures low frequency (1 kc) h_{ib} since this is effective in locating excessive series emitter resistance, which is frequently the cause of trouble in transistors. h_{ib} (the input impedance of the transistor common base with the collector shorted to the base) is given by

$$h_{ib} = r_e' + r_e + (1-\alpha_o) r_b \quad (3)$$

where r_e is the junction diffusion resistance. At room temperatures this is given approximately by $n26/I_E$, where n is usually unity for germanium, but may be as large as 2 for silicon, and I_E is the dc emitter current in milliamperes.

r_e' is any residual series emitter resistance due to contact or spreading resistance.
r_b is the base resistance.
α_o is the low frequency magnitude of alpha.

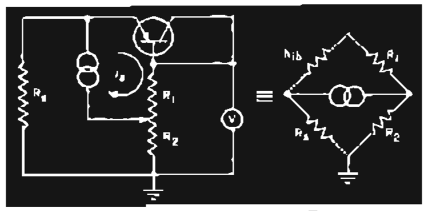
Since the transistor is essentially a power operated device, the power dissipated in r_e' of (3) usually represents a loss in gain.

Now how can the Transistor Test Set 275-A be used to detect this high value of r_e', if it exists? It is reasonable to assume that r_e' and $(1-\alpha_o)$ r_b are constant over the operating range of interest. Then if h_{ib} is measured at two or more values of I_E in the anticipated range of use and these values are linearly plotted as a function of $1/I_E$, a straight line connecting these points will have an intercept on the $1/I_E = 0$ axis equal to $r_e' + (1-\alpha_o)$ r_b. Since $(1-\alpha_o)$ r_b is expected to be very small for high α_o transistors, the major portion of large valued intercepts will in general be due to high r_e'. The fact that all transistors do not have a low value of r_e' is shown in Figure 7, where r_e' has been investigated by the above technique for two different transistors using the 275-A Test Set. Transistor #2 has an intercept of 1.5 ohms which indicates that r_e' is quite low for a transistor of this power level. Transistor #1 on the other hand has an intercept of 20 ohms indicating a high value of r_e'.

In those applications where power gain is not important, this high value of r_e' may be negligible. Let us, however, consider the seriousness of large r_e' in the design of a single mismatched IF amplifier stage using single tuned reactance network coupling (Reference 2). The single tuned interstage coupling network is illustrated in Figure 8. In this circuit, Q is given by $\omega L/R_2$ or $1/\omega CR_2$ at band center frequency. R_2 is the resistance component of h_{ib} (in this example, taken as the measured low frequency value of h_{ib}), and R_1 is the impedance transformed value of R_2 facing the collector of the driving transistor at resonance and given by:

$$R_1 = Q^2R_2 = (\omega L)^2/R_2 = (\omega L)^2/h_{ib}$$

Now suppose that transistor #1 of Figure 7 with an emitter current bias



$$\text{Where } h_{ib} = R_s \times \frac{R_1}{R_2}$$

$$\text{at null, } V = 0$$

Figure 5. AC Equivalent Circuit — h_{ib} Measurement.

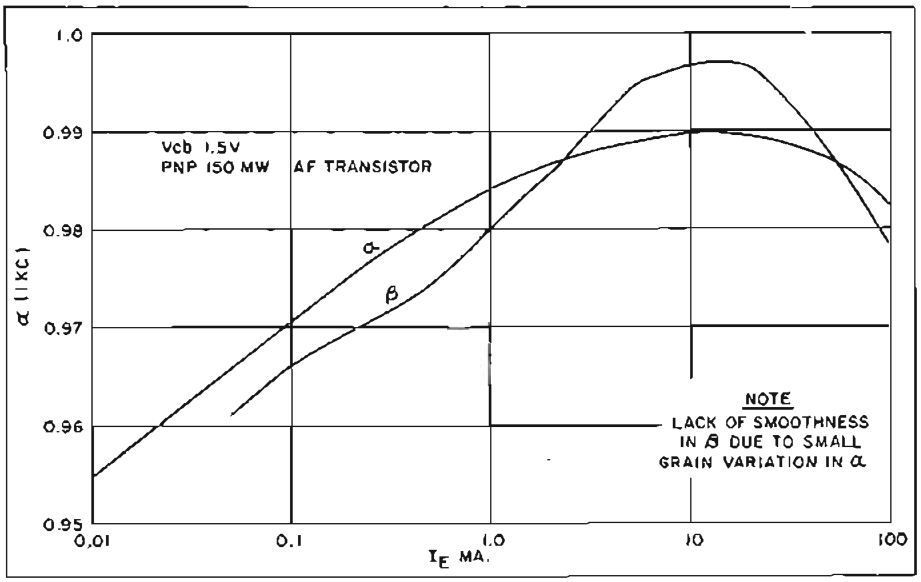


Figure 6. Beta and Delta Versus I_E

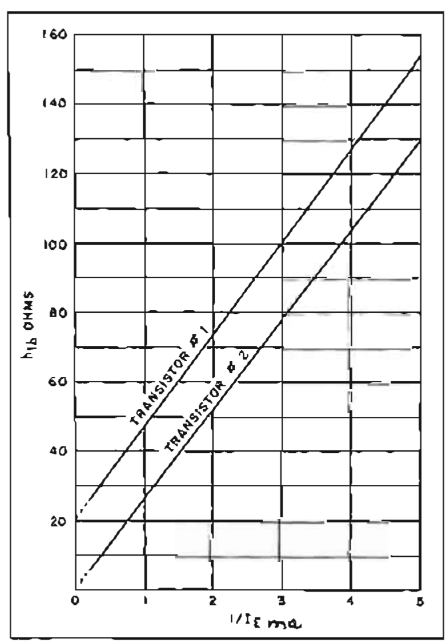


Figure 7. h_{ib} Versus $1/I_E$ For Typical Transistors.

of 2 ma. were used in the common base connection with this circuit. R_2 (h_{ib}) would be 34 ohms whereas if it were not for the high values of r_e' it would be of the order of 14Ω or approximately one-half. Therefore half the power into the transistor is lost by dissipation in a passive resistor. Furthermore, if we had a fixed production circuit and the h_{ib} of the transistor varied from 15Ω to 30Ω we would have a four-to-one variation in power gain and a two-to-one variation in band width.

Nothing has been said about h_{ie}, the common emitter input impedance. This is given by

$$h_{ie} = r_b + \frac{r_e + r_e'}{1 - \alpha} \qquad (6)$$

For high α transistors at frequencies below β cutoff the second term of (6) is usually the larger term. Therefore r_e' is often as important in the common emitter connection as it is in the common base connection.

Many transistor circuits will require at least a reasonably good control of h_{ib}, if not an approach to the minimum value obtainable when r_e' vanishes. It is apparent from the above discussion that some of the frustration and failure which might be experienced in designing transistor circuits can be avoided by a study of the effect of h_{ib} on circuit performance, followed by suitable measurements of h_{ib} on the Transistor Test Set Type 275-A.

Description

There are several circuit features in the 275-A, worthy of note, that have not yet been described.

Expanded Alpha Range

In the specification it was seen that an expanded Alpha range giving four significant figures is provided. This is achieved, simply and stably, as shown in Figure 9. The adjustment R_v is required to correct for unit-to-unit variations in the total resistance of the linear measuring circuit potentiometer, R_T.

Transistor Protection

An important item to the user of the 275-A is the protection provided against accidental burnout of the device under test. First, each current and voltage range is limited to a maximum output value less than twice the full scale value of that range. Secondly, a means is provided to check the polarity of the transistor being tested in a safe manner. This operates as follows: The switch labeled SET-CHECK-MIN (lower left of main

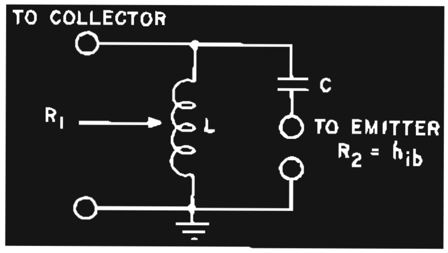


Figure 8. Single Tuned Interstage Coupling Transistor. IF Amplifier.

dial in Figure 1) is moved to SET, disconnecting the transistor socket, shorting the I_E supply, and open circuiting the V_{CB} supply. The desired I_E and V_{CB} are set by manipulation of the proper range switches and coarse and fine controls. The SET-CHECK-MIN switch is moved to CHECK and the METER switch is set at I_E. If the emitter current does not drop below ½ of the value set, the device is correctly poled. If polarity is not correct, the current will drop nearly to zero. If polarity is correct, moving both switches to MIN connects the transistor and permits a minimum to be achieved by rotating the main dial.

Oscillator

The oscillator is a simple RC Wien Bridge type with a fixed frequency of 1 kc. The output, which is coupled to the measuring circuit through an audio transformer of 20:1 ratio, has an internal impedance of 5 ohms and a voltage output of 0.035 volts. Selection of reliable components and use of sufficient feedback make this a stable, trouble-free signal source. The source impedance is suitably varied by series resistors selected by the α, h_{ib}, β, range selector switch, to keep the peak ac-to-dc ratio sufficiently less than unity, to avoid troublesome harmonic generation. The

$$R_T' = \frac{R_V R_T}{R_V + R_T}$$

$$R_f = 9 R_T'$$

Figure 9. AC Equivalent Circuit-Expanded Alpha Scale.

harmonic generation obscures the minimum at balance and excess even order harmonics will cause a bias shift.

Detector

This circuit consists of two amplifiers, a cathode follower, and averaging rectifier. The resulting dc output is applied to the meter and a dc voltage is fed back to the grids of each stage as in automatic gain control, to extend the dynamic range of the indicator and permit a highly sensitive indication of the minimum, without the need for auxiliary level controls. Feed back of dc also prevents any possible damage to the meter by the large signal at the input of the detector when the potentiometer arm is far off the balance point.

Power Supplies

There are three power supplies: two are used for the transistor emitter and collector biasing, and the third is used to supply the internal oscillator and the detector.

The emitter power supply has a very high output impedance and the following ranges: 0-0.1, 0.2, 0.5, 1, 2, 5, 10, 20, 50, and 100 ma., covering most of the low and medium power transistors. The control that varies the emitter current over the entire range has both coarse and fine sections, permitting the user to set the desired emitter bias very precisely. Because of the high internal resistance, the bias magnitude, which is set by shorting the emitter and base terminal for transistor protection, is not affected when the transistor is placed in the circuit. Also, inserting different transistors has no effect on the I_E selected. Since the ac output impedance is many times the dc resistance, the shunting effect on the measuring circuit is negligible. Figure 10 gives a better understanding of the I_E power supply. Current through the tube is adjusted from 0.01 ma. to 100 ma. by the variable cathode resistor, R_C. The emitter current ranges below 5.0 ma. are obtained by a current divider in the plate circuit, since the resulting power dissipation is very low. The breakdown diode (Z in Figure 8) protects the emitter-base junction of the transistor by limiting the high open-circuit voltage that develops because of the large impedance of the I_E supply.

The collector power supply is an electronically regulated constant voltage source with ranges of 0-2, 5, 10, 20, 50, and 100 volts. Like the emitter power supply this also allows easy, accurate selection of the desired V_{CB}. A current regulating tube improves the stability of the reference tube voltage vs. power line fluctuations. This is especially important at low V_{CB} values. The internal ac and dc impedances are so low that neither switching in the transistor nor I_c variation has significant effect on the value of the voltage set. Of course, the ac output resistance, which is less than 15 ohms, acts as an ac short to the several megohm output resistance of the transistor under test.

In addition to the above mentioned circuit features of the instrument, certain additional features are provided. For example, a socket is provided for connection of external power supplies providing up to 5 amperes of emitter current and 100 volts of collector-to-base biasing for Alpha and Beta measurements. (The base current should never exceed 100 ma.) Also, a jack is provided at the rear of the instrument so that a milliammeter can be inserted to measure the dc base current.

Another jack at the rear of the instrument permits the use of a sensitive narrow-band external detector, such as the Hewlett-Packard 302A Wave Analyzer, for measurements at very low signal level and I_E values below 10

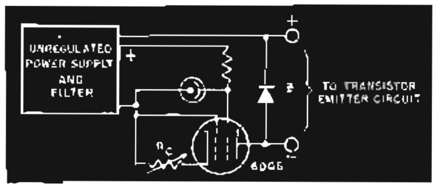


Figure 10. Simplified Circuit — Constant Current I_E Supply.

micro-amperes. The internal power supplies I_E and V_{CB}, which are metered to $\pm 1\frac{1}{2}\%$ of full scale, can be used to energize external circuits when not in use measuring transistors.

A jig, Type 575-A, with different types of transistor sockets, is supplied with the instrument for the customers' convenience, permitting a variety of transistors to be tested with little trouble. A feature of the Type 575-A jig is that the socket mounting plate may be readily removed and duplicated to facilitate mounting of special sockets, heat sinks, or other devices.

Summary

This unique instrument which enables the user to measure, with a new technique, the important dynamic parameters (α, β, h_{ib}) of a transistor is very simple to operate and stable of calibration. It is also evident that dc transistor characteristics such as dc Alpha, dc Beta, I_{CO}, I_{CBO}, I_{EO}, I_{CEO}, I_{CES}, etc., can be measured.

Because of its versatility the limit to the usefulness of the 275-A will be the ingenuity of the user.

Acknowledgment

The authors take this opportunity to express their thanks to Mr. D. E. Thomas of the Bell Telephone Laboratories for his ideas and constructive help while developing the instrument and writing this article. We also wish to thank Mr. J. E. Wachter and co-workers of Boonton Radio Corporation who contributed to the work.

References

(1) Thomas, D. E., "Some Design Considerations for High Frequency Transistor Amplifiers", Bell System Technical Journal, Vol. 38, No. 6, Nov. 1959, pp. 1571-1577.

(2) Thomas, D. E., "Some Design Considerations for High Frequency Transistor Amplifiers", Bell System Technical Journal, Vol. 38, No. 6, Nov. 1959, pp. 1560-61.

New Navigation Aid Test Set

ROBERT POIRIER, *Development Engineer*

Boonton Radio Corporation will soon introduce the Navigation Aid Test Set Type 235-A which provides all of the rf circuitry required for bench testing the ATC (Air Traffic Control) Transponders and airborne DMET (Distance Measuring Equipment — Tacan) sections of the VORTAC navigation system.

At the outset of the development of this instrument, it was proposed that all of the rf circuitry for testing ATC Transponders and airborne DMET be contained in a single package. Although the test procedures of the two devices are different, the Transponder being fundamentally a replying device and the DMET Radio Set an interrogation device, the one package concept is facilitated by the close interlacing of the receiver operating frequencies of the two devices.

The BRC Navigation Aid Test Set Type 235-A contains three basic interconnected units: viz, a crystal-controlled RF Signal Generator, a peak pulse power comparator, and a wavemeter.

Used in conjunction with either the Collins Radio Company's 578X-1 Transponder Bench Test Set or the 578D-1 DMET Bench Test Set and a suitable oscilloscope (Hewlett-Packard Type 150A with 152B Dual Channel Ampli-

Figure 1. Navigation Aid Test Set Type 235-A.

fier) the BRC Navigation Aid Test Set Type 235-A is capable of performing the tests shown in the table in Figure 2.

Crystal-Controlled RF Signal Generator

Referring to the functional block diagram (Figure 3), the receiver test frequencies are generated in three crystal oscillators, heterodyned twice, and doubled in the final modulator stage. The coarse frequency oscillator generates eight 5-mc intervals of frequencies from 25 mc to 65 mc inclusive, and the fine frequency oscillator generates nine 0.5-mc intervals of frequencies from 6.1 mc to 10.6 mc inclusive. After mixing, the sum of the inputs is extracted, providing a frequency spectrum of 31.1 mc to 75.6 mc inclusive in 0.5-mc steps. A tuned amplifier suppresses the spurious products. The bandswitching oscillator and multiplier generate one frequency for each of the two bands; viz., 448.9 mc for the low band and 533.9 mc for the high band. The sum of the outputs of the first mixer and either of the two output frequencies from the bandswitching oscillator is generated in the second mixer, and after filtering and doubling, results in two bands of output frequencies from 960 mc to 1049 mc inclusive and 1130 mc to 1219 mc inclusive in 1-mc steps. Digital readout of the output frequencies, together with a printed dial readout of DMET channel numbers 1 through 63 for the frequencies 962 mc to 1024 mc inclusive and 64 through 126 for the frequencies 1151 mc to 1213 mc inclusive, are provided on the front panel of the instrument.

A total of 180 discrete crystal-controlled output frequencies are generated from 21 crystals, the frequency accuracy of which depends primarily on the two crystals in the bandswitching oscillator. The bandswitching oscillator crystals are specified to ±0.0025%, the coarse frequency oscillator crystals to ±0.005%, and the fine frequency oscillator crystals to ±0.01%. The output frequency accuracy is specified to ±0.005% at room temperature. No crystal ovens are used, since the instrument is intended for use in laboratory ambient conditions.

A similar frequency generator could have been built with 20 crystals, with the frequency accuracy depending primarily on only one crystal, by choosing either the sum or difference products of the second mixer for the high and low bands respectively. The disadvantages of the sum or difference mixing; viz., reduced frequency accuracy of the low

578X-1 Transponder Bench Test Set	578D-1 DMET Bench Test Set
1. Receiver Sensitivity	1. Receiver Sensitivity
2. Receiver Bandwidth	2. Transmitter Power
3. Transmitter Frequency	3. Search Speed
4. Transmitter Power	4. Search Range Limit
5. Side-lobe Suppression	5. Decoder Selectivity
6. Echo Rejection	6. Identification
7. Decoder Tolerance	7. Flag Operation
8. Receiver Dead Time	8. Distance Accuracy
9. Reply Pulse Position	9. Tracking Rate
10. PAR Response	10. Transmitter Pulse Characteristics
11. AOC and Count Down	11. A.G.C. Performance
12. Identification Pulse Delay	12. Distance Indication
13. Transmitter Pulse Characteristics	
14. Image Response	
15. Random Trigger Rate	
16. Transponder Delay Time	
17. Suppressor Output	

Figure 2. Tests Available With The BRC 235-A When Used With A Synchroscope and the Collins 578X-1 or 578D-1 Test Sets.

(difference) band and the reversal in direction of the digital readout between the two bands, justify the use of 21 crystals and frequency switching of the X12 multiplier.

Both grid and cathode modulation are employed in the Navigation Aid Test Set. When the CW output level has been set to the calibration mark of the thermistor bridge meter, the positive pulse modulation signal is applied to the cathode of the final doubler to completely interrupt the CW output for the duration of each pulse. The demodulated interrupted CW may be observed as a relative amplitude either on an integral peak voltmeter or an external synchroscope for a peak pulse calibration of the calibrated CW output level. Subsequently, the doubler stage is biased to cut-off and the positive pulse modulation signal is applied to the grid through a variable modulation level control which is adjusted so that the detected peak pulse amplitude is the same as that for the interrupted CW signal. The peak pulse output level, thus obtained, is referenced to the thermistor bridge calibration of the CW output level. Frequency multiplication is employed in the output stage in order to obtain a high ratio between the peak pulse output level and the CW output, between pulses, in a single stage modulator. The peak pulse to CW ratio is directly related to the purity of the

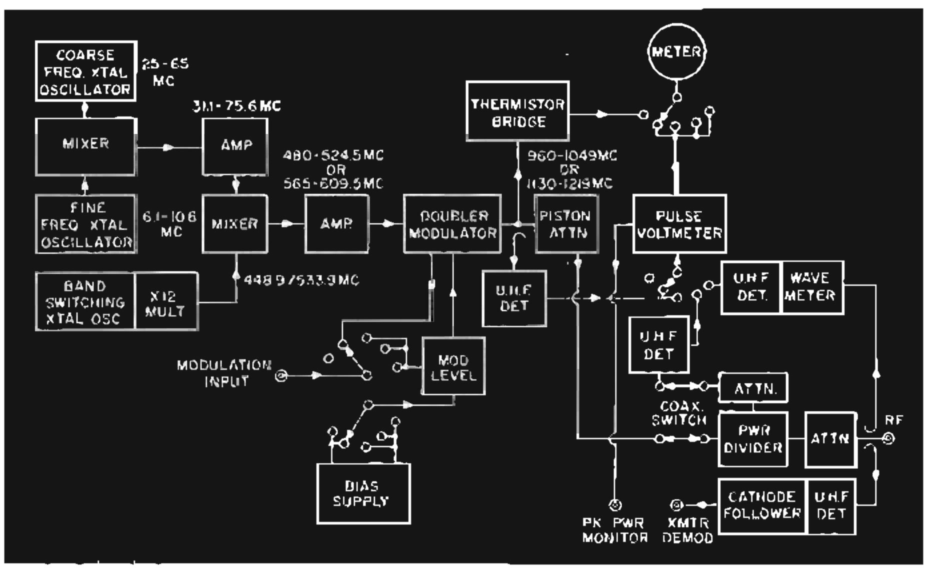


Figure 3 Block Diagram — Navigation Aid Test Set Type 235-A.

driving frequency and, in the 235-A, can be 60 db or better. Spurious outputs are at least 30 db down.

Peak Pulse Power Comparator

Incorporated in the Navigation Aid Test Set is a circuit for measuring the peak pulse output of the transmitters in the ATC Transponders and DMET Radio Sets. In principle, this measurement is made by comparing the relative amplitude of the detected transmitter output pulse after a precisely known suitable amount of fixed attenuation with the variable calibrated output of the RF Signal Generator observed through the same detector. The measurement is made by adjusting the output of the RF Signal Generator, by means of the piston attenuator, for the same detected peak amplitude as was observed for the transmitter output, and reading the piston attenuator dial which is also calibrated in terms of transmitter peak output power over the range of 23 to 33 dbw. The peak power output of the RF Signal Generator is compared with the accurately attenuated transmitter output pulse in the same detector, at the same level but not normally at the same frequency. There is a difference of 60 mc between the transmitter and receiver frequencies of the ATC Transponder, and a difference of 63 mc between these frequencies in the DMET Radio Set. The RF Signal Generator portion of the 235-A is designed for receiver frequencies and does not produce the Transponder transmitter frequency. Some of the DMET transmitter frequencies are incidentally available. The peak power measuring detector is sufficiently flat over 63-mc increments of frequency for an overall accuracy of ±1.5 db for the peak power measurement.

Wavemeter

The wavemeter incorporated in the Navigation Aid Test Set is for the purpose of measuring the transmitter output frequency of the ATC Transponder which obtains from a free-running tunable cavity oscillator. The frequency range of the wavemeter is 1070 mc to 1110 mc in order to accommodate the Transponder operating frequency which consists of a pulse spectrum centered on 1090 mc. Frequency accuracy of the wavemeter is ±0.5 mc over the range of 1070 mc to 1110 mc.

The VSWR of the rf connection on the front panel of the 235-A will be 1.2 or less, and will not be observably dependent on the tuning of the wavemeter. The path between the rf panel connector and the wavemeter includes about 50 db of attenuation. Referring again to the block diagram, the wavemeter indicator is the integral peak voltmeter which provides that the wavemeter indication is dependent only on the transmitter power and spectrum, and is negligibly affected by the reply repetition rate and the number of pulses in the reply code.

Conclusion

The Navigation Aid Test Set provides complete facilities for the accurate testing and calibration of airborne DMET and ATC Transponder Systems and has been specifically developed to serve the needs of BRC customers engaged in the design and operation of this type equipment.

SPECIFICATIONS(1)

Signal Generator Section

OUTPUT FREQUENCY RANGE: Channels spaced 1 mc in the range 960 mc to 1049 mc inclusive and 1130 mc to 1219 mc inclusive.
OUTPUT FREQUENCY ACCURACY: ±0.005% from 60°F to 100°F.
RF OUTPUT LEVEL: —10 to —100 dbm.
RF OUTPUT ACCURACY: ±1 db.
LEAKAGE: Less than —112 dbm.
MODULATION CAPABILITY: Designed for pulse modulation.
Input impedance: 150 ohms.
Input level: +7.5 v peak or more.
Time response: Risetime = 0.14 μsec or less; overshoot 5% or less.
RF Output Impedance: 50 ohms with VSWR of 1.2 or less.

External Power Measurement
FREQUENCY RANGE: 960 to 1215 mc.
INPUT IMPEDANCE: 50 ohms VSWR 1.2 or less.
POWER LEVEL RANGE: 23 to 33 dbw.
200 to 2000 watts
ACCURACY: ±1.5 db.

External Frequency Measurement (ATC Transponder only)
FREQUENCY RANGE: 1070 to 1110 mc.
ACCURACY: ±0.5 mc.

RF Envelope Pulse Detector
FREQUENCY RANGE: 960 to 1215 mc.
SENSITIVITY: External signals; will produce +0.5 volts peak open circuit from a 150-ohm source with 30 dbw input.
Internal signals; will produce +0.5 volts peak open circuit from a 150-ohm source with a level of —13 dbm from the internal generator.
BANDWIDTH: Flat within 3 db to 7 mc.
TIME RESPONSE: Response to Heaviside unit step 0.057 μsec. Risetime, less than 2% overshoot.

Power Requirements: 115/230 volts ac ±10%, 50 to 420 cps.

(1) Based on: Report of Special Committee 58 of the RTCA, "Minimum Performance Standards Airborne Tacan Distance Measuring Equipment Operating Within the Radio Frequency Band 960-1215 Megacycles", dated December 18, 1958.
AEEC Letter No. 57-3-27, entitled, "Status Report of ATC Radar Beacon System Project and Draft Revision of Characteristic No. 532A", dated October 21, 1957.

A NEW UHF Q METER

From the beginning of the Company's history, Boonton Radio Corporation has been active in the development and manufacture of impedance measuring instruments. The first instrument produced by the Company was the Q Meter in the frequency range of 50 Kc to 75 Mc. Later development resulted in a Q Meter making measurements up into the 200-Mc range. Other developments have produced the RX Meter, an RF Bridge operating between the frequencies of 500 Kc and 200 Mc, the QX Checker and subsequently the Q Comparator which offer methods of measuring the deviation of a test sample in Q, L, or C from a standard.

The Q Meters developed in the past have made use of the resonance rise method of measurement. This well known method inserts a known voltage across a very low resistance in series with a resonance circuit. A vacuum tube voltmeter measures the voltage across the capacitor in the resonant circuit while the inductance in this circuit is ordinarily the device of which the Q is to be measured. If the insertion resistor is very low in value and the capacitor has very high Q, the vacuum tube voltmeter across the capacitor can be made to read directly in Q of the test coil. This method has the advantage of being very simple, capable of good accuracy, and entirely direct reading on a precision meter. Beside Q, the Q Meter can also be used to measure capacitance, inductance and resistance over the frequency range of the instrument.

The need for a Q Meter at frequencies above 200 Mc has long existed, but the difficulty in designing a very low and constant value pure resistance insertion impedance and the problems associated with designing a high Q, low inductance variable capacitance over this range have made such an instrument impracticable. It has been known for several years that Q could also be deduced by varying the frequency applied to a resonance circuit so as to obtain a voltage across the capacitance 3 db down from the resonant peak on each side of that resonance. The frequency bandwidth between the 3 db points divided by the resonant frequency yields the Q. Such a method in the low frequency Q Meter, however, does not yield a simple direct reading instrument.

The Boonton Radio Corporation now has a Q Meter, in advanced stages of preparation for manufacture, covering the frequency range of 200 to 600 Mc. making use of the frequency variation method of Q measurement. This instrument consists of a specially designed oscillator, Q capacitor and sensitive detector. The Q capacitor is designed so that inductive effects are automatically removed from the measurement thus yielding a device which exhibits pure capacitance over the frequency range.

The oscillator frequency varies logarithmically with the dial shaft position and the Q of the test sample is indicated directly on a calibrated dial. A dial which reads directly in inductance is also supplied. When using the internal Q capacitor, the instrument will read Q, L, C, and R over a frequency range of 200 to 600 Mc. The Q range is 20 to 2,000. The range of inductance measurements is 3 to 200 milli-microhenries and the capacitance range is 4 to 25 micromicrofarads. A special coupling means is provided so that the resonant frequency and the Q of external cavities located at a distance from the instrument may be measured. The Q range in this application extends upward to 25,000.

A model of the new UHF Q Meter Type 280-A will be exhibited in Booth No. 3101-3102 at the IRE Convention in March of this year.

BOONTON RADIO PLANS EXPANSION

Plans are well under way for the expansion of Boonton Radio's product line and of its facilities. Our Engineering Manager, John P. Van Duyne has been undertaking a broad program of expansion of our engineering personnel to increase the number of precision electronic instruments introduced in the field.

A new site, comprising 70 acres, has been purchased in Rockaway Township four miles west of the Company's present Plant. This site is approximately 36 miles west of New York City and is reached directly by Route 46 from the George Washington Bridge to the newly completed Route 80 which passes within one mile of the new site.

A new building is planned for completion in 1960 which will contain approximately 60,000 square feet. This building will be two and one-half times larger than the Company's present quarters.

The new building will be equipped with modern machinery and new methods which will lead to better control and more precise manufacturing procedures for our present instruments. New techniques and new operations are being included to take care of specialized and more precise manufacturing methods needed for new instruments which are being introduced at the IRE in New York in March as well as new instruments which are planned for the future.

The initial building is part of a master plan which is being completed for the new site. This master plan includes a much larger program of building and expansion which it is expected will be carried out during the next five years.

This planned expansion of our staff and our facilities will place the Company in a position to produce a larger line of precision electronic instruments and to manufacture equipment for which techniques have not been available in our currently used facilities.

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NUMBER 25

The NOTEBOOK

BOONTON RADIO CORPORATION BOONTON, NEW JERSEY

Applications of The Q Comparator

CHARLES W. QUINN, *Sales Engineer*

The Q Comparator Type 265-A has been previously discussed from the standpoint of detailed description and developmental considerations in Issue Number 17 of the Notebook.[1] This article covers the many and varied production applications of the instrument. Included are the obvious applications as well as some which are specialized and tailored to specific measurement problems.

Brief Description

In a few words, the Q Comparator is a sweep-frequency Q Meter that indicates Q, L, and C in relative terms. Mechanically it comprises two units; the Oscillator-Detector Unit and the Indicator Unit (Figure 1).

When using the Q Comparator it must be remembered that the presentation is the result of sweep frequency injection. The indication may occasionally differ from a single frequency measurement. However, with the use of nominal and limit standards, this difference in indication, if any, can be reconciled. A nominal standard is defined as a component which has been selected as being the mean value to which all other units are compared. For example, if ±10% is the specification tolerance, a nominal standard with an absolute value known to ±1% would be used. In this instance, one would work to a specification tolerance of ±9%.

Improved accuracy and better correlation can be provided by means of limit standards. These standards are used in the same way as nominal standards but are made up to provide greater accuracies within the tolerance limits than are attainable with the nominal standards.

YOU WILL FIND . . .

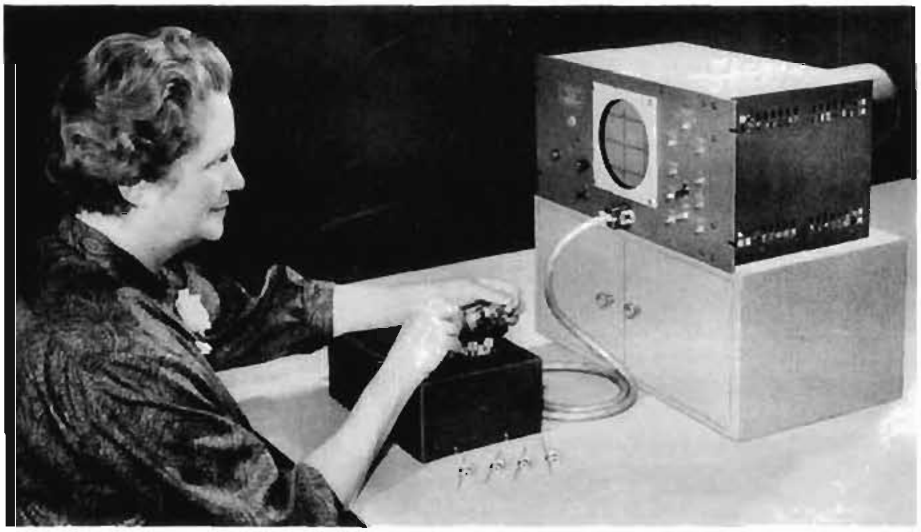

Figure 1. Operator checks components on the Q Comparator.

Purpose and Design Considerations

The purpose of the Q Comparator is the solution of quantity electrical measurements on a comparative basis with speed, ease, accuracy, and reliability. The instrument has been mechanically and electrically designed with the user in mind. First, the cost has been kept to a minimum commensurate with reliability, accuracy, readability, and minimum operator technical ability. The instrument requires a minimum amount of set-up time (about the time required to make one measurement on a Q Meter, which is a laboratory instrument for absolute measurement). Optimum simplicity, elimination of special tubes, speed of readout, and quantity testing of components and circuits were also important considerations.

Electronically and mechanically the Q Comparator has been designed to serve the following industry functions:

1. Incoming inspection.
2. Process inspection and control.
3. Quality control.

Principal of Operation

The heart of the Q Comparator is the Detector Unit or RF Unit which is shown in block diagram form in Figure 2. Once the configuration in Figure 2 is thoroughly understood, applications of the instrument will become apparent to the reader or user who has an unsolved or singular problem in the electrical measurement field. (BRC and the writer are very much interested in applications of this type and hope that the reader will not hesitate to discuss them with us.)

The motor-driven capacitor shown on the left-hand side of Figure 2 sweeps the center frequency of the oscillator. Output from the oscillator is maintained constant over the sweep-frequency range by the dynamic limiter. From this point, the RF signal is shunt fed through a small differential capacitor to the HI—L-C terminals. A so-called "infinite impedance" detector and balancing stage form a differential amplifier to drive the indicator unit. The horizontal trace is generated simultaneously by the sweep capacitor.

Set-Up Procedure

The detailed set-up procedure is given in Notebook No. 17. The production units are as described therein with one exception: the L-C ranges are ±20%

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or ±5%.

Briefly, the nominal component is mounted in a suitable jig and connected to the terminals on the Oscillator-Detector Unit. With the intensity at maximum, the SELECTOR switch is set to CENTER Q and the CENTER Q control is adjusted to center the reference line on the face of the scope. At the same time, the TRACE WIDTH control is adjusted to bring the trace within the dotted lines on the scope face. The switch on the TRACE WIDTH control is set fully counterclockwise (±20% position) at this point. Then, with the SELECTOR set to CAL Q, the CAL Q control is adjusted to bring the reference line to the lower limit on the scope face. The SELECTOR is now set to the USE position and we are ready to set up our nominal component.

Our problem now is to find the desired reasonant point for the component involved. Generally time is saved by readjusting the CENTER Q control so that the base line is visible. The Q capacitor and oscillator frequency are then swept manually until the reasonant curve appears approximately near the center of the scope face. (The Q capacitor is adjusted from its maximum capacitance or fully counter-clockwise position.) The Q centering operation is repeated and the SET Q OF STD-COARSE control is adjusted so that the peak of the resonant curve is near the center of the scope. The dot is positioned at the peak of the resonant curve, the intensity is reduced, and the dot is focused by means of the proper controls. The SET Q OF STD-FINE control is then adjusted to center the dot vertically, and the capacitance or frequency is adjusted to center the dot horizontally.

The above procedure sets up the Q Comparator to the ranges of Q ±25% and L-C ±20% or ±5%, the latter depending upon the position of the L-C TRACE WIDTH switch. However, this is by no means the limit of accuracy or resolution. This subject will be taken up in subsequent paragraphs.

Basic Applications

Inductor testing is the primary function of the Q Comparator and is a direct measurement for values between 1 μh and 15 mh by proper selection of frequency. Indirect measurements can be extended to 0.15μh and 50 mh by utilization of series and shunt techniques. (Dot presentation permits Q measurements from 30 to 500 as specified. However, it has been found that in some cases useful curve presentations can be obtained down to Q's of 15.) The expected accuracy of indirect measurements will usually be less than that of a direct method.

Capacitor testing is the second function of the Q Comparator, and is an indirect measurement in that an external reference inductor must be used. Capacitance is virtually direct; that is, the accuracy is essentially as stipulated for inductors (for values of 500 $\mu\mu$f or greater) and is a function of the internal capacitance setting. By using minimum internal capacitance and maximum sweep expansion, approximately one micromicrofarad can be spread across the face of the scope. This requires a choice of the resonating inductor for minimum distributed capacitance of less than 5 $\mu\mu$f.

Resistance measurements are indicated on the Q scale and are indirect measurements. This means that these two resistance measurements are indicated as a change of Q with respect to the reference inductor.[2]

Specific Applications

Knowing that the Q Comparator is capable of relative measurements of the basic combinations of L, C, and R, we begin to think of other components and circuits to which it may be applied as a two-terminal indicating device. Some of these applications are listed below.

Inductance Measurements

1. Iron cores, shells, toroids and rods (powdered iron and feramic type) may be tested. Relative permeability and Q (or dissipation factor) can be determined at a single glance.
2. Inductors may be trimmed while mounted in position on the Oscillator-Detector Unit. The dot presentation greatly simplifies this operation since no judgement is required to determine the point at which a meter peaks, and the direction of the adjustment (plus or minus) is immediately indicated.
3. L-C and R-C networks can be compared, providing their impedance falls in the general range of Q Meter measurements.[3]
4. The self-resonant frequency of a coil can be determined as follows. A work coil is used to obtain the resonant peak at the nominal frequency, then the coil to be tested (coil X) is connected to the capacitor terminals. If the dot deflects vertically downward only, the resonant frequency is the same as the reference coil. If the dot deflects to the left, coil X is inductive and its resonant frequency is higher than the resonant frequency of the reference coil. If the dot deflects to the right, the resonant frequency of coil X is lower than the resonant frequency of the reference coil. The resonant frequency limits can be established by using Q Meter Type 260-A in conjunction with the limit standards.

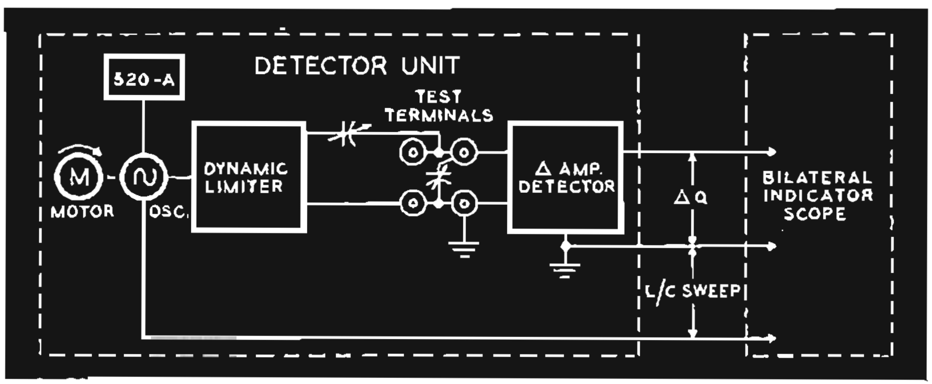

Figure 2. Block Diagram — Q Comparator.

Capacitance Measurements

1. Quartz crystals (blanks or complete assemblies) may be checked for capacitance variations as quickly as they can be inserted in the test fixture. This is performed at frequencies well below resonance.
2. Vacuum tube capacitances; i. e., grid to cathode and plate to grid (in triodes), can be compared quickly.
3. Variable capacitance diodes can be graded simultaneously for Q and capacitance at the rate of approximately

600 to 1000 an hour with simple, manually-operated fixtures.
4. Rapid relative dielectric constant measurements can be made when it is necessary to check for statistical or control purposes.[4]
5. When the dielectric constant of a material is constant, the Q Comparator can be used to indicate relative thickness. A fixture described in Notebook No. 8 could be used for this purpose.[4] The accuracy required will determine how elaborate the fixture must be and what is required to prepare the specimen for thickness and dielectric measurements.
6. Relative moisture measurements of materials can be made. Since the Q Comparator uses an instantaneous two-dimensional presentation, relative loss or resistance is always indicated, even in low-loss materials. If the material is hygroscopic and subjected to a controlled environmental humidity, two indications will be observed. Water, which has a dielectric constant of approximately 80, will theoretically effect the nominal dielectric constant of the material when it is dry. This will be observed as plus C on the indicator for materials with low dielectric constants. However, since the effect of moisture on the loss factor or Q is usually many times the effect on the dielectric constant, a decrease in Q will be much more obvious and would be an indication of the moisture content.

Statistical Studies

After a nominal coil has been set up as previously described, the dot will move on the scope face in accordance with the per cent variations of L and Q of the coils tested. This information is not only useful to an inspection department, but can also be utilized by production engineering groups in specifying manufacturing tolerances. For example, a new component, let's say a small choke coil, is contemplated as an addition to a line of products. This component is first designed in the engineering department. Several preproduction samples are made and checked in the laboratory. The next step is to prepare a pilot production run of a few hundred coils. These can be checked and graded on the Q Comparator in approximately fifteen minutes by a non-technical operator. This statistical information, together with a count for each grade, is then fed back to engineering and tolerance distribution curves, similar to those in Figure 3, are plotted and analyzed. In our example,

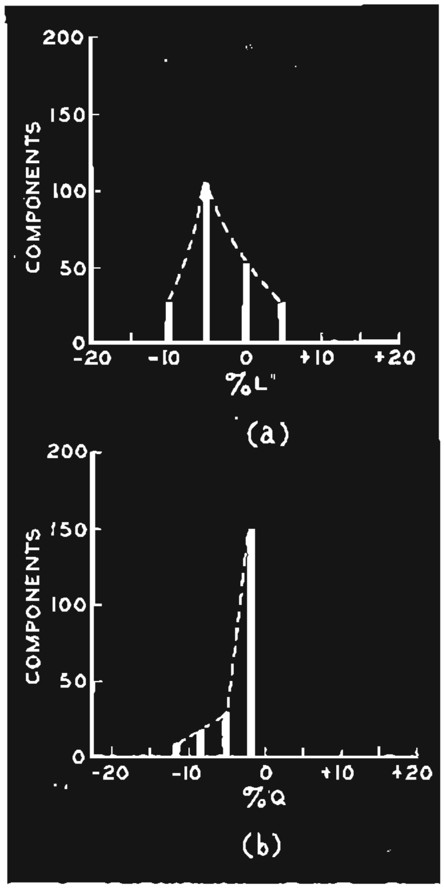

Figure 3. Typical tolerance distribution curves for analyzing pilot and production component runs.

curve (a) in Figure 3 indicates that the inductance is averaging approximately 5% below the expected normal. It is immediately evident that action should be taken to move this distribution curve toward the zero point, or to adjust the specifications. This unexpected picture prompts us to ask, "Why the difference between the prototype and the pilot run?" Possible answers might be differences in coil form diameter, wire spacing, lead mounting, and lead dress. These effects can be checked by controlled production runs, analyzed as above.

Another story is told by the Q Comparator as shown in Figure 3. Here Q behaves virtually the same as the prototype with a few miscellaneous deviations from the proposed nominal value. Greater yield will result if a few simple precautions are taken which are suggested by the curve. Since the Q readings are predominatly below the standard, one would expect, for example, that the difficulty is an erratic tensioning system, defective forms, contaminated coatings, or a hygroscopic problem.

In the above example we have dealt with a case where the components or its specifications may be altered as the result of a statistical study. A similar method can be used to grade components to cover a range of cataloged values when the absolute value is difficult to control. For example, suppose it is desired to grade capacitor diodes over a range of 5 to 30 $\mu\mu f$ in 5 $\mu\mu f$ steps. The circuit in Figure 4 would be used to connect the component to the Q Comparator and to provide the proper bias voltage. A nominal value CN would be set up as follows. Assume C equals 25 $\mu\mu f$. If we set one half C equal to 20% (the maximum range of the Q Comparator) we can compute the set up for our desired display of 5 to 30 $\mu\mu f$. If $\frac{CN}{5} = 12.5\ \mu\mu f$, then $x = 62.5\ \mu\mu f$ nominal capacitance. We now refer to our reactance charts to determine the frequency and inductance required for these conditions. One hundred microhenries at 2 megacycles is one possible combination of frequency and inductance that will be satisfactory. Using this as a starting point, the frequency and internal capacitance can be adjusted to place the limits exactly as desired, and the CRT screen on the Q Comparator can be calibrated in 5 $\mu\mu f$ steps.

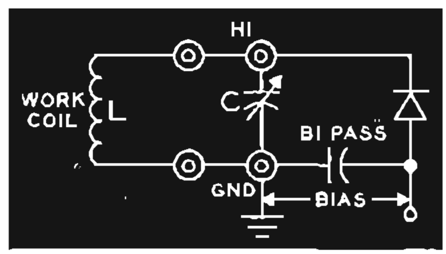

Figure 4. Circuit connections for checking capacitor diodes on the Q Comparator.

A Speed Fixture

Since speed of measurement is a basic advantage of the Q Comparator, we would naturally look for means to utilize its potential. Figure 5 shows a manually-operated fixture designed to minimize handling and loading time, thereby speeding up component testing. This jig, or one constructed along similar lines, permits the checking of 500 to 1000 pieces per hour.

To go further, the Q Comparator can be used as the nucleous of an automated high volume testing system. For example, the component jig could be hopper fed and motor driven, and a photo electric system could be used to reject out of tolerence components. The potential volume with such a system would be limited only by the feed system.

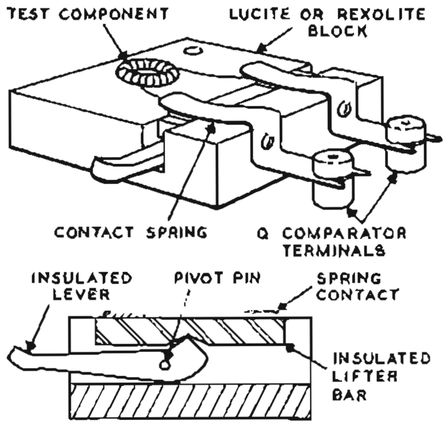

Figure 5. A manually-operated fixture designed to speed up component testing on the Q Comparator.

Capacitors As Calibration Standards

Capacitors are available in virtually limitless values and increments and in tolerence ranges which are adequate for use as standards.

It is obvious that these capacitors can be used to calibrate discrete increments, but not so obvious is the fact that the same scheme can be applied to coil testing. The first step is to measure the resonating capacitance on a Q Meter Type 260-A for the nominal standard. Let us assume the capacitance is 100 $\mu\mu f$. Since L and C bear the same relationship with respect to the resonant frequency, 5% L equals 5% C within the specified limits. If we wish to calibrate the 5% range in 1% increments, 1 $\mu\mu f$ increments will be required. The nominal inductance standard is set up on the Q Comparator with 5 $\mu\mu f$ external capacitance and 95 $\mu\mu f$ internal capacitance. The 5 $\mu\mu f$ capacitance is removed in 1 $\mu\mu f$ steps to calibrate the —L range and an additional capacitance of 5 $\mu\mu f$ is removed in 1 $\mu\mu f$ steps to calibrate the +L range. This operation may be performed extremely rapidly.

Summary

The Q Comparator is a unique instrument specifically designed for high speed production testing of components and networks in the RF range from 200 kc to 70 mc. Offering instantaneous and simultaneous readout of both Q and L-C, operation is extremely simple, and speed of measurement is limited only by the rate at which test components can be fed to the measuring circuit.

References

(1) Wachter, James E., "A Q Comparator", BRC Notebook No. 17, Spring 1958.
(2) Cook, Lawrence O., "A Versatile Instrument — The Q Meter", BRC Notebook No. 4, Winter 1955.
(3) Riemenschneider, Norman L., "The RX Meter or the Q Meter?", BRC Notebook No. 16, Winter 1958.
(4) Riemenschneider, Norman L., "Measurement of Dielectric Materials and Hi Q Capacitors with the Q Meter", BRC Notebook No. 8, Winter 1956.

Typical Performance Data for the Type 225-A Signal Generator

CHANNING S. WILLIAMS, *Production Engineer*

The development of the General Purpose Signal Generator Type 225-A is covered in detail in Notebook No. 21.[1] The catalog specifications define the limit performance capabilities quite completely. However, the collection of data from additional instruments in production more clearly defines the *typical* levels of performance. This article describes the performance levels of seven typical instruments and is written for the guidance of the user who is interested in characteristics not usually specified. It should be noted that, while the data given in this article is typical, it in no way modifies the calalog specifications.

Radio Frequency Characteristics

Frequency stability is defined as the maximum percentage frequency excursion during a stated time interval. For example, the long-term stability specified in the catalog for 1 hour (after a 2-hour warm up) is 0.01% of the carrier frequency or less.

Warm Up Time

A typical instrument achieved the above defined stability in ½ hour at 10 mc, 1 hour at 80 mc, ½ hour at 160 mc, and ½ hour at 320 mc. Warm up time did not exceed 1 hour at any frequency.

[1]Gorss, Charles G., "A General Purpose Precision Signal Generator", BRC Notebook No. 21, Spring 1959.

Long-term Stability

After 2 hours warm up, RF drift for 1 hour typically varies from 0.001% to 0.005% of the operating frequency. The sample stability tape (Figure 1) shows rapid adjustment during the first ½ hour. Drift during succeeding hours is progressively smaller down to the minute changes caused by line variations, FM due to noise, etc. Final stability may take 10 hours to achieve and has been as good as 300 cycles per hour at 100 mc.

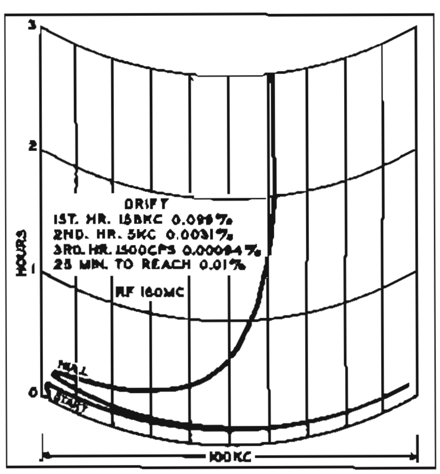

Figure 1. A graphic record of the long-term RF stability of the 225-A.

Short-term Stability

The short-term stability, as specified in the catalog, is 0.001% or less of the carrier frequency for a 5-minute interval after a 2-hour warm up. This stability varies from 0.0001% to 0.00075% of the operating frequency. After 10 hours warm up, the frequency change at 100 mc has been as low as 80 cps for 5-minutes or 0.00008% of the operating frequency.

While measuring short-term stability, sudden perturbations in frequency occurred which displayed steep slopes. These are caused by noise from the oscillator tubes. Note in Figure 2 that the *total excursion* is minute in each case, even though the rate of change may be rapid.

Stability After Operating Controls

During normal operation of a signal generator the frequency controls may be set for one frequency and then changed to another. In a wide range, continuously tunable oscillator, there are many factors which affect the temperature of frequency determining elements. Most important among these factors is the dc power through the tube and the RF current in the inductor. These values are affected by several factors: oscillator efficiency, which is a function of the tank circuit impedance for fixed operating conditions; the amount of feedback, which is also a function of frequency; the operating point of a self-biased oscillator, which is dependant on the foregoing; and the tightness of coupling to the load, which varies from range to range. Any change which affects the level of operation will require some definite time to stabilize, due to the thermal capacity of the parts involved. We may define the point where stabilization occurs as the start of the first 5-minute interval where the change in frequency is less than 0.001% of the operating frequency. Table 1 shows a summary of data for tuning changes of approximately 10% and 20% of the carrier frequency. It should not be assumed

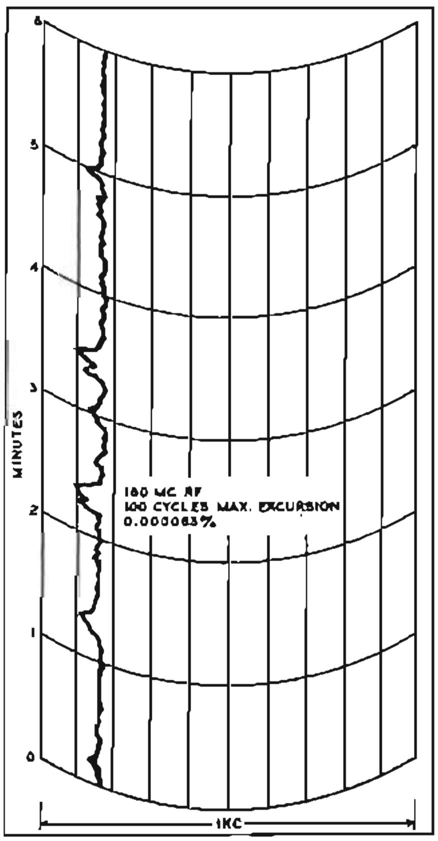

Figure 2. A graphic record of the short-term RF stability of the 225-A.

that a change of 20% will take twice as long to stabilize as a change of 10%. For 31% of the carrier frequencies tested for stabilization, the instrument "stabilized" as defined above, within ½ minute.

Data from a typical instrument indicates that, when operating the Range Switch, an average of 7.7 minutes is required to stabilize to within 0.01% of operating frequency for 1 hour, and 17

TABLE 1

Time Required to Stabilize to 0.001% of the Carrier Frequency for 5 Minutes

Range	Frequency	Maximum Minutes for 10% RF Change	Maximum Minutes for 20% RF Change
1	10mc	9	13
2	35mc	10.5	7
3	68mc	13.8	13.8
4	110mc	4.3	6.2
5	230mc	4.3	4.0
6	270mc	5	2.31
Average of all readings		2.64	6.15

minutes to stabilize to within 0.001% of operating frequency for 5 minutes. The longest stabilization time is required by range three, which requires 15 minutes to stabilize within 0.01% for 1 hour, and 35 minutes to stabilize within 0.001%, for 5 minutes.

For convenience in specifying changes in frequency which occur when varying the RF level control, the 10% and 20% marks on the meter were selected as RF level reference points. For the change in RF level with the frequency dial set near the low frequency end, an average change for all 6 ranges was 0.0016% of the carrier frequency. Near mid range the average for all ranges was 0.0022%, and at the high frequency end, the average for all ranges was 0.007%. A maximum change of 0.018% of the operating frequency occured at 19 mc on range 1. From 10 to 90 mc, an increase in RF level caused an increase in operating frequency. The change occured in less than 5 seconds at all points. Of course, this change is generally unimportant to the user since all calibrations have been made with the RF level set to red line, where the instrument should be used.

If operation of the generator requires frequent adjustment of the attenuator near the 100,000 microvolt level, it is convenient to compensate for the loading effect of the piston on the output tank circuit. Compensation may be achieved by detuning the amplifier trimmer slightly counterclockwise, until variation in the attenuator setting causes no visible change in the RF level meter indication. (Compensation will be accompanied by a slight increase in residual FM.)

The amplifier tank will now have more capacitance than at peak (i. e., will be tuned lower in frequency than the peak). Since the piston loop is in effect a shorted turn which will decrease the inductance of the amplifier tank when coupled closer to it, increased coupling will move the peak of the amplifier resonance curve up in frequency. Therefore, a judicious detuning of the amplifier will allow these two effects to nearly cancel, thus eliminating apparent reaction on the red line reading as the output is varied.

With the attenuator compensated as described in the preceeding paragraph, there was less than 50 cycles frequency change as a result of varying the attenuator from 20,000 microvolts to 100,000 microvolts over the full range of the instrument. When the attenuator was uncompensated the change in frequency due to loading was of similar magnitude.

The detenting resettability, or change of operating frequency, when moving the Range Selector to a high cam position between ranges momentarily and resetting to the same range, is less than 0.01% of the operating frequency when detenting from either side and about 0.006% when detenting from one side. Variation in frequency caused by "rocking" the Range Selector within the detent was less than 0.006% for all frequencies.

Ambient Temperature

An ambient temperature change for a typical instrument caused a 0.03% frequency change per degree centigrade at 320 mc and a 0.006% frequency change per degree centigrade at 20 mc. Frequency stability for a 5-volt line change was 25 cycles at 10 mc and 500 cycles at 320 mc. All frequency changes occurred in less than ½ minute.

Amplitude Modulation Characteristics

One of the characteristics measured during the data taking process was incidental FM due to 30% AM. The method described below was used to measure incidental FM, residual FM, and desired FM. The test set up, shown in Figure 3, is actually a wide band receiver employing a descriminator with very good AM rejection. Limiting is achieved in the H-P 500B frequency meter. The input to this frequency meter must be sufficient to saturate it. The other instruments used are typical of many that will do the job. The audio amplifier used was a 2% instrument covering 10 to

150,000 cps. The local oscillator is another Type 225-A Signal Generator. The local oscillator is set at 100,000 microvolts and the output of the generator under test is set at 20,000 to 50,000 microvolts. This difference is necessary to insure linear operation of the mixer. The setup used permits the reading of deviations of less than 100 cycles. Vertical deflection represents deviation.

To measure the incidental FM, which we define as frequency deviation due to amplitude modulation, the generator under test is modulated 30% with 1000 cps and the scope pattern is measured from peak to peak (less the width of the trace). With no modulation, the indication then represents residual FM from all sources other than AM. A typical 225-A Signal Generator exhibits 400 to 800 cycles incidental FM at 18 mc and even less at other frequencies. For example, 200 to 350 cycles incidental FM is exhibited at 160 mc.

The incidental FM due to 30% AM, as measured on a typical instrument for audio frequencies of 400, 1000, 4000, and 10,000 cps at carrier frequencies up to 20 mc, showed no dependance on modulating frequency.

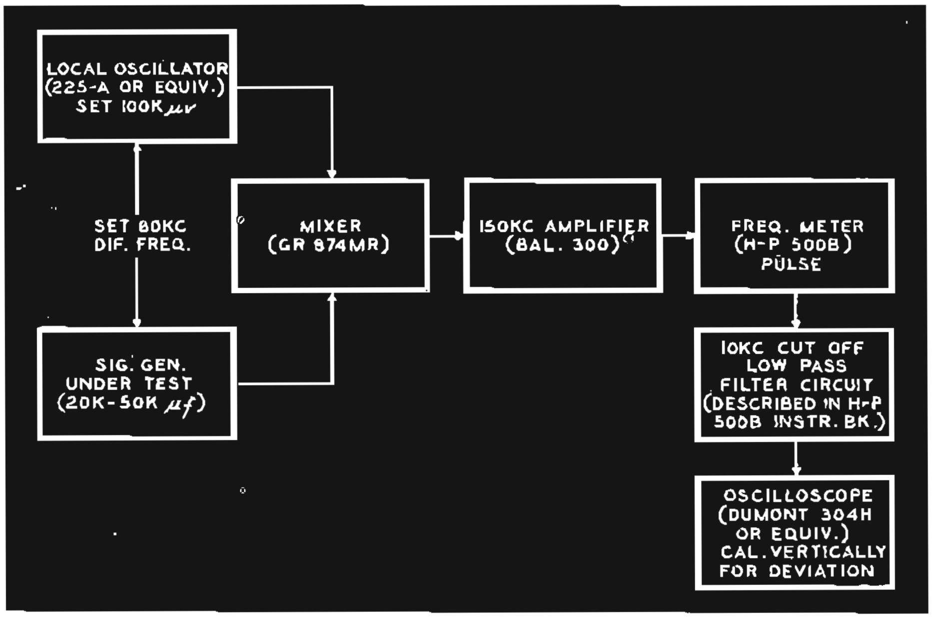

Figure 3. Test set up for measuring incidental FM on the 225-A.

TABLE 2

Range	RF	Deviation Cycles	1000 cps Modulation RMS Voltage	Limiting Factor		
1	10.7	1,740	25	10% Dist.		
2	21	2,170	20	10% Dist.		
2	30	7,550	26	10% Dist.		
3	40	4,360	24	10% Dist.		
4	70	11,000	24	10% AM		
4	92	22,300	26	10% Dist.		
4	100	25,000	24	10% Dist.		
					For 5,000 ~ Deviation	
					RMS Voltage	Resistor*
5	148	7,750	20	10% Dist.	13.7 Volts	
5	155	6,900	18	10% Dist.	12.8	
5	165	10,300	21	10% Dist.	10.6	
5	170	7,050	16	10% Dist.	12.6	
5	174	9,650	19	10% Dist.	10.2	
5	185	22,100	23	10% Dist.	5.3	4,700 Ohms**
5	205	22,100	20	10% Dist.	4.7	5,600
5	216	15,050	17	10% Dist.	5.7	3,900
5	220	25,800	21.5	10% Dist.	4.1	6,800
					For 75,000 ~ Deviation	
					RMS Voltage	Resistor*
6	450	181,000	24	10% Dist.	10 Volts	1,000 Ohms**
6	455	263,000	25	10% Dist.	7.1	2,200
6	460	188,000	16.5	10% AM	6.6	2,700
6	465	211,000	16.5	10% Dist.	5.9	3,300
6	470	460,000	23.5	10% Dist.	3.8	8,200

*Resistor connecting internal oscillator to FM audio input.
**Recovered audio distortion less than 4%.

Frequency Modulation Characteristics

Although the instrument was basically designed for amplitude modulation, provision has been made for frequency modulation from an external source. The resulting FM is useful over the 160 to 500 mc portion of the range.

The audio response of the FM channel is down 3 db at 400 and 12,000 cps. It is also possible to obtain narrow deviation FM from the internal modulation oscillator by connecting a resistor from the AM external modulation binding post to the FM binding post. Use of a resistor as low as 1000 ohms for this purpose will not significantly increase the distortion on the modulating signal. See Table 2 for typical deviations obtainable with different resistor values.

Modulating Oscillator

Output from the internal audio oscillator is available at the AM binding post when the AM Selector is in either the 400 or 1000 cps position. This output is approximately 12 volts RMS and typically has 0.6% distortion with no external load. The distortion of the modulating oscillator in a typical instrument, when grounded externally through 3300 Ω, is 0.9%.

Pulse On-Off Ratio

The generator output may be pulsed by applying a pulse to the AM modulation terminals with the AM level control in Pulse position. Typical DC pulse on-off ratio for −10 volts bias is given below.

10 mc	40 db
15 mc	36 db
150 mc	25 db
700 mc	24 db
450 mc	22 db

Conclusion

The Type 225-A Signal Generator is a truly general purpose generator, providing exceptionally low incidental FM and excellent frequency stability. It is hoped that this additional information regarding the performance of the instrument will prove valuable to the user and increase the utility of his 225-A Signal Generator.

MEET OUR REPRESENTATIVES
EDWARD A. OSSMANN AND ASSOC., INC.

Edward A. Ossmann and Assoc., Inc. has sold and serviced BRC instruments in the Upstate New York area, continuously since 1947. The Company maintains its headquarters in Rochester, with branch offices in Syracuse and Binghamton.

Edward Ossmann, founder of the Company, obtained his EE degree from Manhattan College in 1943. After serving successively as Test Engineer for General Electric Co. and Engineering Manager with the DuMont Laboratories, Mr. Ossmann entered the field of Sales representation in 1946. This was the beginning of a career which was to see his organization progress from a one-man effort to the 23-employee organization it is today.

After the untimely death of Mr. Ossmann in 1959, Mr. Roy Smart, who had joined the organization early that year, became a Director of the Company, and serves now as Vice-President and General Manager. Mr. Smart, a native of England, moved to Toronto, Canada in 1947. He held the post of Works Manager for the Instrument Division of Ferranti Electric, Ltd. from that year until 1954, when he became Manager of the Canadian Division of Helipot Corp. He held the latter position until 1955 when he joined the Ossmann organization.

Sales Manager of the Company is Mr. John Jordan who joined the organization in 1958, bringing with him years of experience as Electronic Engineer with Bell Aircraft and Area Sales Manager with Motorola.

The Company established their headquarters in a new building in Rochester in 1955. Over 6000 square feet in area, the building comprises complete office, service, and warehouse facilities. The main functions of accounting and clerical services are carried out at this location, although both the Syracuse and Binghamton branches are equipped to process and expedite customer orders. All locations have TWX and Western Union service and are in constant communication with the Company's principal factories.

All of the instruments sold by the Company are serviced by Brighton Electronic Laboratories, a Division of Edward A. Ossmann and Assoc. This group is completely equipped to provide calibration and repair service on all BRC instruments.

Edward A. Ossmann & Assoc. headquarters in Rochester, N. Y.

Edward A. Ossmann and Assoc. has endeavored over the years to sell and service only the finest precision electronic instrumentation. It is their firm belief that initial sale of an instrument represents only a small part of their obligation to their customers. To assist in the selection of proper instrumentation for each individual application, and to provide quick and reliable repair service to insure that the instruments they sell continue to fulfill the customer's needs is, they believe, their primary objective.

BRC is proud of its association with Edward A. Ossmann and Assoc. and is grateful for the record of dependable service this Company has rendered to our many customers in the Upstate New York area.

SERVICE NOTE
Checking RX Meter Calibration

The following techniques are given as an aid to those persons responsible for the maintenance and calibration of the RX Meter Type 250-A. It is not intended that the methods described be used to establish absolute calibration of the instrument, but rather, to provide an approximate or relative check as well as an indication of a change in calibration. In many cases, the techniques described will obviate the need for returning to the factory instruments which are thought to be performing improperly.

R_p DIAL

The Type 515-A Coaxial Adapter Kit, with its 50-ohm termination resistor, will check the R_p dial over the entire frequency range at the 50-ohm point. For checking other points on the R_p dial, stable film resistors with short and controlled lead shape and length may be connected to the RX Meter terminals and used to prepare frequency curves of R_p. This should be done after the instrument is received from the factory, or at a time when the calibration is known to be accurate. The film resistors, appropriately labeled, together with the curve data, could then serve as reference standards for the activity responsible for insuring proper operation of the instrument.

C_p DIAL

High quality capacitors with short and controlled lead shape and length may be connected to the RX Meter terminals and used to check the calibration of the C_p dial. When the instrument is known to be accurately calibrated, the capacitors are used to prepare frequency curves of C_p. The labeled capacitors, together with the curve data, are then used as reference standards for subsequent calibration checks of the C_p dial.

A precision variable capacitor may be similarly used to check the C_p dial calibration as follows.

1. Select a coil that will resonate with the precision capacitor at 120 $\mu\mu f$ with the RX Meter C_p dial set to $+20$ $\mu\mu f$ at a frequency in the lowest band, (500-1000 kc).
2. Connect the coil and the precision capacitor to the RX Meter terminals, using the shortest leads possible.
3. Set the RX Meter at zero C_p, the precision capacitor to 115 $\mu\mu f$, and adjust the frequency until a null is obtained.
4. Decrease the capacitance of the precision capacitor in the desired steps (e. g., 10 $\mu\mu f$) and readjust the RX Meter C_p dial for null.
5. Record the C_p dial readings from which a calibration curve can be prepared.

W. J. CERNEY JOINS BRC AS SALES ENGINEER

Many of our customers in the Metropolitan Philadelphia and Washington, D. C. area have already met Willard J. "Will" Cerney, recent addition to the

W. J. CERNEY

BRC Sales Engineering staff, during his visits to those areas. "Will" came to BRC from Link Aviation, Inc. where he worked with instrument trainers, simulators, and associated testing systems, and participated in that Company's training program. Before that time, he was employed by Harnishfager Corp. of Milwaukee, Wisconsin where he assisted in the setting up of a new production control system.

"Will" attended the Milwaukee School of Engineering, the University of Minnesota, and Broome Tech. in Binghamton, New York. While with the U. S. Army from 1948 to 1952, he gained experience repairing radar, navigational, and communications equipment.

During the short time he has been with BRC, "Will" has been instrumental in solving many customer problems and would welcome the opportunity to be of further service to our many customers in this area.

EDITOR'S NOTE
Q Meter Contest Winner

Again this year, the problem coil displayed at the BRC during the IRE show drew a host of hopefuls armed with slide rules, pad and pencil, and crystal balls. Viewing the "monster coil" from every conceivable angle, they slowly lapsed into a stupor, seemingly oblivious to all the commotion around them. Moments later, once again among the living, our friends began their frenzied manipulation of slide rules and delved into page after page of complicated mathematical computations. With the last stroke of the pencil their faces broke into a smile rivaling that of the cat who joined in the search for the missing canary and they quickly jotted down their estimate on the contest card. They were last seen as they disappeared into the stampeding crowd.

The coil has been measured and the story can be told. The Q of the coil, measured at 500 mc on a developmental model of the UHF Q Meter Type 280, is 395. The inductance of the coil is 9.3 mμh.

Winner of the contest and the Type 160-A Q Meter is William F. Byers of General Radio Co. in West Concord, Mass. Other contestants whose estimates are certainly worthy of note are listed below.

Estimate	
386.5	J. H. Marchese, Data Control Systems, Inc., Danbury, Conn.
386.5	E. H. Scannell, Jr., Ft. Trumbull, New London, Conn.
392	F. Hafeed, North Electric, Galion, Ohio
392	J. F. Pryer, Okonite Co., Passaic, N.J.
398	D. T. Walker National Lead Co., South Amboy, N. J.
400	J. Bullinga, National Coil Co., Sheridan, Wyo.
400	W. D. Street, Delta Coil, Inc., Paterson, N. J.
400	Vincent Vinci, Vitro Labs, W. Orange, N. J.
403	Alan Sobel, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.
405	George Kelk, George Kelk Ltd., Willowdale, Ontario, Canada
405	Harry M. Blombaum, Radio Corp. of America, Camden, N. J.

Our congratulations to Mr. Byers and sincere thanks to our many friends who visited us at the show.

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Transistor Current Gain Determination With The Transistor Test Set And The RX Meter

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This article presents a discussion of a method for determining transistor parameters β, β_0, h_{ie}, h_{ib}, f_β, f_T, and Kf_α, using the Transistor Test Set Type 275-A and the RX Meter Type 250-A, together with a step-by-step procedure and an example employing this method. Equations are also included for determining parameters t_r, t_f, and Gm_t. Terms used throughout the discussion are defined below.

Definition of Terms

β — (h_{fe}) Small-signal, short-circuit current gain, common emitter configuration. β will be used in preference to h_{fe} in order to simplify subscripts.

$|\beta_0|$ — Same as β above except that the frequency involved is well below cutoff for transistors with negligible phase shift at 1 kilocycle in the common base configuration ($f_\alpha \geqq 500$ kc).

h_{ie} — Small-signal ac input impedance, common emitter configuration, output short circuited (h_{11e}).

h_{ib} — Small-signal ac input impedance, common base configuration, output short circuited (h_{11b}).

f_β — β cut-off frequency. The frequency at which $|\beta|$ is —3db down from $|\beta_0|$, ($.707|\beta_0|$).

f_T — The frequency at which $|\beta|$ equals unity or zero db. This is also the transistor gain bandwidth product.[1, 8]

K — K is the base grading factor.[1]

α — (h_{fb}) Small-signal, short-circuit current gain, common base configuration.

f_α — f_α is the α (h_{fb}) cut-off frequency where $|\alpha|$ is —3db down from $|\alpha_0|$.

α_0 — Same as α above except that the frequency involved is well below cutoff for transistors with negligible phase shift at 1 kilocycle in the common base configuration.

f — Any arbitrarily chosen frequency of measurement.

t_r — Transistor rise time of the saturated common emitter switch.[4]

t_f — Transistor fall time of the saturated common emitter switch.[4]

Gm_t — Transistor transconductance, grounded emitter $\equiv \dfrac{\delta I_c}{\delta V_{be}}$.

P — Ratio of $|\beta_0|/|\beta|$.

S — Ratio f/f_β.

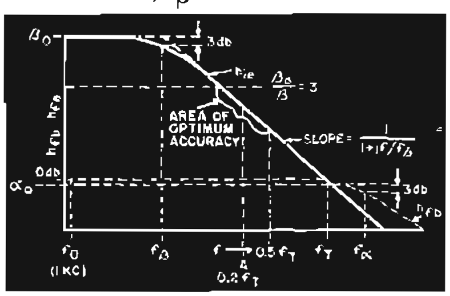

Figure 1. Location of Important Transistor Parameters on Relative Frequency and Amplitude Basis.

Method for Determining Parameters

The method used for determining transistor parameters β, β_0, h_{ie}, h_{ib}, f_β, f_T, and Kf_α, is based on the theory that β (h_{fe}) follows, to a very close approximation, the classical 6db-per-octave slope as a function of frequency (Figure 1).[2, 3] This is true for all transistors that are currently in production. This characteristic is expressed by the equation:

$$\beta = \frac{1}{1 + j\dfrac{f}{f_\beta}} |\beta_0|. \qquad (1)$$

Actually, this method requires but three simple measurements; $|\beta_0|$ on the 275-A Transistor Test Set and h_{ie} and h_{ib} on the 250-A RX Meter. These measurements are then used to characterize many low and medium-power transistors in the frequency range of 1 kc to well above 1 kmc. $|\beta|$ is computed from the two two-terminal RX Meter measurements and compared to $|\beta_0|$ which is read directly on the Transistor Test Set. The ratio of $|\beta_0|/|\beta|$ is then used in conjunction with the curve in Figure 2 to determine the various other parameters.

Two jigs are required for use in making the RX Meter measurements. Schematic diagrams of these jigs are shown in Figure 3A and C on page 3 of Notebook number 19. A suggested design for the jig is shown in Figure 5. It is recommended that the RX Meter measurements be made at a frequency (f) of approximately $0.2f_T$.

Figure 1 shows, graphically, the location of the important parameters on a relative frequency and amplitude basis. Point A ($0.2f_T$) on the straight-line section of the curve is the center of the area recommended for optimum accuracy when making high-frequency RX Meter measurements. A more detailed discussion of accuracy will be undertaken later in the article. It can also be seen in Figure 1 that f_α is always greater than f_T. It has been shown that f_α does not always adhere to a 6db-per-octave slope.[2]

Procedure for Determining Parameters

The following is a step-by-step procedure for determining transistor parameters in accordance with the method discussed in the previous paragraphs. The data obtained may be conveniently recorded on a data sheet (Figure 3) as a part of the procedure.

Determination of β

1. Measure the R_p and C_p values cor-

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responding to h_{ib} and h_{ie} directly on the RX Meter and record these values in column 1 on the data sheet.

2. Convert C_p into X_p ($-C_p$ corresponds to $-jX_p$), using the equation $X_p = \frac{1}{2\pi f C_p}$, and enter the values for R_p and X_p in column 2.

3. Convert the data in column 2 to rectangular coordinates, the series impedance. This conversion can be made by means of the series-parallel conversion chart in the 250-A instruction book and BRC Catalogs L and L-1, or by means of the equations given below. If the conversion chart is used, select a convenient multiplying factor to obtain a location of sufficient resolution on the chart.

General Equations:

$$X_s = -\frac{X_p Q^2}{1+Q^2}$$

$$R_s = \frac{R_p}{1+Q^2}$$

$$Q = \frac{R_p}{X_p}$$

Equations for Q less than 0.1:

$$X_s = -\frac{R_p^2}{X_p}$$

$$R_p = R_s$$

Equations for Q more than 10:

$$R_s = \frac{X_p^2}{R_p}$$

$$X_s = -X_p$$

Enter the data in column 3 on the data sheet.

4. Compute β using the data from column 3 and the equation:[5,6,7]

$$\beta = \frac{h_{ie} - h_{ib}}{h_{ib}} = \frac{R + jX}{R_1 + jX_1} \text{ or}$$

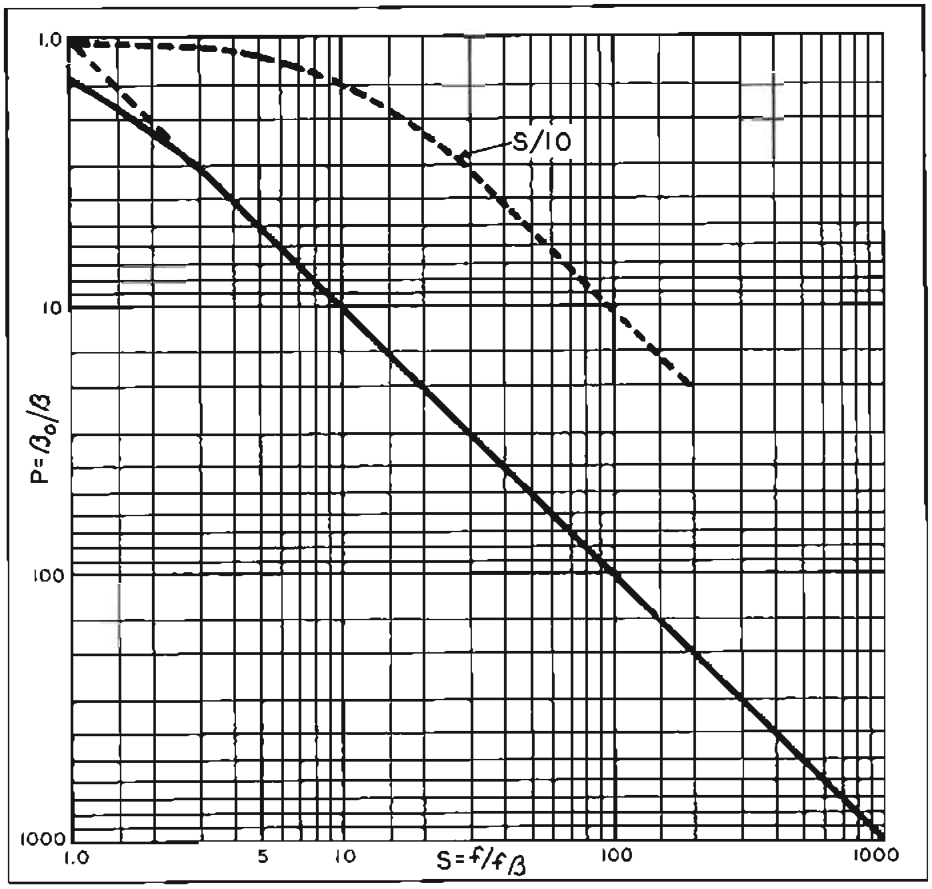

Figure 2. Nomograph for Determining Transistor Parameters

the real and imaginary terms of the numerator divided by the real and imaginary terms of the denominator. Enter this data in column 4.

5. Compute the magnitude of β as follows:

$$|\beta| = \sqrt{\frac{R^2 + X^2}{R_1^2 + X_1^2}}$$

Enter the data in column 5.

6. Measure $|\beta_o|$ and $|\alpha_o|$ directly on the 275-A. Enter these values in columns 6A and 6B.

7. Determine the ratio of the data in column 6A to the data in column 5:

$$\frac{|\beta_o|_{275}}{|\beta|_{250}} = \frac{|\beta_o|}{|\beta|} = P.$$

Enter this ratio in column 7.

8. Locate the ratio of $|\beta_o|/|\beta|$ (P) of column 7 on the vertical axis in Figure 2 and proceed horizontally to intersect the curve. Drop a line vertically and read $\frac{f}{f_\beta} = S$ on the horizontal axis. Enter this in column 8.

9. Compute $f_\beta = \frac{f}{S}$ and enter this in column 9.

Determination of f_T

10. Locate the value $|\beta_o|$, recorded in column 6, on the vertical axis of the curve in Figure 2 and proceed horizontally to the curve. Project this point vertically to the horizontal axis and read S. Compute f_T using the equation: $f_T = Sf_\beta = \beta_o f_\beta$.[3] If $|\beta|$ falls on the 6db-per-octave slope ($P \geqq 3$), $f_T = |\beta| f$. Enter the data in column 10.

Determination of β at a frequency (f) other than that used in the initial 250-A measurement

11. Determine the ratio of $f/f_\beta = S$ and locate this point on the horizontal axis of the curve in Figure 2. Proceed vertically to the curve and read the ratio $P = (|\beta_o|/|\beta f|)$ on the vertical scale. Compute $|\beta| = |\beta_o|/P$ and enter this data in column 11.

Determination of f_α and K

12. f_α may be computed using the following general equation:

$$f_\alpha = \frac{f_\beta\,(1+\beta_o)}{K} \cong \frac{f_\beta\,\beta_o}{K}.$$

In some instances, specifically in the case of a transistor with a 6db-per-octave common base current gain cutoff, K is unity and the equation in (12) becomes: $f_\alpha = f_\beta\,(1+\beta_o)$. If either K or f_α are known, the other may be computed. $Kf_\alpha = f_\beta\,(1+\beta_o)$ and can be computed from the data in columns 1 and 9 of the data sheet. To check an assumed value of one variable, the equation $f_T = \alpha_o K f_\alpha$, may be used.[2] K is a function of the manufacturing process and may be from 0.4 to 1.0. However, K does not vary appreciably from transistor to transistor of the same manufacturing process, having a value of 0.822 for uniform impurity density. The value drops further for accelerating "built-in" fields or "drift" transistors.

Example

The method and step-by-step procedure described above are used in the following example to measure and compute parameters f_β, f_T, β, and Kf_α, for a typical transistor. Data obtained from these measurements and computations is recorded on the data sheet in Figure 3.

Specifications:

$\beta_o = 15$

$f_T = 45$ mc (estimated)

Conditions:

$f = 0.2 f_T = 9.0$ mc

$f_o = 1000$ cps

$V_{CB} = 6$ v

$I_E = 1.0$ ma.

Step 1. h_{ie} and h_{ib} are measured on the 250-A in terms of R_p and C_p, and the values are recorded in column 1 on the data sheet.

Step 2. The C_p reading is converted to X_p; ($X_p = \frac{1}{2\pi f c}$) and R_p and X_p are recorded in column 2.

Step 3. The data in column 2 is converted to rectangular coordinates, series impedance, by means of a series-parallel conversion chart or the following computations.

h_{ib}

Computing Q:

$$Q = \frac{R_p}{X_p} = \frac{100}{7.5K} = 0.0133$$

Using equations for Q less than 0.1:

$$X_s = -\frac{R_p^2}{X_p} = -\frac{10{,}000}{7{,}500} = -j1.3$$

$$R_s = R_p = 100$$

$$h_{ib} = 100 - j1.3.$$

h_{ie}

Computing Q:

$$Q = \frac{R_p}{X_p} = \frac{2.2K}{1.1K} = 2$$

Using equations for Q between 0.1 and 10:

$$X_s = -X_p \frac{Q^2}{1+Q^2}$$

$$X_s = -1.1K \times 4/5 = -j880$$

$$R_s = \frac{R_p}{1+Q^2} = \frac{2.2K}{5} = 440$$

$$h_{ie} = 440 - j880$$

This data is recorded in column 3.

Step 4. Rectangular coordinate, series impedance β is computed using the data in column 3 as follows:

$$\beta = \frac{h_{ie} - h_{ib}}{h_{ib}}$$

$$\beta = \frac{440 - j880 - 100 + j1.3}{100 - j1.3}.$$

Subtracting R terms and j terms separately:

$$\beta = \frac{340 - j879}{100 - j1.3} = \frac{R - jX}{R_1 - jX_1}$$

This data is entered in column 4.

Step. 5. Compute the magnitude of β as follows:

$$|\beta| = \sqrt{\frac{R^2 + X^2}{R_1^2 + X_1^2}}$$

$$|\beta| = \sqrt{\frac{340^2 + 879^2}{100^2 + 1.3^2}}$$

$$|\beta| = \sqrt{\frac{888{,}200}{10{,}000}} = 9.4$$

This data is recorded in column 5.

Step 6. β_o and α_o are measured directly on the 275-A and recorded in column 6.

Step 7. The ratio P is determined as follows:

$$P = \frac{\beta_o}{\beta} = \frac{20}{9.4} = 2.13.$$

This ratio is recorded in column 7.

Step 8. The ratio P is located on the curve in Figure 2 and S is determined to be 1.95. Record S in column 8.

Step 9. f_β is computed as follows:

$$f_\beta = \frac{f_x}{S} = \frac{9}{1.95} = 4.6 \text{ mc}$$

and is recorded in column 9.

Step 10. f_T is determined as follows:

$f_T = S f_\beta$ or $\beta_o f_\beta$

$f_T = 20 \times 4.6 = 92$ mc

and is entered in column 10. Note that $f_T = \beta f$ cannot be used in this case because $P = |\beta_o|/|\beta|$ is less than 3. Note also, that f_T is 92 mc or considerably higher than the 45 mc estimated under "Conditions."

Step 11. $|\beta|$ is determined at 50 mc as follows:

$$S = \frac{f}{f_\beta} = \frac{50}{4.5} = 10.5.$$

S(10.5) is then located on the horizontal axis in Figure 2, and proceeding from this point, vertically, to the curve, P is read on the curve. Then, $|\beta| = \frac{|\beta_o|}{P} = \frac{20}{10.5} = 1.8$. β is recorded in column 11.

	1		2		3		4	5	6		7	8	9	10	11	12
	R_p	C_p	R_p	X_p	RECT. Z R_s	X_s	β RECT.	β POLAR	(A) β_o	(B) α_o	P	S	f_β	f_T	β 50MC	Kf_α
h_{ib}	100+2.3		100+j7.5K		100−j1.3		340−j879 / 100−j1.3	9.4	20	0.9515	2.13	1.95	4.6	92MC	1.8	115MC
h_{ie}	2.2K+16		2.2K+j1.1K		440−j880											

Figure 3. Data Sheet

Step 12. Kf_α or f_α is determined as follows:
$Kf_\alpha = f_\beta (1 + \beta_o) = 4.5 \times 21 = 95$ mc
Record Kf_α in column 12.

This indicates that K must be known to determine f_α or vice versa. If the base layer of this transistor is the uniform impurity type, K = 0.822. Then,

$$f_\alpha = \frac{4.5 \times 21}{0.822} = 115 \text{ mc}$$

Determining Parameters h_{ie}, t_r, t_f, Gm_t, h_{ic}, and h_{fc}

Transistor parameters h_{ie}, h_{ie} and h_{fe} may be readily calculated from direct measurements of α, β, and h_{ib} on the Transistor Test Set:[9]

$$h_{ie} = h_{ie} = \frac{h_{ib}}{(1-\alpha)\ (1-h_{rb}) + h_{ob} h_{ib}} \qquad (2)$$

Since, typically, $h_{rb} << 1$, $(1-h_{rb}) \approx 1$, and $h_{oh} h_{rb} << 1$, under small-signal, low-frequency conditions, equation (2) may be reduced to equation (3) which results in an error of 10% for $\alpha = 0.99$ and decreases with a decrease in α:

$$h_{ie} = h_{ie} \approx \frac{h_{ib}}{1-\alpha} \approx h_{ib}\ (1 + \beta) \qquad (3)$$

$$h_{fe} = \frac{h_{rb}-1}{(1-\alpha)\ (1-h_{rb}) + h_{ob} h_{ib}} \qquad (4)$$

Under the assumptions and within the limits of error set forth for equation (3), equation (4) may be similarly reduced to equation (5):

$$h_{fe} \approx \frac{1}{1-\alpha} \approx -(1 + \beta) \qquad (5)$$

Additional parameters t_r, t_f, (Figure 4) and Gm_t may be determined as follows.[4]

t_r

$$\text{Rise time } (t_r) \cong 0.8 \frac{I_C \text{ (sat.)}}{I_B 2\pi f_T} \qquad (6)$$

$\frac{I_C}{I_B} = \beta_{DC}$ at operating current range.

t_f

$$t_f \cong 0.8 \frac{\beta_C \text{ off}}{2\pi f_T} \qquad (7)$$

"β_C off" is the turn-off circuit β, using the circuit in Figure 4, and can be measured on the 275-A by adjusting to the proper dc bias point.[4]

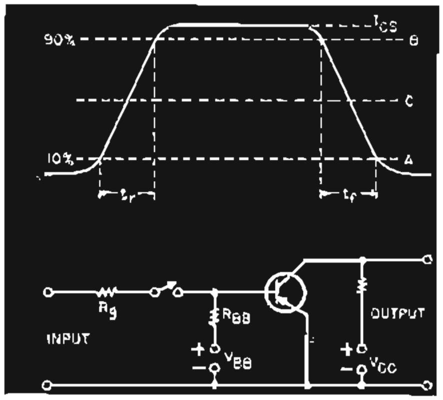

Figure 4. Transistor Switching Parameters

Gm_t*

$$Gm_t = \frac{\beta_o}{h_{ie}} \qquad (8)$$

$$h_{ie} = h_{ib}\ (1 + |\beta_o|) \qquad (3)$$

$$Gm_t = \frac{|\beta_o|}{h_{ib}\ (1 + |\beta_o|)} \cong \frac{1}{h_{ib}} \qquad (9)$$

h_{ib} can be measured directly on the 275-A.

Gm_t is derived as follows:

$$Gm_t = \frac{\delta i_c}{\delta V_{be}} =$$

$$\frac{\delta i_c}{\delta i_b} \times \frac{\delta i_b}{\delta V_{be}} = \frac{\frac{\delta i_c}{\delta i_b}}{\frac{\delta V_{be}}{\delta i_b}}$$

Since $\beta = \frac{\delta i_c}{\delta i_b}$ and $h_{ie} = \frac{\delta V_{be}}{\delta i_b}$,

$$Gm_t = \frac{\beta}{h_{ie}}.$$

An example of the procedure for determining t_r, t_f, and Gm_t will not be included in this discussion.

Accuracy and Limitations

It is obvious from the equations presented in the foregoing procedure, that the accuracy of $|\beta_o|$ is of prime importance, since all of the parameters; f_β, f_T, t_r, t_f, β, and Kf_α are based on the ratio $P = |\beta_o|/|\beta|$. If the accuracy of $|\beta_o|$ can be made to exceed the accuracy of all other measurements, the overall accuracy will be improved accordingly.

*Pointed out by H. Thanos, RCA, Somerville, N. J.

This is the case when the measurements are made on the 275-A Transistor Test Set, where the accuracy of $|\beta_o|$ is $(0.6 + 30/\beta)$% or usually less than 2%. The ratio $P = |\beta_o|/|\beta|$ is employed to take care of the possibility that the measured β at $0.2f_T$ (as per the published specifications or estimate) falls above the 6db-per-octave slope, and to preclude the need for an additional measurement. This actually happened in the example given above. For optimum accuracy, the ratio P should be greater than 3 and f should not exceed 0.5 f_T for the average transistor.

The accuracy of the high-frequency β (h_{fe}) measurements is dependent upon the RX Meter accuracy equations (See pages 10 and 11 of the 250-A instruction book.) and the relationship:

$$h_{fe} = \frac{h_{fe} - h_{jb}}{h_{jb}}.$$

For most values of h_{ib}, which is usually resistive, 3% accuracy is about average. For h_{ie} 5% accuracy can be expected. Generally, accuracies better than 10% can be expected for h_{fe}. Since the h_{fe} and h_{jb} real and imaginary terms are in quadrature, potential errors in R_p and C_p are not directly additive, but are a function RMS of the respective errors.

The above discussion of this procedure also considers some of the limitation imposed by the original assumption that β (h_{fe}) adheres strictly to the 6db-per-octave fall off common to R-C filters.

Another important consideration involves the design of the jigs for the RX Meter measurements. Good high-frequency techniques and practice must be followed to achieve the accuracies mentioned. Figure 5 suggests a jig design to minimize most of the problems encountered. It is also possible to "calibrate out" RX Meter and jig residuals by means of the technique described on page 4 of Notebook number 22, when making measurements above 20 mc.

The accuracy of t_r and t_f is affected by the variation of $|\beta|$ as a function of the current range over which the transistor is to be operated. Improved accuracy can be obtained if $|\beta_o|$ readings are taken at points approximating A, B, and C in Figure 6, and the results are averaged. If it is found that the bias current for $|\beta_o|$ average is such that h_{ib} is below the range of the RX Meter, a convenient bias current can be used and f_T can be corrected by the ratio of $|\beta_o|$ at the current of the RX Meter measurement to $|\beta_o|$ average. If $|\beta_o|$ proves to be

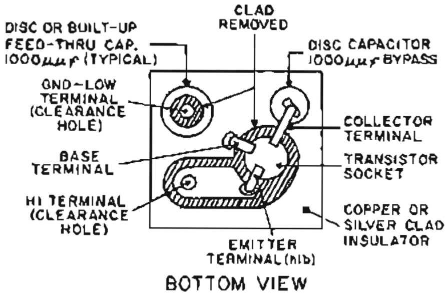

Figure 5. Suggested Design for a VHF h parameter Jig for Measuring Transistor Parameters on the RX Meter

quite constant as a function of bias conditions, which is the case for well designed switching units, corrections are not necessary and accuracies better than 20% can be expected.

Still another step may be taken if accuracies greater than those already mentioned are desired. If the frequency characteristics of the transistor are such that when $S = f/f_\beta = 0.1$, f is 500 kc or greater, a $|\beta_o|$ measurement may be made on the RX Meter at 500 kc. The $|\beta_o|$ measurements on both instruments (250-A and 275-A) can then be compared and a correction factor computed. See the broken-line curve in Figure 2. This correction, applied to subsequent β measurements on the 250-A, will yield improved accuracies.

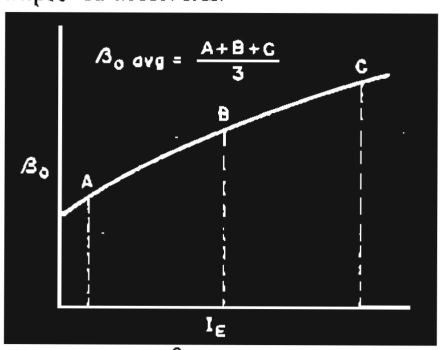

Figure 6. Typical β_o versus I_E curve for Determining β_o Average

Conclusion

We have shown that a transistor current gain characteristic can be readily determined for the common emitter configuration using the 275-A Transistor Test Set and the 250-A RX Meter. Since RF measurements, with this procedure, are made at $0.2f_T$, devices with f_T's up to 1.25 kmc can be accommodated.

The author wishes to thank Mr. C. D. Simmons of the Philco Corp., Mr. H. Thanos of RCA, and the BRC Engineering Department for their assistance during the preliminary search for material for this article.

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Diode Measurements on the Transistor Test Set Type 275-A

WILLARD J. CERNEY, *Sales Engineer*

There are many instances in diode applications when it is desirable to know the AC impedance of the diode as well as the DC resistance. A few examples of these applications are:

1. When a diode is used as a limiter or clipper.
2. When a Zener diode is used as a voltage regulator and the AC impedance is of importance.
3. When diodes are used in the design of power supply circuits. Here, knowing the AC impedance of diodes will be of value in designing ripple suppression and low-frequency coupling networks.

The terms and expressions used in this article are defined in Figure 1.

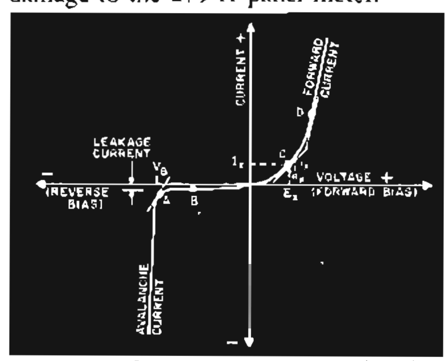
Typical values for a 1N1522 diode at points A, B, C, and D on the diode curve in Figure 1, are given in the table below.

Point	Volts	DC Res. $= \frac{E_X}{I_X}$	AC Res. $= \frac{e_x}{i_x}$
A	−8.2	820	0.2
B	−6	1200	6100
C	+1	20	1
D	+2	18	0.5

The Transistor Test Set Type 275-A may be used to measure the DC forward resistance and the forward biased AC resistance of all diodes, directly and simply, within the current and impedance ranges of the instrument (0 to 100 MA. and 0.3 to 3000 ohms). Some of the high-current diodes may also be measured using external equipment. In addition, the 275-A may be used to measure the DC breakdown voltage and resistance and the AC resistance of Zener or Regulator diodes at any point.

MEASUREMENT PROCEDURE

A step-by-step procedure for making diode measurements on the 275-A is given below. Before a diode is connected across the terminals of the 275-A, the controls should be set up in accordance with the initial set up procedure to prevent damage to the instrument or the diode under test. All of the measurements must be limited to 100 milliamperes unless external equipment is used. Care should be taken not to short the E and C terminals as this might result in damage to the 275-A panel meter.



V_B = Avalanche region, commonly referred to as the voltage breakdown or Zener point. Actually the exact voltage and current points are defined arbitrarily.

DC Resistance = Resistance at any point on curve = $\frac{E_X}{I_X}$ as shown above.

AC Resistance = Impedance at any point on curve = $\frac{\Delta E}{\Delta I}$ or $\frac{e_x}{i_x}$ as shown above.

Figure 1. Diode Curve

Initial Set Up Procedure

1. Set the α-h_{11}-β Selector to the h_{11} position.
2. Set the SET-CHECK-MIN switch to the SET position.
3. Set the Meter switch to the V_{CB} position.
4. Turn the V_{CB} Volts Range Selector and the V_{CB} control fully counterclock-

wise; i.e., $V_{CB} = 0$.
5. Set the Meter switch to the I_E position.

Forward Biased Measurements (All Diodes)

To measure DC forward resistance and AC resistance when forward biased, set up the instrument controls in accordance with the initial set up procedure then proceed as follows:
1. Connect the diode to be measured between terminals E and B on the 275-A.
2. Set the I_E controls for the desired biasing current.
3. Set the SET-CHECK-MIN switch to the CHECK position and the NPN-PNP switch to the position that gives the highest reading on the panel meter.
4. Connect an external DC VTVM (such as the HP 412A) across the diode under test.
5. Set the SET-CHECK-MIN switch to the MIN position and read the VTVM and I_E. The DC forward resistance equals the VTVM reading divided by the I_E reading.
6. Set the Meter switch to the MIN position and adjust the h_{ib} dial for a null meter reading.
7. The h_{ib} reading is the small signal forward biased AC resistance.

Reversed Biased Measurements (Zener or Regulator Diodes)

If the breakdown voltage of the Zener or Regulator diode to be measured is less than 6 volts DC, perform the following procedure in addition to the procedure for the forward biased measurements.
1. Set the Meter switch to the I_E position.
2. Set the I_E controls for the desired biasing current.
3. Set the NPN-PNP switch to the position opposite to that used in step 3 in the forward biased measurement procedure.
4. Read the external VTVM and I_E.
5. The DC breakdown resistance equals the VTVM reading divided by the I_E reading.
6. Set the Meter switch to the MIN position and adjust the h_{ib} dial for a null meter reading.
7. The h_{ib} reading is the breakdown or operating AC resistance.

If the breakdown voltage of the Zener or Regulator diode is greater than 6 volts DC, perform the following procedure in addition to the procedure for forward biased measurements.
1. Set the SET-CHECK-MIN switch to the SET position.
2. Change the diode connection from the B terminal to the C terminal on the 275-A. Do not disturb the connection at the E terminal.
3. Set the Meter switch to the I_E position.
4. Set the I_E controls for the desired biasing current.
5. Set the NPN-PNP switch to the position opposite to that used in step 3 in the forward biased measurement procedure.
6. Set the SET-CHECK-MIN switch to the MIN position.
7. If the I_E reading on the panel meter increases, turn the V_{CB} control clockwise until the meter indicates the desired biasing current.
8. Read the external VTVM and I_E.
9. The DC breakdown resistance equals the VTVM reading divided by the I_E reading.
10. Set the Meter switch to the MIN position and adjust the h_{ib} dial for a null meter reading.
11. The h_{ib} reading is the breakdown or operating AC resistance.

Measurements Above 100 MA.

To measure diodes above 100 milliamperes on the 275-A, proceed as follows:
1. Connect an external power supply as shown in Figure 5 on page 10 of the 275-A instruction book. The resistance of the supply should be high (constant current) so that it does not shunt down the diode impedance.

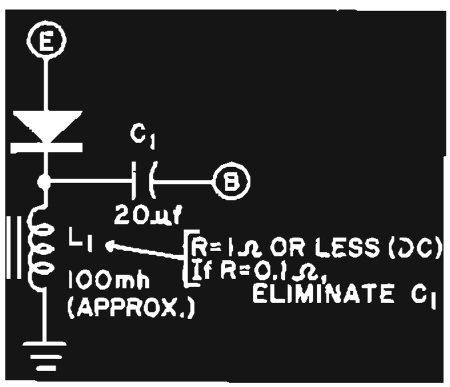
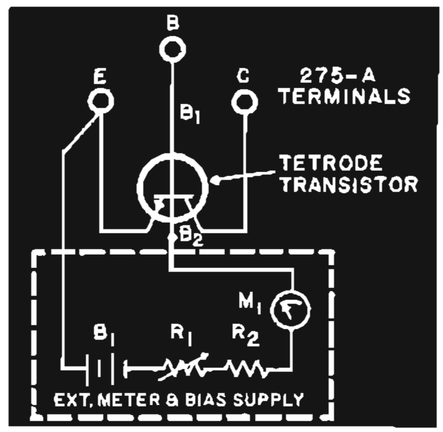


Figure 2. Connections for Measuring Diodes Above 100 MA.

2. Connect the diode to be tested and the choke to the 275-A terminals as shown in Figure 2.
3. Follow the procedure for forward biased measurements.

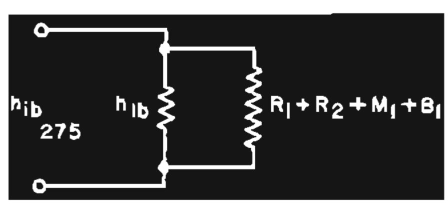
Measuring Tetrode Transistors on the 275-A

Tetrode transistors can be conveniently and directly measured on the Transistor Test Set Type 275-A with the same accuracies specified for standard triode units, by employing a simple external meter and bias supply. The correct connections for this measurement are shown in the figure below. After the connections are made to the 275-A, the I_E, V_{CB}, and I_{B2} in the external bias supply are set to the desired values and the 275-A is operated in the normal manner to measure α, β, and h_{ib}.



Connections for Measuring Tetrode Transistors on the 275-A

In making these measurements, the following notes should be observed:
1. The external bias supply impedance should be very large with respect to h_{ib} in order to reduce the loading effect of the bias supply on the h_{ib} reading.
2. B_1, R_1, R_2, and M_1 in the figure should be selected to give the desired current range and at the same time provide the very large impedance required per Note 1. If it is not possible to select values of R_1 and R_2 that are much larger than h_{ib}, the effect of the external components may be determined by calculation as follows:



Equivalent Circuit

$$h_{ib275} = \frac{h_{ib}\,(R_1 + R_2 + M_1 + B_1)}{h_{ib} + (R_1 + R_2 + M_1 + B_1)}$$

Solving for h_{ib}:

$$h_{ib} = \frac{h_{ib\,275}\,(R_1 + R_2 + M_1 + B_1)}{(R_1 + R_2 + M_1 + B_1) - h_{ib\,275}}$$

Note: B_1 and M_1 can usually be chosen to be negligible compared to R_1 and R_2.

Typical measurements for a 3N36 tetrode transistor are:

$V_{CB1} = 5V$
$I_E = 1.5MA.$
$I_{B2} = 0.91MA.$
$h_{fe} = 12$
$h_{fb} = 0.9227$
$h_{ib} = 28$ ohms

Typical values for external bias used in the above measurements are:

$R_1 = 10K$ ohms
$R_2 = 2.2K$ ohms
$M_1 = 0\text{-}2MA.$
$B_1 = 2V$

MEET OUR REPRESENTATIVES

VAN GROOS COMPANY

The Van Groos Company was formed in 1945 by J. C. Van Groos and presently operates with headquarters in Woodland Hills, California and a branch office in Los Altos.

"Van" grew up with the West Coast electronics business. He began his engineering career at the University of California. Later on he was Maintenance Engineer for the McClatchey Broadcasting Chain and prior to World War II he entered the sales engineering field in California. During the latter part of the War, he was in charge of ground electronic equipment at all Naval Air Stations.

J. C. Van Groos

Since its inception, the credo of the Van Groos Company has been complete service to the customer, and to implement this customer service concept. Van conceived the idea of a mobile demonstrator to meet the needs of the dynamic electronic industry in California.

In 1956 a 30-foot Flexible bus was converted into a mobile demonstrator known as "Groosvagen". During the first year this mobile unit was used to demonstrate to more than 15,000 engineers in the California area. The "Groosvagen" was so effective, in fact, that in 1958 the Van Groos Company put into service another mobile demonstrator known as "Groosvagen II". This unit was bigger and better with such added features as a mobile telephone, air conditioning, and a self-contained generator, and proved to be more popular than ever with the West Coast engineers. Van advises, in fact, that it has been difficult to keep up with the demand for mobile demonstrations.

Since the beginning of 1960 an air conditioned service and calibration laboratory, under the supervision of Mr. Vic Howard, has been in operation at the Van Groos Company's Woodland Hills office. This new facility has been added to provide local repair service on all instruments with emphasis on minimum down-time for the customer.

Interior of the Van Groos Mobile Demonstrator "Groosvagen II"

During his career Van has been an active member of the IRE and is currently a senior member. Two years ago he founded and served as Chairman of the San Fernando Sub-Section which now has 1500 members. He also has been an Amateur Radio enthusiast since 1930 and is holder of Radio Amateur License W6GFY.

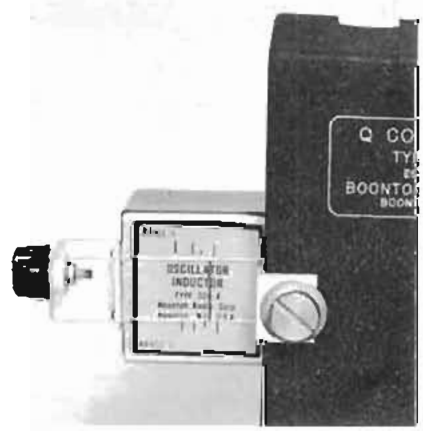
The Van Groos Company has built its success in the electronic test equipment field by emphasizing service combined with integrity. We at BRC proudly salute the Van Groos Company for their faithful service to our many valued customers throughout California.

SERVICE NOTE

Modification of Type 265-A Q Comparator for Improved Stability

Beginning with Q Comparator Type 265-A, Serial No. 70, an auxiliary mounting bracket has been added to the plug-in Type 520-A Oscillator Inductors to provide a more rigid mounting for the inductor when it is plugged into the oscillator circuit on the Detector Unit. The new bracket securely clamps the 520-A inductor to the top of the Detector Unit cover so that it cannot shift in its socket with vibration from the capacitor drive motor. Some customers had advised BRC that this vibration would often cause a shift in oscillator frequency noticeable as a shift in the CRT display on the Indicator Unit.

In order to provide a means whereby this feature could be incorporated into equipment already in the field, a special field modification kit has been prepared and distributed to our representatives and in some cases, directly to our customers. If there are any owners of Q Comparators with serial numbers below 70 who have not received this modification kit, they may be obtained by calling or writing BRC.



The New Type 520-A Oscillator Inductor Mounting Bracket

EDITOR'S NOTE
Q Meter Winner

Many Notebook readers have written to the Notebook inquiring about the winning estimate in the Q Meter Contest held last March at the IRE show. We announced the name of the winner in our Spring issue, but neglected to give his estimate. Mr. Byers' estimate was 394.

Mr. Byers informed us in a letter that he was both pleased and greatly surprised when he heard the news of his winning the contest. "I frequently use both your models 160-A and 190-A in working with coils for signal generators," he writes, "but your contest coils usually bear little resemblence to coils with which I am familiar."

It is odd that Mr. Byers should make this comment, because we have received similar complaints from other contest hopefuls. Be assured that we have passed these complaints along to the Engineers responsible for the design of the contest coils. Each year, however, the coils become more and more incredible. Apparently they have become obsessed with their fiendish task.

William F. Byers

Mr. Byers, our contest winner, was graduated from Ohio State University with a BS in Electrical Engineering in 1943. After teaching at the University for a brief period, he joined the General Radio Company and has been with that Company ever since, engaged in the design and development of special purpose and standard signal generators, frequency modulation equipment, and broadcast monitoring equipment.

Mr. Byers is a member of the Institute of Radio Engineers and the American Radio Relay League, and is holder of Amateur Radio Station License W1NXM.

Our congratulations again to Mr. Byers. We are sure that he will make good use of the Q Meter.

BRC APPOINTS NEW SALES REPRESENTATIVES

Boonton Radio Corporation is pleased to announce the appointment of the George H. Sample Company and the S. Sterling Company as sales representatives. The George H. Sample Company will be our exclusive representative in Australia with its headquarters in Melbourne. The S. Sterling Company will be our exclusive representative in Michigan, West Virginia, and sections of Ohio and Pennsylvania, with its main offices in Detroit.

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BOONTON RADIO CORPORATION

WINTER, 1961

NUMBER 27

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

Design of a UHF Q Meter

CHARLES G. GORSS, *Development Engineer*

With the intensive development presently going on in the ultra-high frequency area of the frequency spectrum, the need for a Q Meter capable of measurements in this frequency range has become evident. It was inevitable that the task of designing such an instrument would be undertaken by Boonton Radio Corporation, pioneer in the field of Q Meter design.

Q Meters presently in use are limited to measurements at frequencies below 300 Mc. This limitation is due mainly to certain design characteristics inherent in these instruments which have precluded the possibility of their use at UHF; namely, the injection resistance does not remain constant, resulting in poor calibration at higher frequencies; high series inductance is introduced into the measuring circuit at higher frequencies; and the oscillator design is not suited to UHF operation.

This article describes how these and other design problems were solved by the BRC Engineering Department during the course of the development of the new UHF Q Meter Type 280-A (Figures 1 and 2), an instrument which measures Q from 10 to 25,000 over a frequency range of 210 to 610 Mc.

Figure 1. UHF Q Meter Type 280-A

Direct Reading, Self-Correcting Q Capacitor

The key to the development of the UHF Q Meter lay in the design of the Q capacitor, for without a workable Q capacitor, a UHF Q Meter would not be practical. A concept of a true reading capacitance was selected for the Q capacitor design in the UHF Q Meter.

If a capacitor (C) has a series inductance (L), which is characteristic of all capacitors, the equivalent capacitance (C_{eq}) is given by the equation:

$$C_{eq} = C \times \frac{1}{1-\omega^2 LC}, \text{ where } \omega =$$

operating frequency times 2π.

In the usual case L may vary with C. For example, in a butterfly-type capacitor, L and C vary in the same direction. In certain other type structures, such as the capacitor structure used in the BRC Type 190-A high-frequency Q Meter, L is almost constant.

As an interesting possibility, assume that L will vary inversely with C, so that L times C is a constant. This is equivalent to the series resonant frequency being constant, and independent of capacitance. Then, at a given frequency, C_{eq} would be equal to a constant times C, and the error (difference between C_{eq} and C) would be a constant percentage; this percentage being a function of frequency only.

In this case, if the readout scale for C were made logarithmic, a simple single motion of the readout index would produce a constant percentage correction in the C readout, and the system would provide a true capacitance reading at any frequency level.

Construction of the Q Capacitor

Having established the fact that series L times C should be a constant and that the law of variation of capacitance should be logarithmic, a practical way of constructing such a device had to be found. With high Q as an objective, sliding contacts were ruled out because

YOU WILL FIND . . .

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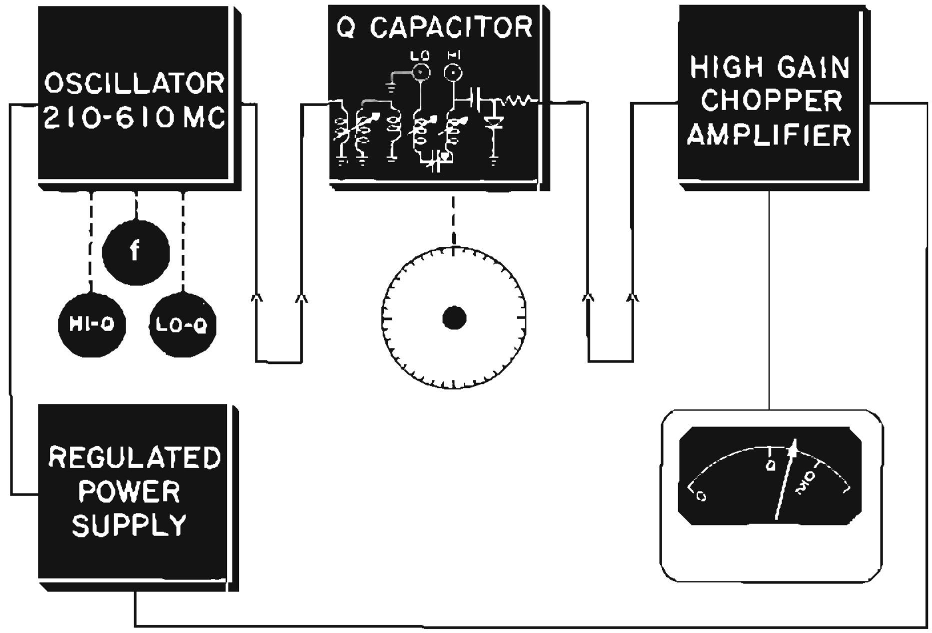

Figure 2. Block Diagram — UHF Q Meter

they are known to introduce unwanted and unpredictable variable resistance. A logical solution was a two-stator capacitor with insulated movable plates meshing both stators. The plate material would need to be of highest conductivity and the unit would have to be small and well shielded to prevent radiation loss. The constant series resonant frequency of the capacitor should be at least twice the maximum frequency of use.

Using the relationship that the equivalent high frequency capacitance equals the low frequency capacitance times the factor $\dfrac{1}{1-\left(\dfrac{F}{F_c}\right)^2}$ (derived from previous equation), where F equals the operating frequency and F_c equals the constant series resonant frequency of the unit, the correction factor is 1.33 (when $F/F_c = 1/2$). This correction factor is quite high compared with the anticipated accuracy of $\pm$ 5%.

The constant L times C product suggested a capacitor with an average internal path length which would decrease as the capacitance was increased. The two most likely motions to accomplish a varying capacitance are rotation and translation, with translation being defined as motion in a straight line, and rotation as the angular movement of a shaft about its axis. Translation was chosen for our purpose because it allowed the plate area to be moved toward the capacitor terminals at the same time that the capacitance is increased.

From the start it was obvious that binding posts, as we know them, would introduce too much inductance for the high resonant frequencies anticipated. Therefore, the plane of reference concept was adopted, with the two capacitor stators presenting a common plane surface, separated by an air gap. The stator surfaces would be tapped for terminal screws which would be used for connection of the components to be measured. However, the calibrated capacitance would appear at the reference plane only.

These design requirements were met in the Q capacitor structure shown in Figure 3. The translator plates approach from the bottom allowing the effective path from the plane of reference to the capacitor to be advanced at the same time that the capacitance is increased. The curve on the rear section of the translator plates is designed to give the translator plates the logarithmic capacitance variation required. The plane of reference is slanted 20° to allow the translator to approach the plane of reference as closely as possible. This tilting device maintains the approach distance to a point where the inductance is kept low, and at the same time provides a heavier stator section for attachment of components.

To give the reader some idea of the size of the capacitor structure in the instrument, the total width of the plane of reference is only 0.5 inch, and the air gap between the high stator and the ground stator is only 0.020 inch.

The electrical requirements of the capacitor were translated into mechanical dimensions by considering the structure to be a series of transmission lines of various impedance levels. The structure was then analyzed as a series of three transmission lines, one butted on to the next, with an open end and with a constant total length. The shaped translator plates approach from the bottom (Figure 3) out of the rectangular ground stator which is large enough to contain the entire translator, except for two small support tabs. The resulting self-resonant frequency is around 2000 Mc, or higher than the two-to-one requirement previously mentioned. The distance from the translator to the front terminals varies at approximately the proper ratio. The stators are wedge shaped to provide a larger section for the terminal screws and to give additional support to the structure. The entire unit is well shielded to prevent spurious resonant structures.

Linear ball bushings are used to support the translator plates so that there is virtually no play in the translator plates as they are moved toward the plane of reference.

L-C Dial Correction

The reason previously given for designing a capacitor with constant L

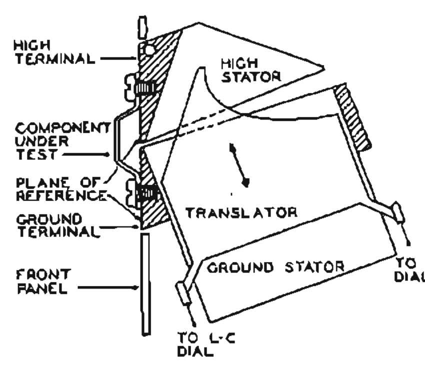

Figure 3. Q Capacitor — UHF Q Meter

times C, and for providing logarithmic capacitance variation was that the dial would be correctable by a constant percentage at any given operating frequency, and that this percentage would be equivalent to a given angular rotation of the readout hairline with respect to the capacitance scale. This is accomplished in the UHF Q Meter by pivoting the readout hairline for the capacitance dial on the same center as the capacitance dial. The hairline is rotated as a function of the frequency dial rotation by means of cam devices designed in accordance with the previously mentioned correction formulae. Readout of effective RF capacitance is therefore automatically accomplished with the tuning of the oscillator.

As an aid to computation, a concentric spiral logarithmic inductance dial is pivoted on the same shaft as the capacitance dial. The spiral inductance dial and the capacitance dial are held together by means of a friction disk. For each operating frequency used, there is an alignment of these two dials which results in the inductance scale reading that inductance which will resonate with the effective capacitance. The mechanics of driving the translating capacitor from a rotary shaft motion are accomplished by a conventional rack and pinion drive which is spring loaded to prevent backlash.

Circuit Coupling

Input is inductively coupled to one side of the high stator and output is capacitively coupled to the opposite side of the high stator, with the high stator serving as a shield between the two. The output is a voltage probe and the input is a current probe.

In order that the input coupling is only as much as is needed to give suitable output on the voltage probe for a wide range of circuit Q conditions, the input coupling has been made variable. Output from the oscillator is terminated in the movable probe of a cut-off type piston attenuator. The movable probe is a 50-ohm termination, resulting in a low standing wave ratio on the oscillator output line. A small loop at the end of the attenuator tube couples to a 50-ohm line which in turn enters the Q capacitor enclosure. This line is shorted near the front terminals of the Q capacitor with a small loop which couples to the Q circuit. The entire 50-ohm line is very short and nearly lossless and resonates at approximately 1400 Mc. Therefore, if the attenuator piston is decoupled, negligible loss is injected into the Q circuit. The advantage of this scheme is that for low Q circuits, where loss is less important and high injection level is needed, the piston is closely coupled; and for high Q circuits, where low injection level is required, the piston with its resistive component is decoupled from the Q circuit.

The voltage probe consists of a 1N82 diode coupled very loosely by a capacitive probe to the high stator. This diode looks like 4500 ohms in parallel with 0.5 pf capacitance. Voltage from the Q circuit is divided by a very small coupling capacitor providing a voltage ratio of 25 to 2, and resulting in a resistance ratio of 156 to 1. The diode appears across the Q circuit as roughly 0.7 megohms, limiting the Q of the Q capacitor to somewhat over 3,500. This value is considerably higher than the Q of small components suitable for measurement at the Q capacitor terminals.

Oscillator

In order to circumvent the previously mentioned problems associated with measuring resonant rise in the UHF range, a different principle of Q measurement has been employed in the UHF Q Meter. This principle is derived from the well known relationship that Q is equal to the frequency of resonance divided by the bandwidth from 3db point to 3db point on the resonance curve. This relationship is extremely accurate for values of 10 and above.

A number of automatic methods for sweeping this bandwidth to provide an automatic Q readout were considered, but it is likely that these methods would complicate the instrument and render it less accurate and reliable. It was decided, therefore, that the most direct approach to this type of Q measurement would be by manual, mechanical tuning of the oscillator. If this mechanical tuning were properly coupled to a dial, the dial itself could be made to readout Q directly. The oscillator frequency in this case would be an exceptional function of shaft rotation, and a given angular rotation of the oscillator shaft would be the same percentage of the oscillator frequency, regardless of the shaft position. The oscillator vernier would be calibrated in Q, starting at infinity and progressing down the Q range.

To measure Q with this system, an operator would start a measurement with the vernier dial reading ∞ at one 3db point and then tune through resonance, stopping at the other 3db point (Figure 4). The dial could be made to read Q directly. This system is simple and straightforward and a modulation system is not required. Use is made of basic mechanical elements which are necessary in an oscillator in any case, and the only additional requirement imposed on the design is that the oscillator vernier be somewhat refined and calibrated in Q. The indicator is a simple square-law diode detector, free of complex demodulator circuitry. This represents a Q measurement broken down into its fundamental essentials.

The oscillator frequency, as a function of shaft rotation, is very important in reading Q accuracy. The law must follow the general form: $f = Ae^{K\theta}$, if the Q dial calibration is to be accurate. Therefore, the oscillator structure should be repeatable. Two forms are apparent in which the frequency is controlled chiefly by mechanical parts: one employs tuned transmission lines and the other is a "butterfly" type construction. Both types employ a rigid mechanical resonator, but because the transmission line oscillator would tend to be noisy and cause frequency jumping over small motions, the "butterfly" type was chosen for this application.

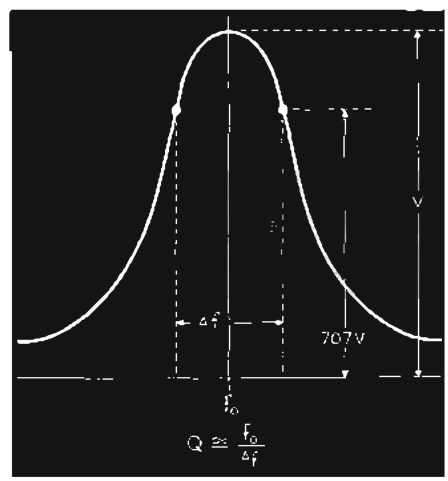

Figure 4. Q Resonance Curve

Vernier Tuning System

Q can be measured by means of a gear reduction system on the oscillator shaft up to a point; however, when the motion of the gear drive is reduced to its most infinitesimal increment, this motion becomes erratic. This could be expected for the measurement of high circuit Q values. To avoid this problem, an independent vernier was devised for the measurement of Q values beginning at 200. Basically, this fine vernier is achieved by rotating the entire butterfly stator by a micrometer driven torsion spring system, with the rotor shaft held fixed. The system is shown schematically in Figure 5. The disk represents the stator support and the four lines marked "S" are spring-tempered beryllium copper. These springs are stiff in a radial direction but permit rotation when the micrometer is advanced. The micrometer screw provides the precise uniform motion required. The springs flex elastically, in exact relationship to the micrometer motion, so there is no lost motion or backlash in the system.

The oscillator main drive is a double-ended shaft driven at right angles by a precision worm. Both vernier drives have a lock and a clutch between the shafts and the dials which are operated by means of front panel controls.

Oscillator Output System

In order to insure sufficient isolation between the test circuit and the oscillator, the oscillator output should be high. For the existing voltmeter sensitivity 0.1 watt would be sufficient for most cases. Because of the losses which may occur in makeshift external resonator couplers, it was decided that the oscillator output should be close to 1 watt RF. This output is just high enough for most requirements, without sacrificing stability.

The oscillator tube is a GENELEX DET22, with dc power handling capabilities of 10 watts. It is a planar type tube, and therefore has a very high series resonant frequency, assuring consistency of oscillator design and consequently uniformity of the law of frequency variation as a function of rotation ($f = Ae^{K\theta}$) from unit to unit. Ideally, a plot of θ versus the log of frequency should be a straight line with a slope which is the same for all instruments. This slope is held within $\pm$ 15% of nominal for all instruments at all points on the curve from 210 to 610 Mc. Operating conditions of the butterfly oscillator tend to vary considerably across the band. In order to keep the power level in the oscillator tube reasonably constant, a constant current pentode is connected in the cathode return of the tube. This holds the current change within reasonable limits without a large series dc drop. The oscillator has been carefully designed to eliminate spurious parasitic resonances which might cause the output amplitude or frequency to change at a rapid rate and thereby affect

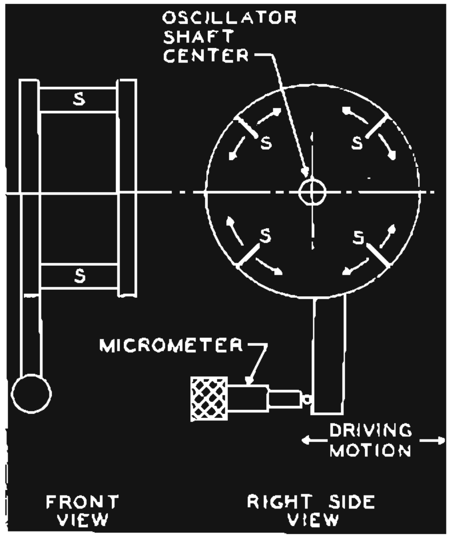

Figure 5. High Q Vernier Tuning System

the accuracy of the measurement.

Mechanical Design Parameters

Mechanical design parameters were derived from existing formulae for butterfly resonators. These parameters did not accurately elevate the inductance in the minimum capacitance condition, when the rotating plates fill the open area. The final shape of the capacitor plates was arrived at by first designing from the best known formulae, and then correcting this shape on the basis of data taken from this structure.

With nearly 10 watts being lost in the plate structure of the oscillator tube, it was necessary to provide excellent heat conduction from the plate to the general mass of metal in the stator, where it could be dissipated by radiation. The plate mount itself is a solid copper casting soldered fast to the butterfly stator. Ventilation around the outside of the oscillator is maintained at a relatively high level by the use of adequate clearance between the oscillator and other units in the instrument, and by means of perforations in the instrument cabinet.

Voltmeter System

The output from the diode probe is approximately 20 microvolts dc when the resonant peak voltage is 0.025 volts rms. In order to work with this low voltage level, a high-gain dc amplifier was necessary. A photo-conductive chopper amplifier circuit,[1] is used. This device employs light-sensitive resistance elements. Light is interrupted periodically by a mask rotated by a synchronous motor. Sharply tuned filters are used to remove noise and synchronous detection is used to better the efficiency of recovery. This unit can be operated on a 50-cycle power source by merely changing a plug-in filter unit.

Five steps of sensitivity have been provided by means of a front panel switch. The switch has five positions which equally divide the sensitivity ranging between 25 millivolts RF and 250 millivolts RF fullscale. Each step is approximately 3 to 1 dc sensitivity. Since the detector is square law, a 100 to 1 dc range represents a 10 to 1 RF level range. The switch steps, therefore, are roughly 5db.

If the Q to be measured is high and the test voltage is not critical, the least sensitive range would be used because the fluctuation noise is least and the response time of the voltmeter is shorter. At the higher sensitivities, the time of response is somewhat longer and zero fluctuation noise is noticeable. For very low Q devices, where it might not be possible to develop 0.25 volt across the resonator, maximum sensitivity would be used.

Provisions for Measurement of External Resonators

The coaxial cable which connects the oscillator to the Q capacitor and the cable which connects the dc voltmeter to the Q capacitor are jumpered at the rear of the instrument to allow for connection of external resonating devices. A suitable inductive probe connected to the oscillator output, to present a reasonable 50-ohm termination, could be used to lightly couple to the circuit under test and a small shunt diode could be used for the pickup. In this manner the external resonator would simulate, very closely, the internal resonator, and its Q and resonant frequency could be determined readily.

This application represents a remarkable advance in the Q Meter art. Previously, the Q Meter could only resonate on its terminals and therefore presented to these external resonating devices a non-reducible minimum shunt capacitance. With the UHF Q Meter the devices can be measured without significantly adding capacitance and changing the internal impedance of the external resonating device. In this respect, it would be well to point out that the UHF Q Meter has a lower minimum capacitance at its terminals (only 4 pf)

[1] This circuit is similar to that which is used in the H-P 425A dc voltmeter.

than the previous Q Meters in any frequency range.

A more detailed discussion of external resonator measurements will be given in Notebook Number 28.

Power Supply

A 300-volt, electronically regulated power supply furnishes all of the power for the oscillator and the dc voltmeter circuit. Power voltages for the voltmeter are supplied through dropping resistors. Two regulated filament dc power supplies are required: one for the oscillator which has common cathode and heater connections, and the other for the constant-current pentode and low-level stages in the dc voltmeter. These dc supplies are transistor regulated with a dc Zener diode reference. Two 6.3-volt ac supplies are provided for noncritical filament and bulb lighting.

Conclusion

Development of the UHF Q Meter Type 280-A has brought about a number of significant advances to the Q Meter measurement art. First, it has made possible the direct reading, without correction, of Q, inductance, and capacitance. Second, the frequency range for Q measurements has been extended to 610 Mc. Third, the resonant voltage has been lowered from approximately several volts, which was a function of Q, to 0.025 volts, which is constant for a measurement; opening the field for measurement of semiconductor devices and other non-linear impedances. Finally, a unique means has been provided for measuring the Q of external resonators with Q's up to 25,000, with negligible circuit loss due to the measurement.

Specifications

RADIO FREQUENCY CHARACTERISTICS

RF Range: 210 to 610 MC
RF Accuracy: ±3%
RF Calibration: Increments of approximately 1%
RF Monitor Output: 10 mv. minimum into 50 ohms*
* at frequency monitoring jack

Q MEASUREMENT CHARACTERISTICS

Q Range:
Total Range: 10 to 25,000*
High Range: 200 to 25,000*
Low Range: 10 to 200
* 10 to approx. 2,000 employing internal resonating capacitor
Q Accuracy: ± 20% of indicated Q
Q Calibration:
High Q Scale: Increments of 1-5% up to 2,000
Low Q Scale: Increments of 3-5%

INDUCTANCE MEASUREMENT CHARACTERISTICS

L Range: 2.5 to 146 mμh*
* actual range depends upon measuring frequency
L Accuracy: ± 11 to 15%*
* accuracy depends upon resonating capacitance
L Calibration: Increments of approx. 5%

RESONATING CAPACITOR CHARACTERISTICS

Capacitor Range: 4 to 25 μμf
Capacitor Accuracy: ± (5% + 0.2 μμf)
Capacitor Calibration: 0.05 μμf increments, 4-5 μμf
0.1 μμf increments, 5-15 μμf
0.2 μμf increments, 15-25 μμf

MEASUREMENT VOLTAGE LEVEL

RF Levels: 25, 40, 80, 140, 250 mv. nominal*
* across measuring terminals

PHYSICAL CHARACTERISTICS

Mounting: Cabinet for bench use; by removal of end covers, suitable for 19" rack mounting
Finish: Gray wrinkle, engraved panel (other finishes available on special order)
Dimensions: Height: 12-7/32" Width: 19" Depth: 17"
Weight: Net: 72 lbs.

POWER REQUIREMENTS

280-A: 105-125/210-250 volts, 60 cps, 140 watts
280-AP: 105-125/210-250 volts, 50 cps, 140 watts

A VHF Telemetering Signal Generator System

WILLARD J. CERNEY, *Sales Engineer*

The Type 202-G FM-AM Signal Generator and the Type 207-G Univerter (Figure 1) were designed specifically for measuring the performance of telemetering systems and equipment.[1] With a frequency range of 195 to 270 Mc, the 202-G Signal Generator is ideally suited for checking telemetering receivers since this frequency range completely blankets the recently extended 215 to 260 Mc telemetry band. The 207-G is a unity gain frequency converter which is used in conjunction with the 202-G to provide additional frequency coverage of 0.1 to 55 Mc in the intermediate frequency range.[2]

Description of the Type 202-G

A functional block diagram of the Type 202-G FM-AM Signal Generator is shown in Figure 2. The instrument consists essentially of an oscillator, a reactance modulator for FM, a pair of frequency doublers, an audio oscillator, a regulated power supply, an output network, and associated monitoring meters.

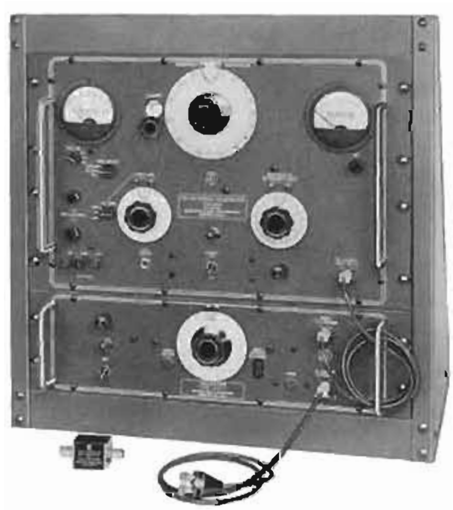

Figure 1. Rack Mounted View of 202-G and 207-G

When a voltage is applied to the grid of the reactance tube, a proportional shift in frequency is accomplished. This modulating voltage is applied to the FM terminals and coupled through a network to the grid of the reactance tube. The band pass of this network is 30 cps to 200 kc. Modulating voltage may be selected internally from any one of seven standard RDB subcarrier frequencies, or from any suitable external oscillator capable of furnishing frequencies in the range of 30 cps to 200 kc. AM may be obtained by means of a front panel internal modulation switch which places the audio signal on the screen grid of the second doubler tube. Simultaneous AM and FM may also be obtained by using an external oscillator.

The RF unit in this instrument is mounted on a rugged aluminum casting and is thoroughly shielded. This results in excellent stability, low leakage, and reliability.

The instrument is designed for bench use or for installation in a standard 19-inch rack. All of the operating controls

1. H. J. Lang, "A Telemetering FM-AM Signal Generator," BRC Notebook Number 21, Spring, 1959.
2. The 207-G may also be used to extend the frequency ranges of the Types 202-D and 202-F Signal Generators.

are arranged in convenient functional order on the front panel. All calibrated controls are direct reading.

Description of the Type 207-G

A functional block diagram of the Type 207-G Univerter is shown in Figure 3. The instrument consists essentially of a mixer, a local oscillator with a nominal center frequency of 195 Mc, two broad-band amplifiers, an output stage, and a power supply. Input from the 202-G Signal Generator is fed into the mixer from which is subtracted the output frequency of the local oscillator. The resulting difference signal is then separated by filtering and amplified. The second stage amplifier drives the output stage which is a cathode follower used to provide a suitable source impedance. Like the 202-G, the 207-G may be operated on a bench or installed in a standard 19-inch rack. All operating controls are located on the front panel.

Interconnection of Types 202-G and 207-G

The frequency range of the 202-G Signal Generator may be extended to provide intermediate frequencies of 0.1 to 55 Mc by interconnecting the instrument with the 207-G with the output and patching cables furnished with the instruments, as shown in Figure 1. The Type 501-B Output Cable is connected to the 207-G unity gain output and the Type 502-B Patching Cable is connected between the 202-G output and the 207-G input. These connections reproduce the FM and AM characteristics of the 202-G at the 207-G output.

A Type 509-B Attenuator is supplied with the 207-G for use in cases where the signal level required is low compared to the constant noise level of the 207-G. Used at the output of the 207-G, the 20-db pad attenuates both the signal level and constant noise level. This permits the use of a higher input signal from the 202-G thus improving the signal-to-noise ratio.

Applications

The Type 202-G Signal Generator and Type 207-G Univerter, when used together, provide a signal generator system especially suited for measuring performance and aligning VHF telemetering equipment. Some of the major applications of this system are listed below.

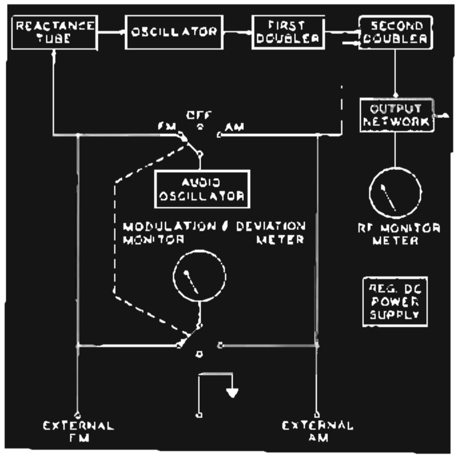

Figure 2. Block Diagram — Type 202-G

1. *Receiver Alignment* — The system may be used to check and align IF and RF amplifiers and local oscillators. Tracking may also be checked in cases where VFO equipment is used. RF checks are performed with the 202-G and IF checks are performed with the 207-G and 202-G interconnected.

2. *Receiver Bandwidth Measurements*—The calibrated output system in the 202-G makes the instrument a convenient tool for performing normal bandwidth measurements. These measurements are made by first disabling the AGC in the receiver and then determining the half-power points on the response curve. Through the use of a precision backlash-free gear train, approximately 2,200 vernier logging divisions are provided to aid in making frequency measurements. Each logging division changes the output frequency approximately 34 kc. The minimum bandwidth that can be measured without use of the 207-G is determined by the required accuracy of the bandwidth measurement. For example, if the required accuracy of the measurement is approximately 10%, then the minimum bandwidth directly readable from the 202-G would be 350 kc. If the bandwidth is narrower or requires greater accuracy, it will be necessary to measure the bandwidth of the IF amplifiers at intermediate frequencies, using the calibrated dial on the 207-G. This technique is usually acceptable, since the IF amplifiers generally control the overall bandwidth of the receiver.

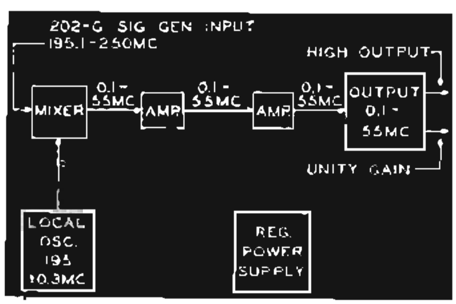

Figure 3. Block Diagram — Type 207-G

3. *Receiver Sensitivity Measurements*—To fully analyze FM receiver performance, three measurements must be made. These are maximum sensitivity, quieting sensitivity, and deviation sensitivity. The 202-G provides a calibrated output of 0.1 μv to 0.2 volt which is ideally suited for these measurements. Maximum sensitivity is generally defined as the minimum amount of a specified carrier with a standard modulation that will produce a standard output with all controls set for maximum gain. In modern telemetering systems, it is not uncommon to have receivers with sensitivities of 0.5 μv or better. Quieting sensitivity is usually specified as the minimum unmodulated carrier signal required to reduce the output noise by a specified amount below the output with standard test modulation. A typical receiver required 3 μv to attain 20 db and 5.5 μv to attain 30 db of quieting. A deviation sensitivity test is the measurement that characterizes the discriminator. This measurement is generally accomplished by having a specified carrier level, with minimum deviation, produce a standard test output signal at the discriminator with all controls set for maximum gain.

4. *Receiver AGC Characteristics* — Generally the telemetering receiver detects fluctuating RF signals from the telemetering transmitter due to the relative motion between the receiver and transmitter. Because of the motion problem, good AGC response and control are necessary to insure reliable reception. The fact that the 202-G has a continuously calibrated RF signal output which can be continuously varied, makes this instrument particularly suited for measuring the AGC level and response. The AGC level is determined by feeding a predetermined RF level into the receiver and measuring the amount of dc level in the AGC circuit. Normally, AGC time response is checked using a low-frequency square wave to amplitude modulate an RF source of a predetermined level and observing the rise time of the AGC voltage on an oscillograph or oscilloscope.

5. *Recovered Audio Distortion* — The

signal generator system may be used to measure distortion on recovered audio in TCM (tone code modulation), PM (phase modulation), and other overall systems. This determination would normally be made with a distortion analyzer, such as the H-P 330B, connected to the output of the receiver under test.

6. *Pulse Response Check* — Occasionally in PAM (pulse amplitude modulation), PCM (pulse code modulation), and PDM (pulse deviation modulation) types of modulation it is important to know the overall pulse response of a receiver. If excessive delay, rise time, fall time, and overshoot occurs with the recovered data, the overall data handling capability of a system is adversely affected. The pulse response check is accomplished by using a square-wave generator to modulate the 202-G and observing the output pulse on a good oscilloscope.

7. *System Performance Check* — By modulating the 202-G with an external subcarrier generator, the overall performance of a complete data system may be checked. A check of the system can also be made to determine system reliability versus signal and modulation level. The modulating signal may also be recorded and compared with the recovered data.

MEET OUR REPRESENTATIVES

Instrument Associates, Inc.

The main sales offices of Instrument Associates, Inc. are situated in Arlington, Massachusetts, in close proximity to Massachusetts' famous "electronics highway" (Route 128). Organized in 1955 by James F. McCann the company has continued to expand its operations to keep pace with the growth of the electronics industry in the New England area. In 1956, to meet the ever-increasing demands for competent applications engineering assistance, Instrument Associates, Inc. established a second sales office in Hartford, Connecticut. Completely modern, well-designed, and well-staffed facilities in both locations include general sales offices, sales order departments, display and demonstration areas, seminar rooms, and fully-equipped engineering service laboratories. Customer applications engineering service is available from factory-trained field engineers who serve the states of Maine, Vermont, New Hampshire, Massachusetts, Rhode Island, and Connecticut.

Instrument Associates, Inc. Headquarters in Arlington, Mass.

Important past experiences in various positions have aided "Jim" in evolving his company's concept of "sales through service". After securing his BSEE degree from Tufts University, he pursued graduate studies in Business Administration at Boston University. His engineering career started at the General Electric Company where he spent seven years as design and development engineer. The next two years involved government sales with the same company. A commissioned officer in the United States Navy, he served in both World War II and the Korean War.

James F. McCann

We at BRC are proud of our association with Instrument Associates, Inc. and salute them for their continuing record of faithful service to our many customers in the New England area.

275-A As A Bias Supply for Making Transistor Measurements on the RX Meter

The 275-A Transistor Test Set not only provides a highly accurate source of measurement for α_o, β_o, and h_{ib}, but can be used to furnish I_E and V_{OB} bias for measurements on the RX Meter. The power supplies in the 275-A are ideal for this purpose because they are continuously variable, well regulated, and easily reversed. In addition, both I_E and V_{OB} are monitored with an accurate meter and the voltages are readily accessible from the front panel test terminals.

When operating the 275-A in this manner, it is advisable to remove the 1000-cps signal superimposed on the E and B terminals, as this signal may cause an indefinite null on the RX Meter. The signal is easily removed by connecting a suitable capacitor; e.g., 30 μf with a minimum working voltage of 15 vdc, across the E and B test terminals, and setting the α-h_{ib}-β Selector to the .9 — 1.0 α range. The positive terminal of the capacitor must be connected to the B terminal on the 275-A when the NPN-PNP switch is in the NPN position, and to the E terminal when the NPN-PNP switch is in the PNP position.

A detailed discussion of a method for determining transistor parameters using the Transistor Test Set Type 275-A and the RX Meter Type 250-A is presented in Notebook Number 26.

ARVA, INC. NEW BRC SALES REPRESENTATIVES

Boonton Radio Corporation is pleased to announce the appointment of ARVA, Inc. as sales representatives for BRC in the states of Alaska, Washington, Oregon, Idaho, and Montana, and the Western Canadian Provinces of British Columbia, Alberta, Saskatchewan, and Manitoba. Headquarters of ARVA, Inc. is located in Seattle, Washington. Other sales offices are located in Spokane, Portland, and Vancouver, D. C. Canada.

EDITOR'S NOTE

Hewlett-Packard S. A. Appointed European Distributor for BRC Products

In order to further improve our service to our many European customers, BRC, effective July 1, 1960, appointed Hewlett-Packard S. A. Rue Du Vieux Billard No. 1 Geneva, Switzerland, exclusive sales coordinator and distributor for our products in the following countries: Austria, Belgium, Denmark, Finland, France, Western Germany, Greece, Italy, Netherlands, Norway, Portugal, Spain, Sweden, Switzerland, Yugoslavia, and United Kingdom. Under their direction, qualified Engineering representatives have been established in each of these countries to provide local application engineering, order processing, and repair services.

BRC products are now available for immediate delivery from HPSA's "duty-free" warehouse in Basel, Switzerland with one minimum uniform surcharge added to the U. S. catalog price to cover shipping and handling costs. Bulk air shipments to the Basel warehouse are made periodically from our plant in Boonton, New Jersey thereby eliminating expensive export packing charges and, by consolidation, actually reducing the shipping and handling charges on individual items.


Bill Doolittle, Managing Director HPSA

For more than a decade, BRC has intensively trained its domestic Sales Representatives through frequent Sales Seminars at our factory in Boonton, New Jersey but, because of the long distances involved, it has not always been possible for a majority of our European representatives to regularly participate in this program. Under our new arrangement with HPSA, special training seminars are now scheduled in Geneva, Switzerland, exclusively for our European staff. Representatives from our factory will make available to all of our European customers the very latest application engineering data on both new and established products. The first of these seminars will be held in April, at which time BRC will be represented by Harry J. Lang, Sales Manager.

Import control regulations in many European countries have made it difficult or even impossible for potential BRC customers to receive demonstrations of BRC instruments. To overcome this problem, HPSA has equipped a specially constructed Mobile Demonstration Laboratory, outfitted with many of the newest BRC instruments, to make on-the-spot demonstrations in any part of Europe.

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NUMBER 28

The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

The BRC UHF Q Meter
A New and Versatile Tool for Industry

CHARLES W. QUINN, *Sales Engineer*

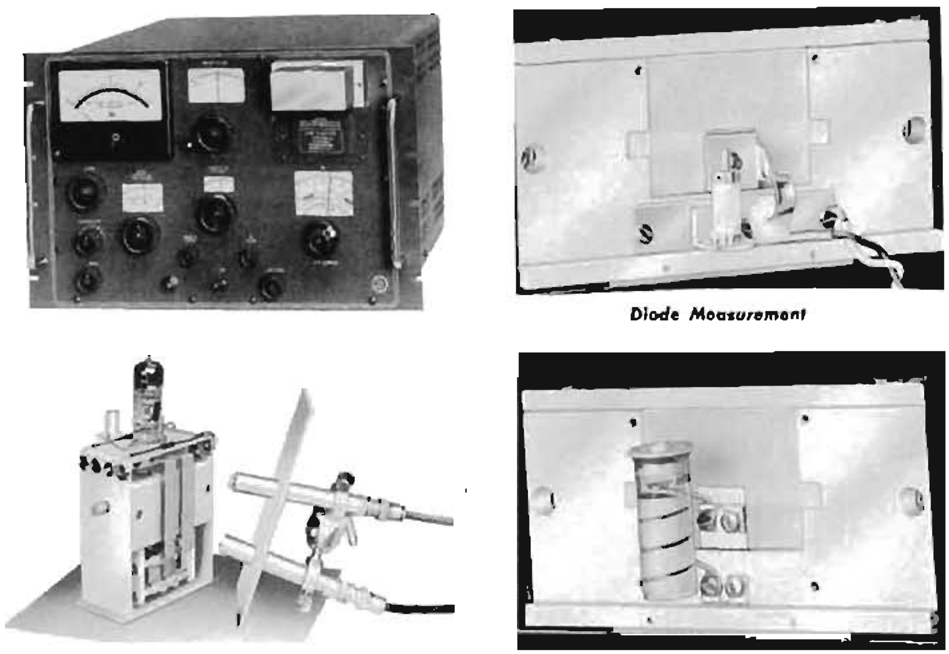
Q Meters have been serving the electronic industry for more than 25 years. Their original application was in the design of resonant circuits, in the early days of radio-frequency communication and broadcasting. Since that time, Q Meter applications have multiplied many times.[1-5] The basic theory of Q Meter operation, however, had not changed in all these years, until the development of the new Type 280-A UHF Q Meter.[6] With this change in Q meter theory, the applications will be again multiplied. It is these applications which are the subject of this article. Conventional measurements, as well as unconventional measurements, which include measurements of external resonators and components, and "in circuit" Q measurements, will be described.

PURPOSE

The prime purpose of the UHF Q Meter is to provide industry with a versatile impedance measuring device that will extend Q Meter measurements into the UHF region. The UHF Q Meter is a completely self-contained instrument capable of measuring, rapidly, conveniently, and directly; Q, capacitance, and inductance. Until the advent of the UHF Q Meter, a signal generator, a frequency measuring device, a dc amplifier, and coupling devices were required to make these tedious measurements. Inductance and capacitance, which are now measured directly on the calibrated capacitor, could not even be measured with the above-mentioned equipment.

OPERATING PRINCIPLE

To aid the reader in understanding the theory of the Type 280-A UHF Q Meter it might be well, at this point, to compare its operation with the lower frequency Q Meters, Types 260-A and 190-A. This comparison is especially necessary if use of the instrument, beyond the obvious, is to be understood.



Diode Measurement

"In-circuit" Q Measurement

Coil Measurement

Figure 1. Typical Applications of the UHF Q Meter

Preivous Q Meters utilized the definition that:

$$Q = \frac{X_{LS}}{R_S} = \frac{R_P}{X_{LP}},^*$$

*S and P subscripts indicate series and parallel configurations respectively.

as well as the fact that the voltage (V_C), measured across C (the Q capacitor), has the following relationship at resonance:

$$V_C = QV_S, \text{ or } Q = \frac{V_C{}^*}{V_S}$$

*Within the Q Meter Q limits (10 to 625).

V_C is the voltage injected in series with the resonant circuit (Figure 2A). If V_S is held constant, then Q is directly proportional to V_C. This basic principle, employed in all BRC Q Meters to date, is known as the "resonant rise" system of making Q measurements.

YOU WILL FIND . . .

ESTIMATE THE Q WIN A Q METER

Yes, that is all that is necessary to win the factory reconditioned Type 160-A Q Meter which will be on display in the BRC exhibit at the IRE show to be held in the New York Coliseum from March 20th through March 23rd. The Q Meter will be awarded to the person whose estimate is closest to the actual measured Q of the resonator circuit to be displayed with the Q Meter. Complete information will be furnished by engineering personnel on duty in BRC Booths 3101 and 3102.

THE BRC NOTEBOOK is published four times a year by the Boonton Radio Corporation. It is mailed free of charge to scientists, engineers and other interested persons in the communications and electronics fields. The contents may be reprinted only with written permission from the editor. Your comments and suggestions are welcome, and should be addressed to: Editor, THE BRC NOTEBOOK, Boonton Radio Corporation, Boonton, N. J.

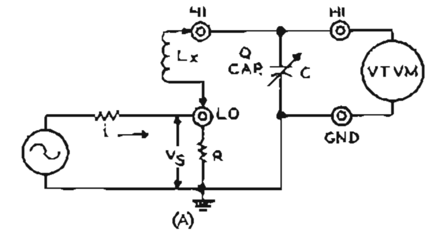
The UHF Q Meter uses the peak of the resonant rise to indicate resonance, but employs the bandwidth relationship to determine Q, where:

$$Q = \frac{f_r}{\Delta f} \qquad (1)$$

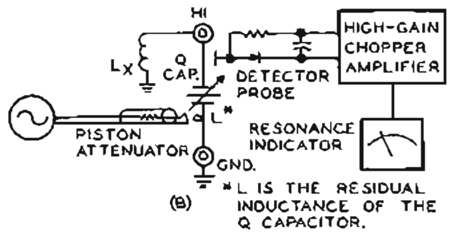
This relationship is shown in Figure 3. Δf is the frequency between the two 0.707 voltage or half-power points, and f_r is the frequency at the resonant peak. As is indicated in Figures 2B and 4, there are other more subtle differences between the UHF Q Meter and the lower frequency Q Meters. These will be discussed later in this article.

FIELDS OF APPLICATION

Because of its frequency range, the UHF Q Meter will serve many fields of the electronic industry. Some examples of these fields ar given below.



Simplified Circuit — Conventional Q Meter (Type 260-A)



Simplified Circuit — UHF Q Meter Type 280-A

Figure 2. Comparison of Q Measuring Circuit in Conventional Q Meters and the UHF Q Meter

means of the appropriate controls.

3. The Q capacitor is adjusted until output is indicated on the resonance indicating meter.

4. The Q capacitor or Q (frequency) control is adjusted in conjunction with the Level Set control until the resonant peak is indicated at full scale on the meter.

FIELD	SPECIFIC APPLICATIONS
Missile and Rocketry	Telemetry and remote control systems.
Communications	Commercial, mobile airborne, relay networks, amateur radio, UHF television, and military mobile.
Navigational Aids	Glide slope
Radar and ECM	
Components and Materials Manufacturers	Inductors, cores, capacitors, UHF diodes, insulators, and resistors.
Other Fields	Accelerator, medical research, and basic research of new materials.

BASIC OR CONVENTIONAL MEASUREMENTS

Set-up Procedure

A condensed set-up procedure will be given at this point to aid in the understanding of the instrument. The same procedure is used for both conventional and unconventional measurements. Arbitrarily, it will be assumed that the Q and inductance of a small inductor is going to be measured.

1. The component to be measured is clamped to the Q capacitor terminals by means of the clamps provided (Figures 1 and 5), or by other suitable means.

2. The oscillator is adjusted to provide the desired operating frequency by means of the appropriate controls.

5. The appropriate Q dial is locked and its knob is turned clockwise to the proper half-power point which is indicated by the Q mark on the meter.

6. The Q dial is unlocked and the knob is rotated in a counter-clockwise direction, through the resonant peak, to the opposite half-power point; also indicated by the Q mark on the meter.

7. Q is read directly on the appropriate Q dial, capacitance is read directly on the Q capacitor dial, and inductance is read directly on the integral calculator dial.

Inductance Measurements

Inductance measurements are a primary function of all Q Meters. The UHF Q Meter capacitance dial is provided with a spiral calculator to compute inductance from the capacitance reading and the operating frequency. The direct-reading inductance range is 2.5 to 146 millimicrohenries (Figure 6). Circuit Q is read directly from the CIRCUIT Q dial.

Capacitance Measurements

Capacitance measurements are second nature to a Q Meter, but are indirect measurements in that a reference inductor or "work coil" must be used. The clamps provided with the instrument permit individual connection of the work coil and the unknown capacitor for parallel measurements. Standard Q Meter procedure is then employed to make the parallel capacitance measurements and all general Q Meter equations [2,7] apply. Q_1 and C_1 of the work coil are measured; then, with the unknown capacitor (C_X) connected, Q_2 and C_2 are also measured. The capacitance of the specimen is determined by the equation:

$$C_X = C_1 - C_2$$

and

$$Q_X = \frac{Q_1 Q_2}{Q_1 - Q_2} \times \frac{C_X}{C_1}$$

Dissipation factor measurements can be estimated by referring to Figure 7. For example, a 20-pf capacitor with an R_p of 0.3 meg. ohms can be detected at 210 Mc, using a work coil with a Q_1 of 300. The dissipation factor would then be computed as follows:

$$D = \frac{1}{Q} = \frac{X_C}{R_P} = \frac{40}{0.3 \times 10^6}$$
$$= 130 \times 10^{-6}$$
$$= 0.00013$$

Consider the possibilities if higher Q inductors or resonators are used. One precaution must be observed if a false value for C_2 is to be avoided. The Q dials (frequency dials) should always be returned to their original positions, indicated by the resonant peak of the work coil before C_X was connected.

Direct parallel capacitance measurements, over a range of 0.1 to 20 pf are possible on the UHF Q Meter. It is also possible that capacitance measurements can be extended by a "step-shunt" technique. This technique requires that an external capacitor or capacitors (C_A and C_B), within the capacitance range of the instrument, be calibrated at the frequency of measurement. The external capacitors are then connected in parallel with another work coil and the Type 280-A internal capacitor is adjusted to peak. The external capacitors are removed as required when the unknown capacitor (C_X) is connected. Then:

$$C_X = C_A + C_B + (C_1 - C_2) \quad (2)$$

Series techniques may also be used. Some suggestions on this subject are taken up in the resistance measurement section which follows.

Resistance Measurements

Resistance measurements are also indirect measurements, and the procedure used is identical to that used for capacitance measurements. In this case, however, we are interested in the major parameter of resistance. Figure 8 shows approximate limits of measurable resistance for indicated Q_1 values of 300 and 500, Q_2 values of 20 and 10, and a $\triangle Q$ of 10. Approximate limits for both parallel and series measurements are given. The upper limits of parallel measurements may be extended by utilizing higher Q reference inductors and smaller $\triangle Q$ values. The lower limits of parallel measurements may be extended, slightly, by using additional external capacitance.

At ultra-high frequencies, series measurements present a more difficult problem. First, shunt capacitance and series inductance of the series jig must be small relative to the resistance to be measured. Secondly, a low inductance and low resistance short-circuiting device must be employed.

In the Type 280-A, circuit component contact resistance is basically the lower limiting factor in series measurements. This contact resistance usually becomes a function of the component shape and may require a special machined fixture for a given component.

A short cut to solving the multiple computations of the real component of parallel impedance measurements is illustrated in Figure 7. Curves for a given work coil, with Q_1 and frequency held constant, are plotted as a function of Q_2 and R_{P}. If the work coil is stable, well designed, rigid, well placed, etc., these curves, or a group of curves, can be used for general measurements over long periods of time.

SPECIAL OR NONCONVENTIONAL MEASUREMENTS

The basic parameters of L, C, and Q are often affected when brought near, or in contact with, a component to be tested. Let us consider some specific instances and determine what measurements may be made.

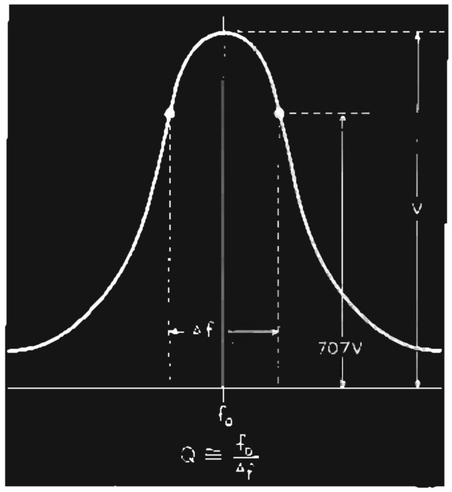


Figure 3. Q Resonance Curve

Measurements Involving Change in Inductance and Resistance

Iron cores, shells, toroids, and rods may now be tested simply, at higher frequencies, with the UHF Q Meter. It has been found that some defects are detectable in the resistive or Q_2 indication at these frequencies (210 to 610 Mc) that do not show up at the lower operating frequencies.

The ferro-resonant frequency of some ferro-magnetic components may be detected on the resonance indicating meter, if this resonance falls within the instrument frequency range.

Figure 9 suggests a possible jig design for coupling these and other components, liquids, and materials into the inductive field of a test coil. The plastic plug can be machined to precisely position the specimen so that the change in C, L, or Q falls within the range of the instrument. A change of inductance indicates a change in effective permeability and a change in Q indicates a change of specimen resistivity. A high degree of precision can be achieved in these measurements, since both the work coil and plug can be fabricated on precision machines.

A work coil and two plastic plugs, patterned after those shown in Figure 9, were made and attached to the Q capacitor terminals on the UHF Q Meter, and a few experiments were performed which produced some interesting results. In the first experiment, a group of small shell cores were inserted in the plastic plug and tested at 400 Mc. Q_1 was determined to be within 5% of 630, and Q^2 was within 5% of 284 for all cores. Inductance increased by 5%, indicating permeability greater than unity, even at 400 Mc. Core #4 showed a 5% decrease in inductance, with a drop to 135 in Q_2. This core was obviously of low-frequency material acting like a poor short circuit. This experiment indicates a technique for evaluating inductive tuning or adjustment devices and their effects upon circuit Q at ultra-high frequencies.

The author has long been curious about the effects of liquids on circuit Q. This curiosity led to the second experiment, performed to determine the effect of tap water on circuit Q, with and without a few salt crystals added. Q_2 measured for the clear water was 610. Low losses, very little change in inductance, and approximately 1% increase in distributed capacitance were noted. A pinch of salt (NaCl) was then added and the effects noted. Q_2 dropped to 255, with no inductance change apparent. It can be concluded

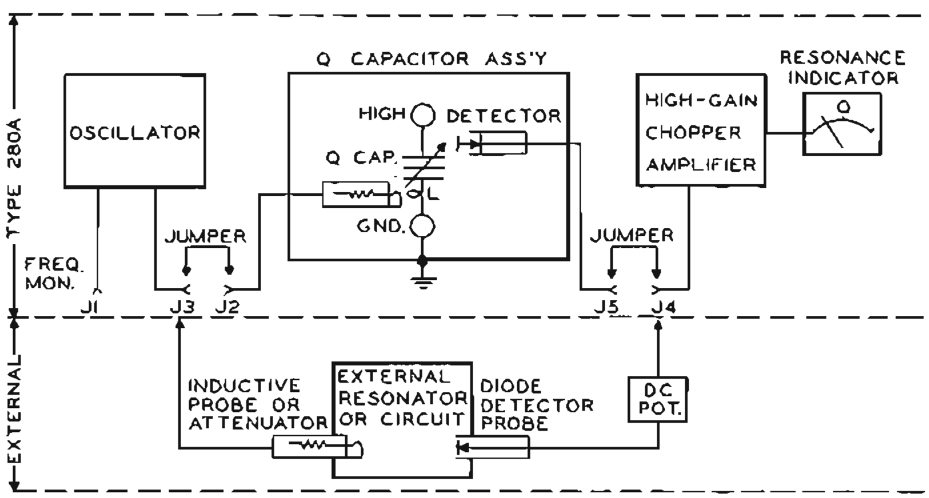


Figure 4. Block Diagram of UHF Q Meter Showing External Resonator Connections

then, that the RF resistivity or losses only change in a positive direction with the addition of salt.

These experiments point up the application of the UHF Q Meter in the UHF inductive heating field (cooking of foods, curing of adhesives and resins, etc.) where it is important to know the frequency of optimum energy absorption.

Jigs similar in theory to the one discussed above, but more sophisticated, may be constructed to detect, test, and measure more complex components and materials and to solve more exacting problems. For example, a capacitance-loaded or "end-tuned" coaxial resonator could be adapted to check toroidal behavior under truly inductive conditions and with the flux lines in a specific plane.

Measurements Involving Change in Capacitance and Resistance

The measurement of the dielectric loss factor of Teflon, Polyethlene, etc., is one of the most difficult measurements to make with any degree of accuracy. For example, high-grade Teflon is known to have a loss factor of approximately 0.00014.

The Type 280-A UHF Q Meter, with its frequency range of 210 to 610 Mc and Q range of 10 to 25,000, makes this equivalent high shunt resistance more readable. Further, since the Type 280-A employs a bandwidth measuring system; i.e., Δf is measured between the half-power points, permitting the use of frequency counting techniques; calibration and readability of the Q dials can be eliminated as a source of error and ΔQ becomes more readable, limited only by our ability to measure Δf. Let us consider the order of Δf or frequency changes that will be encountered for such a measurement.

Conditions:

1. The specimen capacitance (C_x) should be about 10pf.
2. If a plate area of 0.6 inches is used, material thickness should be 1/32 inch for approximately 10 pf C_x.
3. C_1, under these conditions, should be approximately 15pf.
4. Q_1 should be at least 500.
5. Operating frequency is 300Mc.

Solving for ΔQ: We can now solve for the expected ΔQ for a 0.0001 dissipation factor. In this case:

$$D = \frac{1}{Q_x}, \; Q_x = 10{,}000.$$

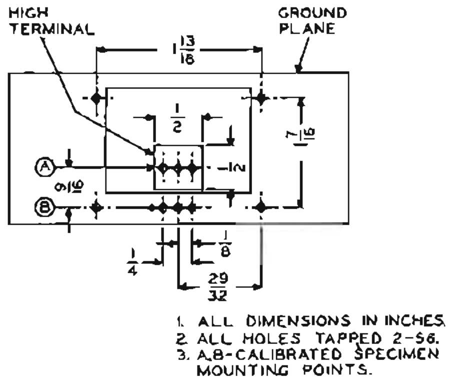


Figure 5. Q Capacitor Terminal Dimensions

Using the standard equation for Q:

$$Q_x = \frac{Q_1 Q_2}{\Delta Q} \times \frac{C_x}{C_1}. \qquad (3)$$

Let $C_x/C_1 = K = 0.66$ which is a practical ratio adjustable by manipulation of inducance or frequency and specimen thickness. Then:

$$Q_x = K \frac{Q_1 Q_2}{\Delta Q} \qquad (4)$$

since

$$Q_2 = Q_1 - \Delta Q; \qquad (5)$$

and

$$\Delta Q = \frac{K Q_1^2}{Q_x + K Q_1}. \qquad (6)$$

Example (for above conditions):

$$\Delta Q = \frac{.66\,(500)^2}{10{,}000 + .66 \times 500} = \frac{.66\,(25 \times 10^4)}{10{,}330} = 16$$

Calibrated dial divisions at this Q

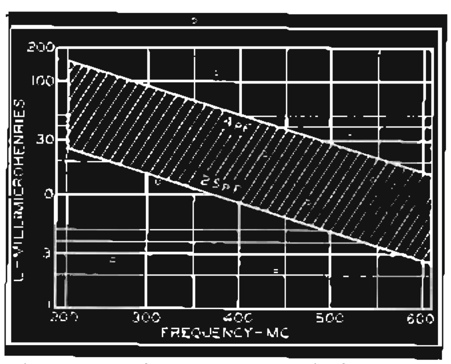


Figure 6. Inductance Range of the UHF Q Meter (Direct Reading)

value are 10 units. This means that estimates from the dial reading can be within approximately 20% of this ΔQ value. With a ΔQ of 16 at a frequency (f_r) of 300 Mc, what is the frequency bandwidth change? Let us refer to this change as Δf_3. The derivation of the equation used is as follows:

$$Q_1 = \frac{f_r}{\Delta f_1}, \; \Delta f_1 = \frac{f_r}{Q_1} \qquad (7)$$

$$Q_2 = \frac{f_r}{\Delta f_2}, \; \Delta f_2 = \frac{f_r}{Q_2} \qquad (8)$$

$$\Delta Q = Q_1 - Q_2, \; Q_1 > Q_2$$

$$\Delta f_3 = \Delta f_2 - \Delta f_1 \qquad (9)$$

$$= \frac{f_r}{Q_2} - \frac{f_r}{Q_1}$$

Clearing:

$$\Delta f_3 = \frac{Q_1 f_r - Q_2 f_r}{Q_1 Q_2} = \frac{f_r\,(Q_1 - Q_2)}{Q_1 Q_2} = \frac{f_r\,(\Delta Q)}{Q_1 Q_2} \qquad (10)$$

To compute the above example:

$$\Delta f_3 = \frac{300 \text{ Mc} \times 16}{500 \times 484}$$

$$= .0198 \text{ Mc or } 19.838 \text{ kc}$$

From the above example, two factors stand out as important to the accuracy of measurement: First, the value of the ratio K in equation 4, especially if the Q dial readout is to be used, should approach as close to unity as possible to optimize readability. Secondly, equations 7, 8, 9, and 10 indicate that a frequency measurement technique can be used to measure Q_1, Q_2, and ΔQ.

Use of an Auxiliary Frequency Counter to Measure Loss Factor

Fortunately for those with dielectric loss measurement problems, the art of frequency measurement is highly refined and is really a simple solution to the loss-factor measurement problem. A popular frequency measuring device found in most laboratories is the frequency counter. This instrument, with a suitable transfer oscillator, has more than sufficient accuracy and resolution for this application. The frequency counter is connected to jack J1 at the

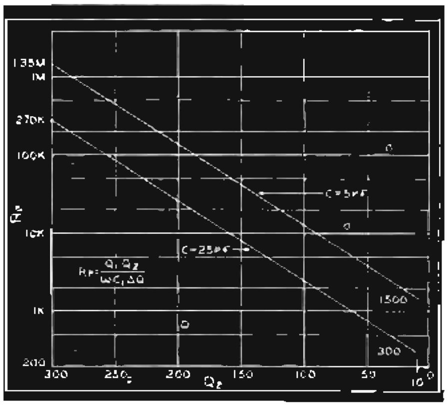


Figure 7. R_p versus Q_T and C_X

rear of the Type 280A (Figure 4) which is provided especially for monitoring the UHF Q Meter frequency. With this technique the accuracy of measurement is determined by the short-term frequency stability of the Type 280-A and the stability of the half-power indicator in its most insensitive position (position of maximum stability). In this mode of operation, 0.5 kc per minute per 100 megacycles can be resolved with good repeatability. For this method of dielectric measurements, it is convenient to derive the equations for Q_x and D_x (dissipation factor) in terms of frequency. Considering the relationships of equations 7 and 8, equation 3 can be written:

$$Q_x = \left[\frac{\dfrac{f_r}{\Delta f_1} \times \dfrac{f_r}{\Delta f_2}}{\dfrac{f_r}{\Delta f_1} \times \dfrac{f_r}{\Delta f_2}}\right] \times \frac{C_x}{C_1}$$

$$O_x = \frac{f_r^2}{\Delta f_1 \times \Delta f_2} \times \frac{\Delta f_1 \times f_2}{\Delta f_2 \times f_r - \Delta f_1 \times f_r} \times \frac{C_x}{C_1}$$

$$= \frac{f_r^2}{f_r\,(\Delta f_2 - \Delta f_1)} \times \frac{C_x}{C_1}$$

$$= \frac{f_r}{(\Delta f_2 - \Delta f_1)} \times \frac{C_x}{C_1} \qquad (11)$$

$$D_x = \frac{\Delta f_2 - \Delta f_1}{f_r} \times \frac{C_1}{C_x} \qquad (12)$$

Dielectric loss factor measurements in this range, were heretofore obtained by refined techniques and extreme skill. The Type 280-A Δf technique can achieve $\pm 10\%$ accuracy (or one part in the fifth place) with considerable simplification of the measurement procedure in this frequency range.

Measurement of Semiconductor Components and Materials

Since one of the key features of the new UHF Q Meter is high detector gain, low RF levels are available across the component to be tested. The level can be selected by the front panel SENSITIVITY control from 25 to 250 millivolts. Of the many components measurable in this RF voltage range, the variable-capacitor diode is one of the best examples. Here, one is most concerned with the behavior of Q and capacitance as a function of bias and frequency. With 0.025 volts RF across the diode, investigations to almost zero bias (0.1v dc) can be made. RF impedance of detector and mixer diodes can be determined using standard Q Meter equations[7]. A suggested design for a diode jig, with provisions for biasing, is shown in Figure 1. Other parametric and nonlinear components, including h_{ie}, h_{oe}, and h_{ob} of some UHF transistors, may be measured in a similar manner. Semiconductor material resistivity can be measured in the electrostatic manner previously described under "Measurements Involving Change in Capacitance and Resistance", or relative resistivity can be obtained using the inductive jig previously described under "Measurements Involving Change in Inductance and Resistance."

External Resonator and "In Circuit" Measurements

One of the most interesting phases of the new UHF Q Meter application is the measurement of external resonators and "in circuit" measurements. Referring to Figure 2B and 4, observe that there is really no direct connection to the injection and detection circuits. The RF signal is actually magnetically coupled or induced into the Q capacitor by a piston-type inductive attenuator. This device is a tubular probe, with a single turn of wire at its end. The detector circuit is similar to a conventional diode probe used on many RF vacuum tube voltmeters and is coupled to the Q capacitor by merely bringing one end of it near the electrostatic field of the stator structure.

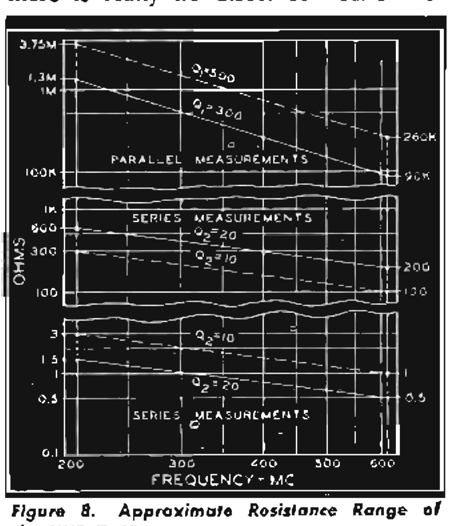


Figure 8. Approximate Resistance Range of the UHF Q Meter

The fact that there is actually no conductive connection to the circuit under test suggests many possible configurations for making measurements. As shown in Figure 4, connections to the Q capacitor assembly have been made through a series of jacks and jumpers located at the rear of the instrument. This means that the oscillator and high-gain amplifiers may be disconnected from the Q capacitor.

External Resonators

First, let us assure that we have a coaxial resonator and need to know its Q and resonant frequency. Due to the physical size of the component, it can not be mounted on the Q capacitor terminals. Even if it could be mounted, the minimum capacitance of 4pf would prohibit uncorrected measurements. The Type 280-A, with appropriate accessories, can make these measurements on the bench rather than on the instrument. Figure 4 shows the connections for a typical resonator circuit. The piston attenuator and diode probes shown in Figures 1 and 4 will be made available as optional accessories for the Type 280-A.

The procedure for making this measurement is basically the same as for making conventional measurements, except that the "Level Set" controls (Q capacitor piston attenuator and Q capacitor controls) are no longer operative. The motion of the attenuator probe and adjustment of the dc potentiometer serve as the "Level Set" control once the detector probe has been positioned. The frequency or CIRCUIT Q dials are then tuned to obtain the resonant peak. The resonant frequency is read directly on the frequency dial, or by means of external frequency measuring equipment if desired. The Q measuring procedure is the same as described above for inductors.

Care must be taken to avoid unexpected loading of the resonator. Prevention of this loading is one function of the coupling block and is also the reason that an adjustment is provided on the attenuator probe. Two Q readings, at different detector probe and attenuator probe settings, will establish the extent of loading. If there is any loading, Q_2 will be different than Q_1.

A plot of two or three Q readings as a function of coupling will show that Q approaches a limit, asymptotic to the Q value, at which the Type 280-A injection and detection circuit reflected losses are negligible. This Q value is the actual unloaded Q of the resonator under test.

In resonators of this type, Q is important as a method of determining bandwidth in receivers. The effects of circuit loading can be determined and optimized.

As a power handling device, Q is related to efficiency (E) as follows:

$$E = 100 \left(1 - \frac{Q_L}{Q_{UL}}\right) \%, \quad (13)$$

where Q_L = Q loaded and Q_{UL} = Q unloaded.

"In Circuit" Measurements

A distinct advantage of the UHF Q Meter is its ability to measure the Q of resonant circuits (resonators) as they are connected and mounted in actual use; i.e., "in-circuit" measurements. This is extremely important, since the behavior of most resonators is a function of many things. Resonators may take many forms; i.e., coaxial, cavity, open lines, strip lines, butterfly tanks, etc. An example of a typical "in-circuit" measurement problem is shown in Figure 1. Here, flat strips are used to form a resonator for a developmental RF amplifier. It is important to know the Q_L and Q_{UL} of the resonator to determine the optimum efficiency versus bandwidth compromise. Coupling was achieved as illustrated, and the following example readings were made at 400 Mc: Q_{UL} = 400, Q_L = 40, E = 100 (1 - 40/400) = 90%. It was found that due to radiation losses, Q_{UL} dropped to 300 with the shield removed, resulting in an efficiency of 100 (1 - 40/300) % = 84%. These efficiencies were adequate, but a different tube type and aluminum shields resulted in a Q_{UL} of 100. Efficiency was 60% under these conditions and, therefore, this may prove to be an unusuable configuration.

An extension of this type of measurement can be applied to mating components, or may be used to determine

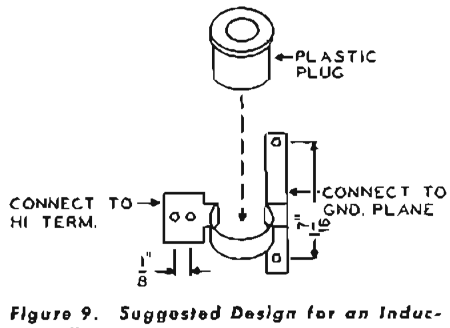


Figure 9. Suggested Design for an Inductance Jig

Q at the self-resonant frequency of an inductor. The components are placed on a small ground plane in the vicinity of the probes, or in a convenient shield, to limit radiation losses and body capacitance effects. By this means, any tuning or fixed capacitor desired may be employed.

It is important to realize that measurements made in the manner described in this section yield Q_e; i.e., the effective Q of the component and associated circuit inperceptably influenced by the Q Meter, if care is used to determine sufficient probe decoupling. This is the actual "in-circuit" Q and can be used directly in circuit computations. The Type 280-A UHF Q Meter is the only Q Meter in existance that can measure, directly, the Q of a circuit that is resonant at the frequency of measurement.

To measure circuit "stray" capacitance, a coil may be calibrated on the Q capacitor and then soldered into the circuit at the desired points. The circuit capacitance can then be computed from the relationship for resonance:

$$f = \frac{1}{\omega\sqrt{LC}}$$

or

$$C = \frac{1}{f^2 \omega^2 L} \quad (14)$$

The same technique can be applied to circuit inductances.

CONCLUSION

We have attempted to describe some of the applications of the new UHF Q Meter Type 280-A, but realize that there will be many more jobs for this versatile instrument; some of which are not apparent at this writing. These will provide worthwhile subject matter for future articles in The Notebook.

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SERVICE NOTE

RX Meter Null Indicator

Proper operation of the Type 250-A RX Meter is dependent upon the correct balancing of the bridge circuit, and the bridge circuit cannot be correctly balanced if the NULL INDICATOR is not functioning properly. To check the operation of the NULL INDICATOR, proceed as follows:

1. Select the desired measuring frequency by means of the OSC RANGE and OSC FREQ controls.
2. Set the C_p dial to "O" and the R_p dial to ∞.
3. Unbalance the bridge by shorting the two binding posts and adjusting the DETECTOR TUNING knob until maximum deflection is obtained on the NULL INDICATOR. The meter pointer should indicate about 35 scale divisions. A peak of substantially less than this amount is usually an indication of an unusuable harmonic response instead of the desired fundamental. At higher frequencies, two fundamental frequency peaks will be observed, either of which represents satisfactory tuning of the detector. Several secondary or harmonic peaks, which may be recognized by their relative sharpness and low amplitude, will be observed between the fundamental peaks. Care should be taken not to tune to one of these harmonics, since this will produce erroneous readings or make bridge balance impossible. When maximum meter deflection has been obtained, remove the short from across the binding posts and tighten the binding posts nuts.
4. Balance the bridge by adjusting the three ZERO BALANCE controls, alternately, until a minimum deflection is obtained on the

NULL INDICATOR. The indication should not be more than 3 scale divisions on the meter. At frequencies above 100 Mc, the COARSE R control should be adjusted to its approximate midpoint position before null is sought. Since a slight interaction exists, at high frequencies, between the FINE R and C controls, it is important to use all three controls to obtain final balance. When an apparent null has been obtained, the circuit should be tested for true balance by slowly rocking the R_p dial above and below the setting, and observing the NULL INDICATOR. If a deeper null is observed at some R_p value other than ∞, the R_p dial should be returned to the latter indication and a new balance obtained with the ZERO BALANCE controls.

NOTE: When the measurement frequency is changed, steps 2 through 4 above should be repeated.

5. After the bridge is balanced as described above, set the frequency controls for 0.5 megacycles and change the R_p dial setting from ∞ to 100K. The NULL INDICATOR pointer should deflect upscale and indicate approximately 7 to 12 divisions.

A 10–500 Mc Signal Generator Power Amplifier

ROBERT POIRIER, *Development Engineer*

An increasing demand has developed for higher RF power output levels, in the 0 to 10 dbw maximum output range, over the frequency range from 10 to 500 Mc, for the testing of communications systems and for general laboratory measurements. The need for higher power output signal sources results mainly from strong signal and cross modulation requirements of certain receiver tests and the large input signal requirements of bridge type devices. Because of the large number of existing signal generators in the 0 dbm maximum output category, BRC has developed a tunable signal generator power amplifier for use with these instruments. The signal generator power amplifier is to be an accessory for use with any signal generator having a maximum output in the vicinity of 0 dbm to provide a maximum output level in the vicinity of 4 dbw.

The new Signal Generator Power Amplifier Type 230-A conceived by the Boonton Radio Corporation, consists essentially of three tracked tuned, cascaded stages of grounded-grid amplification. The choice of grounded-grid triode amplification was established primarily by a desire to provide a maximum operating frequency of 500 Mc. Two other advantages which are accrued for grounded-grid triode amplification as compared with grounded cathode tetrodes are: a low untuned input impedance which can be made nominally in the vicinity of 50 ohms, and a gain

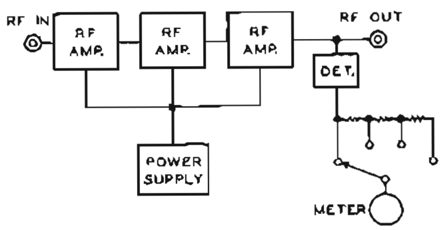


Figure 1. Block Diagram of Signal Generator Power Amplifier Type 230-A

and maximum power output which are less sensitive to variations in load impedance. A minimum of 34 db power gain is to be provided for a frequency coverage of 10-500 Mc which will be continuously tuned in six slightly overlapping ranges. The gain will be linear within 9.0% up to 10 volts output in a 50-ohm termination. This provides that a maximum of 91% AM of a 5-volt carrier level, with 10% distortion of the modulation envelope, will be obtained for a 100% modulated (with no envelope distortion) input signal for which the carrier level approaches 0.1 volt or —7 dbm. The changes in percentage of modulation and envelope distortion which may be developed in the Signal Generator Power Amplifier at the maximum output levels, become negligible for modulation crests of 0.5 watt (5.0 volts rms in 50 ohms) or less. The linearity characteristic of the Signal Generator Power Amplifier is such that, in general, if the outgoing modulation crests exceed 0.5 watt, the modulation index will always be less than the incoming modulation by an amount not exceeding 9.0% of the incoming modulation. Whether, and in which direction, the envelope distortion may be affected at the maximum output levels, depends on the magnitude and phase of the incoming envelope distortion components, if any. The effect should be within ±10% for modulation crests of 10 volts rms in 50 ohms, diminishing to 2% or less for modulation crests of 5 volts rms in 50 ohms or less. The absolute maximum power output over most of the frequency range is 4 watts or 6 dbw (14.14 volts rms in 50 ohms), but the linearity (and gain) is not specified beyond 2 watts or 3 dbw. The overall bandwidth of the three-stage power amplifier is not less than 700 kc and is considerably greater over much of the frequency range.

A block diagram, Figure 1, shows that a self-contained power supply and an output RF voltmeter are included with the Signal Generator Power Amplifier. The RF output voltage is metered from 0-15 volts in four convenient ranges. The detector and the metering circuit will withstand the high voltages which can be developed at the RF output jack when it is unterminated, or terminated in a load having a very high VSWR. The accuracy of the RF output voltage indication is specified at the output jack to be ± 1.0 db of full scale over a frequency range of 10 to 250 Mc and ±1.5 db from 250 Mc to 500 Mc for a 50-ohm termination having a VSWR of 1.0 (0 db) in each case.

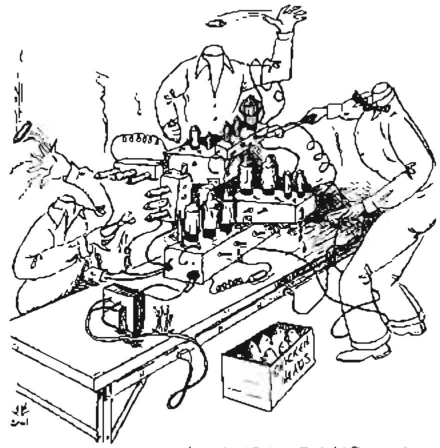
An electronically-regulated power supply is incorporated in the Signal Generator Power Amplifier to maintain a constant final amplifier plate voltage against the large variations in final plate current which occur over the range of 0.5 to 4 watts RF output. Other features include 50 ohms input and output impedance with a VSWR of 2.0:1, or less, over the frequency range of 10-500 Mc. RF leakage is sufficiently low to permit measurements at 0.1 volt.

Since the demand for higher power signal generators comes almost exclusively from sources already supplied with low-power signal generators, it is felt that the Signal Generator Power Amplifier will conveniently and readily fulfill this demand, offering up to 2 watts output for AM applications, or or up to 4 watts output for CW and FM, where amplitude linearity is unimportant.

EDITOR'S NOTE
New Look for BRC at IRE

The few weeks preceding the IRE show in March are pandemonium at BRC. Engineering and Sales are steeped in the problems of readying new instruments for showing and assuring that enough advance information is disseminated to stimulate customer interest. Many last-minute details are being attended to and the loose ends are being gathered and knotted .The last days before the show are tumultuous, but those in the midst of the turmoil are aware of the impact of the job they are doing, and in this there is solace.

This year, BRC will show its instrument line in a new display booth; designed not only to provide an attractive setting for instrument display, but to make it easier for BRC engineers in attendance to handle demonstrations and inquiries.



EACH YEAR, IN THE MONTH OF MARCH, A HIGHLY COORDINATED EFFORT IS MADE......!

Of particular interest at the show will be the UHF Q Meter Type 280-A (the subject of the lead article in this issue), the Navigation Aid Test Set Type 235-A (described in Notebook Number 24), and the new Signal Generator Power Amplifier Type 230-A (described in this issue).

Another "guess the Q" contest will be featured for those friends of BRC who welcome the challenge of a perplexing problem. Our engineers have, true to form, devised a resonant circuit which will be on display at the BRC booth. Contestants will be asked to estimate the Q of the circuit, enter this estimate on a contest card, and drop the entry into a special, locked receptacle. After the show, the Q of the resonant circuit will be measured on the UHF Q Meter Type 280-A, by means of the "in circuit" technique. Several measurements will be made and averaged. The entry which is closest to this average measured Q will be awarded a factory-reconditioned Q Meter Type 160-A. In case of a tie, a drawing will be held to determine the winner.

Plan to visit the IRE show at the Coliseum in New York City and stop at the BRC exhibit (Booths 3101 and 3102). Our engineering personnel on duty will be grateful for the opportunity to help you with your test equipment problems.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

CALIBRATION OF A UHF Q METER[1]

CHARLES G. GORSS, *Development Engineer*

Introduction

This paper describes the development of coaxial line impedance standards for the UHF Q Meter Type 280-A, a modified two-terminal Q measuring instrument. (These standards are currently being readied for production by BRC and will be available to customers in the near future.) Improved methods for machining pure copper are described. The methods of deriving the reactance and series resistance of the coaxial are also described.

The ideal way to establish calibration of an impedance measuring device and maintain that calibration in the field is to utilize a stable, intrinsically accurate, reliable, and easily used impedance standard. If more than one of these standards exist with various known values of impedance the calibration is more exact. What is more, if these standards can be duplicated by precise methods, duplicates can be placed in the field where they are needed. The 280-A UHF Q Meter is a device which needs such standards.

There is no precise instrument which will cross check measurements made by this instrument with the required accuracy in the frequency range of the 280-A (210-610 Mc). The resonating capacitance varies between 4 and 25 pf with ±5% accuracy.

The internal resonating capacitance accuracy indicates the need for accurate inductance standards to check the actual effective resonating capacitance the internal capacitor presents to the instrument terminals. What is more, the instrument measures circuit Q so that if the internal losses of the resonating capacitor are to be evaluated, the Q of the standard inductor must be well known. In this way the losses in the internal capacitor can be unwound. The standard must therefore be an inductor whose inductance and Q are both accurately known and preferably calculable from reliable physical relationships.

Figure 1. Don Gann, BRC Lab Technician, Checks the 280-A UHF Q Meter with an Impedance Standard

Design

The most logical calculable form for an inductance standard to assume is a coaxial line shorter than $\lambda/4$ and short circuited by a perfect short circuit. Ideally, there should be no dielectric other than air, the dimensions should be precisely known, the metal completely homogeneous and of a precise conductivity, and the surface roughness should be nil compared with the skin depth.

[1] This article will appear in the 1961 IRE International Convention Record.

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WE ARE MOVING

Boonton Radio Corporation will be moving to its new plant and offices in August. Our new address and telephone number will be as follows:

Mailing Address: Boonton Radio Corporation
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Address of Plant and Offices:
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Green Pond Road
Rockaway Township, New Jersey

Telephone: OAkwood 7-6400

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Effective date of the move will be announced subsequently.

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The metal picked for this development was copper. Oxygen free, high conductivity copper was chosen for its purity and relative freedom of conductivity from the effects of cold working as well as its high conductivity; actually exceeding the conductivity of the IACS (International Annealed Copper Standard).

The standards are to be of essentially 3 basic parts: the outer conductor, the short circuit, and the inner conductor. The outer conductor is a straight cylinder into which the short circuit fits. The inner conductor fits into a hole in the short circuit. Each fit is made an interference fit. The parts are joined by shrinking the inner line in liquid nitrogen and inserting it into the short circuit. These two are then shrunk and inserted into the outer line. The result is extreme pressure and virtually a welded contact without heat or solder to add resistance. To mount this structure to the terminals of the Q Meter, an outer flange is provided. This flange is soldered into place using high temperature solder. The flange is placed ¼ inch from the end of the coax in order to allow for attachment of the removable mounting plate. This mounting plate clears the coax line by 10 thousandths of an inch and is 5 thousandths short of the end of the coax line. A 5 thousandths ridge is provided at the top for contact with the mounting surface. This assures contact of the coax line itself with the ground plane and not the brass plate. The contacts cover approximately 100° of arc. The gap between the copper line and the brass mounting plate tends to keep the currents in the copper piece. This structure and its relation to the mounting surface is shown in Figure 2.

In order to contact the hot stator, a precise hole is bored into the center of the center conductor. A solid coin silver set of spring fingers plugs into this hole. A 2-56 stud on the reverse side connects this with the high post. This is placed on the high post with a torque of 35 inch ounces.

The calculability of this standard depends to a great extent now on how well the surface and dimensions agree with theoretical assumptions. The bulk dc conductivity of this copper checks out at 101% IACS. Theoretically this should be the conductivity used in calculating resistance in the surface where the current flows. This will be true if the surface is not rough, torn, or contaminated to a depth which is small compared with skin depth. This is assured by the methods used to machine the surface.

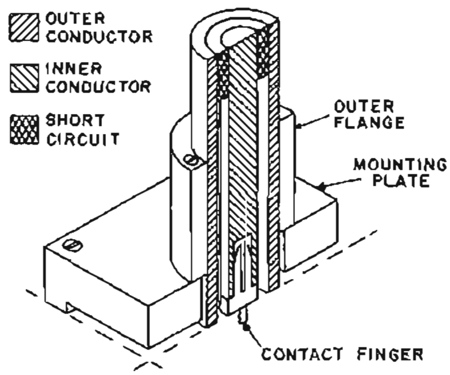


Figure 2. Cutaway View of Impedance Standard

Fabrication

In general, the proper machining of copper of this purity must be approached with a great deal of thought. Ordinary high-speed steel tools are quickly dulled by the abrasive nature of the copper to such an extent that accurate work is impossible. Silicon carbide can be used for preliminary shaping but it too is limited. All metal tools will tear the surface to a slight degree due to the tendency of the copper to stick to the tool and tear. The final cut of ½ thousandth of an inch must be cut with a diamond cutting tool. The finish obtainable from this type of tool with proper cutting rates is better than 4 microinches. The work was all done on a precision Hardinge toolroom lathe. The short circuit and the center conductor were cut in a conventional manner using the carbide and diamond tool. The outer conductor cylinder was cut out of a solid rod, first, by gun drilling within 10 thousandths. The tube was then mounted in a holder on the carriage which supported it over its full length. The boring tool was rotated between lathe centers and the carriage passed by it. Chips were forced out by continuous flow of coolant. The carbide tool was used in many fine successive cuts until the bore was within 0.0005 inch of nominal. The diamond tool was then inserted in the bar precisely without disturbing the work. A single pass with the diamond tool brought the work to final size and finish.

After machining and assembly with precision jigs using liquid nitrogen for shrink fits, the entire piece was reduced in a hydrogen atmosphere at 230°C.

Credit should be given to the Bureau of Standards at Boulder, and in particular to Howard E. Bussey for the valuable assistance he gave us in the techniques of machining copper with diamond tools, and the further use of hydrogen reduction to maintain the surface conductivity.

Of course, no other surface finish is used. Plating or lacquer on the cleaned surface could only increase the losses in some nonrepeatable and nonpredictable manner. There is no evidence that electroplating can really approach the conductivity of the pure metal closely enough to use it for the conducting surface.

Evaluation

The highest frequency these standards are presently used at is 610 Mc. The skin depth in copper here is very close to 200 millionths of an inch. Since the surface finish is in the order of 4 microinches and of a regularly repeating nature, because the surface was developed by turning, the surface conductivity can be considered that of pure copper.

The calculation of the basic impedance of this structure is then undertaken from transmission line equations using reasonably exact relationships which take the copper losses into consideration. Basically, the impedance of a shorted transmission line can be given as:

$$Z = Z_0 \left[\frac{(\alpha l \cos\beta l + J\sin\beta l)}{(\cos\beta l + J\alpha l \sin\beta l)} \right] \quad (1)$$

$$\gamma = \alpha + J\beta \quad (2)$$

$$\alpha \approx \frac{R}{2\sqrt{L/C}} + \frac{G\sqrt{L/C}}{2} \quad (3)$$

$$R = \left(\frac{1}{a} + \frac{1}{b}\right) \sqrt{\mu_0 f/4\pi\sigma} \quad (4)$$

$$\beta = \omega\sqrt{LC}\left(1 + \frac{R^2}{8\omega^2 L^2}\right) \quad (5)$$

$$Z_o = \sqrt{L/C}\left[\left(1 + \frac{R^2}{8\omega^2 L^2}\right) + j\left(\frac{-R}{2\omega L}\right)\right] \quad (6)$$

$$L = \frac{\mu_0}{2\pi} ln \frac{b}{a} \quad (7)$$

$$C = \frac{2\pi\epsilon}{ln(b/a)} \quad (8)$$

$E = 8.855 \times 10^{-12}$ farads/meter

$\mu_o = 4\pi \times 10^{-7}$ henrys/meter

$\sigma = 5.85 \times 10^7$ mhos/meter

l = length of line

a = radius of inner conductor

b = inner radius of outer conductor

These relationships give the series reactive and resistive components of the basic coaxial inductors.

This picture would be complete if the device were attached to a coaxial device. However, the Q Meter is an unbalanced device and a discontinuity will exist at the junction of the standard line and the Q capacitor terminal. Figure 3 illustrates the standard line superimposed upon the high terminals. The unbalanced currents result in excess inductance and resistance in the Q standard. The presence of the high post in the field of the line places a discontinuity capacitance across the line end. The exact calculation of these values would be very laborious because of the strange discontinuity configuration.

As a result of this limitation, a series of measurements were made which would define the reactive components, and, from an experimental knowledge of the reactive components, predict the effect of the current around the junction and then calculate the most probable excess resistance. The internal inductance of the resonating capacitor was first measured, at all settings in use, by short circuiting the terminals with a strap which covered the full ½ inch width of the terminals which are only 0.018 inch apart. When shorted, the resonant frequency of the structure was measured using lightly coupling probes which are a part of the Q Meter. The frequency was accurately measured with an electronic counter. The low frequency capacitance was then determined by comparing the same settings with a GR 722D precision capacitor and a precision bridge. From the capacitance and resonating frequency series L was computed

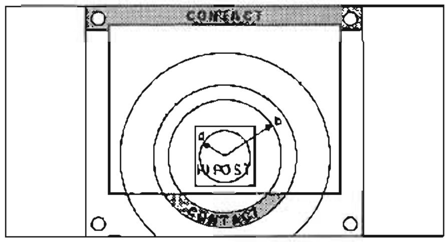


Figure 3. Standard Line Superimposed Upon the High Q Capacitor Terminals of the 280-A

$$L = \frac{1}{4\pi^2 f_o^2 C}$$

The effective X_C present at the terminals at a given test frequency would then be $X_C - X_L$. This then gave a reliable RF figure for X of the internal capacitor.

A series of measurements was then made of the resonating capacitance of various length lines at various frequencies within the range 210-610 Mc. Since the inductance of the coaxial standard could be computed from dimensions, and the X_c of the capacitor could be computed from series resonant frequency and low frequency capacitance, the discrepancy between X_L and X_c could be attributed to the presence of the discontinuity L and C. By graphical plotting it was possible to determine values of L_d and C_d which resulted in better than 2% agreement between the computed X_L and the computed X_c at all frequencies in the 210-610 Mc range. The discrepancy remaining could most likely be reduced by using a more complicated model but this is quite satisfactory for reactance calibration of a 5% instrument. As a result of this experiment, L_d was set at 0.60 nanohenry and C_d at 0.2 picofarad.

The next step is to use this knowledge in a calculation of the most probable discontinuity resistance. It is assumed that the current at the end of the coax line is at its maximum where the perimeter of the line actually contacts the ground stator. However, current does not stop at the end of this area but most likely tapers off gradually toward the noncontacting side because current flows on the end of the coax line. The assymmetrical current flow results in higher order TE modes. If the amplitude of these higher modes were known at the boundary, the value at any other point up the line is approximated by n nepers attenuation per average radius since the line is well beyond cutoff for these modes; where n is the order of the modes being considered. An integration of the excess mean squared current vs. axial travel up the line permits determination of total excess loss due to the presence of higher order TE mode waves. This excess loss can be expressed as an equivalent resistance in series with the TEM mode model of the reactance standard.

Assume that axial current at the discontinuity has a known distribution around the periphery represented by Fourier Series of

$$\frac{1}{2\pi b}\left[1 + P_1 \cos\phi + P_2 \cos 2\phi + P_3 \cos 3\phi + \ldots . P_n \cos n\phi + \ldots\right]$$

Assume further the coax line has 50 ohms characteristic impedance, and the frequency is very much less than the cutoff frequency of the higher modes. Then:

$$L_d = 2.22 \times 10^{-9}\, \Sigma K_L P_n$$
where b = 0.0111 meters
$$R_d = r_s \Sigma K_r (P_n)^2$$

Using the above relationships a number of plausible distributions were tested for which the Fourier coefficient are known. From this a relationship was developed which fits most distributions within a ±5% error. This is quite satisfactory, since the total correction is only a small part of the total resistance. The approximate relationship between L_d and R_d is as follows:

$$\frac{R_d}{r_s} = 0.126 \times 10^9 [L_d]^{1.4}$$

r_s = surface resistivity ohms/sq.

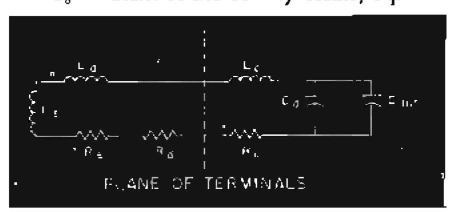


Figure 4. Equivalent Circuit with Impedance Standard Connected to 280-A Q Capacitor Terminals

Until such time as the actual current distribution can be established this relationship will do quite well. As a typical situation, where the discontinuity resistance is 1/10 the resistance of the TEM line, the error in Q will only be 1% if a 10% error in R_d exists. This is certainly in line with the present state of development of these standards.

Credit must be given here to Bernard D. Loughlin, Electronic Research Consultant, Huntington, Long Island, for developing this method of evaluating the discontinuity parameters and for his many invaluable contributions to the concept of the standards.

Measuring Technique

The method in which these devices are used will also contribute to their precision as standards. As previously mentioned the contact button screws into the center hole of the high capacitor stator. When this is inserted it must be clean. It must be also be seated with a precise torque value of 35 inch ounces. This torque value will not break the 2-56 stud and yet makes adequate contact so as to assure no Q deterioriation. The value was derived experimentally as that value which is 25% above the torque value where no readable change occurs with additional torque.

The four screws which hold down the mounting flange are also tightened to this torque. Care is taken to tighten each of the four screws a little at a time and in succession. This is to assure that the standard line is seated properly on the Q capacitor.

The temperature of the copper is also monitored with a thermocouple during the measurement to allow corrections for conductivity and dimension changes which occur with changes in temperature.

The Q is measured by determining the frequency interval between the 3 db points on the resonance curve. The 280-A Q Meter is equipped to measure this internally and, as well, provides an external monitor jack to be used with a precision counter. Q is equal to the frequency at the peak of the curve divided by the bandwidth.

Application

The significant applications of these standards are as calculable Q standards on the UHF Q Meter and as a means for evaluating the internal losses in the self-contained resonating capacitor of the 280-A. A knowledge of the effective inductance of the coaxial standard, as previously described, will define what capacitance should resonate with the standard at a given frequency and thereby give a precise standard for checking capacitor calibration. However, computing the series resistance of the internal capacitor in the 280-A from a knowledge of measured circuit Q is a more involved procedure. Q is fundamentally defined as:

$$\omega_0 \times \frac{\text{Energy stored in ckt.}}{\text{Average power lost}}$$

One may sum the power stored in the inductive reactance of the line and the two lumped reactances $L_d + L_c$. If Q is divided into the product of ω_0 and this total stored energy, the average power lost will be derived. If the power loss in the line is summed and added to the power loss in the discontinuity, the remaining power loss would be attributable to the series resistance of the capacitor. The derivation of this resistor is as follows:

$$Q = \omega_0 \frac{\text{Energy stored}}{\text{Avg. power loss}}$$

Lumped Inductance $L_t = L_c + L_d$

Lumped Resistance $R_t = R_c + R_d$

V_1 = Voltage of transmitted wave

$$I = \frac{2V_1}{Z_0} \cos \beta l$$

Energy stored in line Inductance

$$U_m = \frac{L}{2} \int_0^l \frac{4V_1^2}{Z_0^2} \cos^2 \beta l \, dl$$

$$= \frac{2V_1^2}{Z_0^2} L \left[\frac{l}{2} + \frac{1}{4\beta} \sin 2\beta l \right]_0^l$$

Energy stored in Lumped Inductance

$$U_L = \frac{1}{2} L_t \times \frac{4V^2}{Z_0^2} \cos^2 \beta l$$

$$= \frac{2V_1^2}{Z_0^2} L_t \cos^2 \beta l$$

Average Energy Lost in Line

$$W_R = \int_0^l \left[\frac{2V_1 \cos \beta l}{Z_0} \right]^2 \frac{R}{2} - dl$$

$$= \frac{2V_1^2}{Z_0^2} R \int_0^l \cos^2 \beta l \, dl$$

$$= \frac{2V^2R}{Z_0^2} \left[\frac{l}{2} + \frac{\sin 2\beta l}{4\beta} \right]_0^l$$

Average Energy Lost in Lumped Resistance

$$W_{RL} = \frac{I^2R}{2} = \frac{4V_1^2}{Z_0^2} \cos^2 \beta l \frac{R_T}{2}$$

$$= \frac{2V^2}{Z_0^2} R_t \cos^2 \beta l$$

Then Summing Stored Energy and Average Power Loss and cancelling term: $\frac{2V^2_1}{Z^2_0}$:

$$Q = \omega_0 \left[\frac{\frac{Ll}{2} + \frac{L}{4\beta} \sin 2\beta l + L_t \cos^2 \beta l}{\frac{Rl}{2} + \frac{R}{4\beta} \sin 2\beta l + R_t \cos^2 \beta l} \right]$$

$$Q = \omega_0 \left[\frac{2L\beta l + L \sin 2\beta l + 4\beta l + \cos^2 \beta l}{2R\beta l + R \sin 2\beta l + 4\beta R + \cos^2 \beta l} \right]$$

$$2R\beta l + R \sin 2\beta l + 4\beta R_T \cos^2 \beta l = \omega_0 \frac{[2L\beta l + L \sin 2\beta l + 4\, \beta L_t \cos \beta l]}{Q}$$

$$R_t = \frac{\omega_0 \, [2L\beta l + L \sin 2\beta l + 4\beta l + \cos^2 \beta l]}{4\beta Q \cos^2 \beta l} - \left[\frac{2R\beta l + R \sin 2\beta l}{4\beta \cos^2 \beta l} \right]$$

$$R_c = R_t - R_d$$

The resistance derived by this method can be considered in series with the internal Q capacitor. For precise measurements of external high Q components, this resistance and the series inductance of the capacitor must be considered in series with the externally connected circuit. With this knowledge, the Q of a capacitor can be measured whose losses may be even less than the internal Q capacitor. Without the internal C loss, such a capacitor might even seem to be one with negative losses; and the result would be meaningless.

The present standards under development include 7 different lengths which are designed to be used as a check at high, medium, and low frequencies in the 210-610 Mc range; with 3 capacitance values at each of the three frequencies. These are able to describe a relatively accurate picture of the internal resonating capacitor.

Future Work

As a future check on the relationship between the conductivity of the copper and its performance in the skin of the line, a long line, shorted at both ends, will be constructed from the same copper and machined by the same methods. By means of tiny probes through the wall of the tube, the resonant frequency as a half wave resonator and the bandwidth can be determined. This will give the Q and hence the surface resistivity working backward from the relationship

$$Q = \beta / 2\alpha.$$

This experiment will give an independent check on the conductivity of the copper in the surface, free from the effects of any discontinuities. This is of interest as a final check on the use of these as standards. All previous work has assumed that conductivity at RF is equal to the dc value. This is accurate, most likely, to within two percent (2%) but it will be of great value to verify this experimentally and will perhaps improve the absolute accuracy by some measurable degree.

Conclusion

In summary, the devices described above are stable repeatable standards of impedance specifically for use on the 280-A UHF Q Meter. They are useful for laboratory and field calibration of this instrument within 2% of reactance and very close to that in Q. Further investigation of these pieces should place the Q value within 5%. However, the knowledge of such a small resistance in series with such a high reactance will always have uncertainties. The techniques used here are applicable to standards for any similar impedance measuring system, and in a sense are more applicable to coaxial systems because of the simpler discontinuity picture. Like any standardizing program, this is a continuing one. The needs for better standards are constant. The advances in techniques of copper machining and fabrication described here are not an end in themselves, nor are the methods of analysis, which should be improved by future study.

Checking The New DME And ATC Airborne Equipment With The Navigation Aid Test Set

WILLARD J. CERNEY, *Sales Engineer*

The BRC Navigation Aid Test Set Type 235-A provides all of the RF circuitry required for bench testing the new ATC (Air Traffic Control) transponders and DME (Distance Measuring Equipment) portions of the VORTAC navigation system. The test set (Figure 1) contains three basic interconnected units: a crystal-controlled RF signal generator, a peak pulse power comparator, and a wavemeter. The wavemeter is used for measuring the frequency of the ATC Transponder transmitter, and the signal generator and power meter are used for making both ATC and DME measurements.

An engineering description of the 235-A is given in Notebook Number 24. This article will describe some measurements that can be made with the test set when it is used with the Collins Radio Company's 578X-1 Transponder Bench Test Set or the 578D-1 DME Bench Test Set, and a suitable oscilloscope. Before these measurements are described, a brief history of the navigation aid systems will be given.

NAVIGATION AID SYSTEMS

The first radio navigation aids for aircraft were the low-frequency radio direction finder and radio range equipment. These systems were used for navigating aircraft during cross-country flights, for orienting aircraft at or near the airport, and for instrument landings. Later a system was developed which provided a new and improved technique for instrument landings. This system was called ILS (Instrument Landing System). VOR, a system for measuring bearing to a radio station, was introduced a short time later. The new ILS and VOR systems operate in the VHF region. BRC's types 211-A and 232-A signal generators were designed specifically for use in checking the ILS and VOR systems.

About that same time, FAA put into service Airport Surveillance Radar (ASR) equipment to navigate aircraft in case of loss of radio contact with ground stations, and Precision Approach Radar (PAR) equipment to aid in the landing of aircraft without ILS or with ILS which was not working properly.

These navigation aid systems have played an important part in commercial, military, and private air travel, and should be given a good deal of credit for air travel being as safe as it is today. However, with more and faster aircraft being put into service everyday, the need for new and faster techniques for navigating and identifying aircraft became apparent. Recognizing this need, FAA, in conjunction with the military, installed a new system designed to give not only bearing but range to the radio station. This system, called TACAN, has been installed by the Government at the same locations as the VOR equipment. The two systems may also be combined to form a hybrid system known as

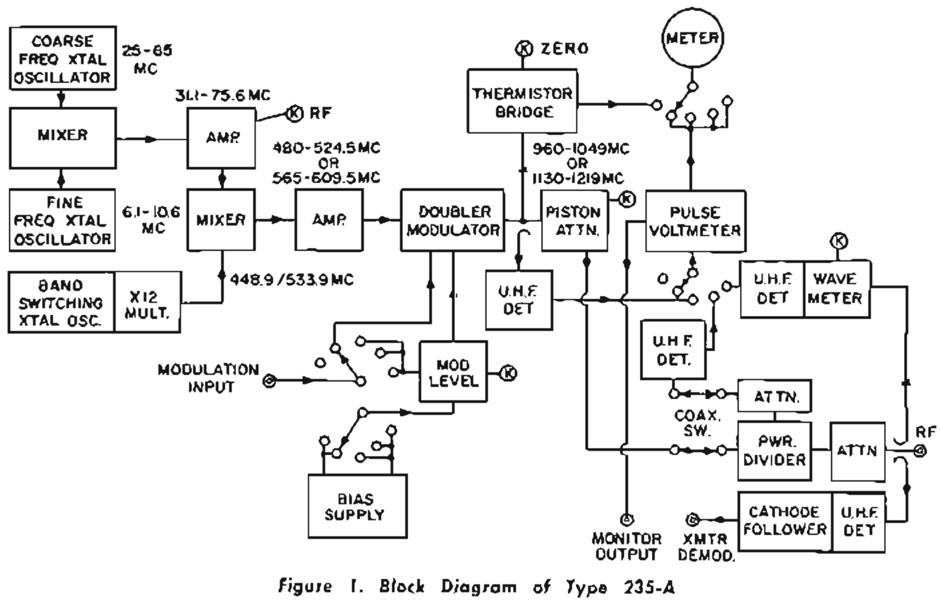


Figure 1. Block Diagram of Type 235-A

VORTAC; with the VOR transmitter being used to determine bearing and the TACAN system being used to determine the distance from the aircraft to the ground station.

The ATC transponder is an automatic receiver-transmitter installed in the aircraft. (This system is similar to the IFF system used during World War II.) When the Air Traffic Controller on the ground wishes to identify an airborne plane, he merely presses a button on his radar console. This operates a radio circuit which automatically transmits a series of coded interrogation pulses to the receiver in the aircraft. A series of coded reply pulses is then automatically sent to the ground station from the plane's transmitter, and appears on the Air Traffic Controller's radar scope. The system is positive and fast enough to fill the requirements of the fast-flying aircraft in use today.

DME MEASUREMENTS

A typical DME radio set consists of an interrogation generator or synchronizer, an encoder, a modulator transmitter-receiver, a decoder, distance measuring circuits, and the indicator and controls in the cockpit. DME measurements which can be made with the Navigation Aid Test Set Type 235-A may be broken down in three groups; transmitter characteristics, receiver characteristics, and distance measuring circuit measurements. The basic setup for performing DME measurements is shown in Figure 2.

Transmitter Power

The 235-A measures, on a comparison basis, the peak power of the pulse train transmitted from the DME equipment. First, the peak of the DME transmitter is measured in a pulse voltmeter circuit and read out on a panel meter. The pulse voltmeter and detector are then switched to read the calibrated output of the signal generator through an adjustable precision attenuator which is adjusted to provide the same level measured for the DME transmitter. The power level is read directly on an attenuator dial.

Transmitter Pulse Characteristics

Certain pulse shapes and positions are required to insure proper operation of the DME transmitter. The following typical DME pulse requirements can be checked with the 235-A, the DME mod-

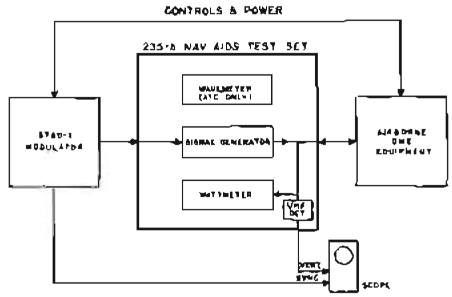


Figure 2. Basic setup for DME Measurements

ulator, and an oscilloscope of suitable dynamic range.

Pulse Characteristic	Typical Requirement (Nominal)
Rise Time	2.5 μsec
Fall Time	2.5 μsec
Duration	3.5 μsec
Pulse Top	5% of maximum amplitude
Repetition Rate	150 pulse pairs (in search position) 30 pulse pairs (in track position)

Receiver Sensitivity

A typical DME receiver sensitivity requirement is that the receiver be capable of locking on a fixed distance 9 out of 10 times. This requirement can be checked with the 235-A, used in conjunction with the DME modulator.

Distance Measuring and Memory Circuits

The functions of the distance measuring circuits are to search for a returned pulse, lock on, maintain lock on in case of momentary loss of signal, and to read out distance. The 235-A, in conjunction with the DME modulator, will check that the search time, memory and prememory time, and distance accuracy are within the tolerances specified by the DME equipment manufacturer.

ATC MEASUREMENTS

A typical ATC transponder consists of a receiver, decoder, encoder, modulator, transmitter, and the necessary cockpit controls. ATC measurements may be broken down in three groups: receiver characteristics, decoder and encoder characteristics, and transmitter characteristics. The basic setup for making ATC measurements is shown in Figure 3.

Receiver Sensitivity

A typical requirement for ATC receiver sensitivity, is that the ATC transponder should give at least 90% replies with a signal level of −74 dbm, and that the sensitivity should be reduced a nominal 12 db for low-sensitivity operation. This requirement can be checked with the 235-A.

Receiver Bandwidth

To measure receiver bandwidth, an oscilloscope is connected to the monitor output connector on the 235-A test set. With the signal generator output set at the level where the ATC transponder is just triggered at center frequency, the attenuator reading is noted. The signal generator output frequency is then changed the desired amount, and the signal generator output is increased until the transponder just triggers again. The difference in attenuator indication is the attenuation for the frequency increment used.

Receiver Dead Time

To check receiver dead time, the second interrogation delay control on the ATC modulator is adjusted so that a full display of second interrogation pulses is observed on the oscilloscope and the receiver response time is measured. A typical requirement is that the receiver be capable of responding in not less than 25 μsec nor more than 145 μsec after the first pulse of the first reply group.

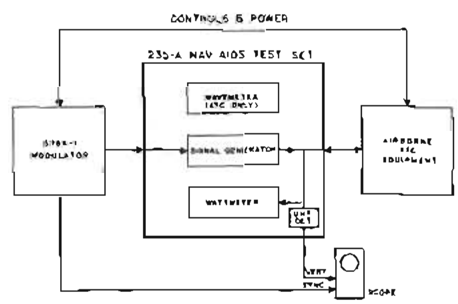


Figure 3. Basic setup for ATC Measurements

Decoder Tolerance

The function of the ATC decoder is to reject all improper signals, such as random pulses, sidelobe pulses, reply pulses from other equipment, etc., that may resemble an interrogating pulse. The decoder pulse spacing is checked by varying the interrogation pulse spacing control on the ATC modulator in a plus and minus direction and observing that the spacing, as displayed on the oscilloscope, is within the tolerances specified. The ability of the decoder to reject sidelobe interrogations is checked by varying the amplitude of the second pulse. The pulse width capability is checked by varying the width of either pulse.

Encoder Measurements

The primary purpose of the ATC encoder is to produce selected reply codes. Presently, there are 64 different reply codes which are set up binarily. Each reply code is made up of 2 framing pulses and 6 code pulses. The spacing between the framing pulses and the code pulses must be within specified tolerances. The 235-A, in conjunction with the ATC modulator and an oscilloscope, can be used to check these tolerances.

Transmitter Frequency

The frequency of the ATC transponder is measured by adjusting the wavemeter in the 235-A for a maximum indication on the front panel meter, with the function selector set for frequency measure operation, and reading the frequency on the wavemeter dial.

Transmitter Power and Pulse Characteristics

The ATC transponder power and pulse characteristic measurements are made in the same manner as the DME transmitter power and pulse characteristic measurements. Pulse characteristics requirements are obtainable from the ATC equipment handbook.

SPECIAL MEASUREMENTS

The measurements described in this article are the basic measurements which can be made using the 235-A. Other measurements such as overall transponder delay, AGC characteristics, AOC measurements, etc., may also be made. Complete ATC and DME measurement procedures are given in the 235-A instruction book and in the instruction books for the ATC and DME equipment.

NEW FM STEREO MODULATOR TYPE 219-A

With the recent FCC approval (Docket 13506) of a system providing entertainment stereo and subsidiary communications in the 88 to 108 Mc FM broadcast band, a definite requirement has developed for a suitable modulator to generate the specified multiplex signals to, in turn, modulate an FM signal generator for the testing of receiving systems.

The new Type 219-A FM Stereo Modulator is designed to provide stereo modulation outputs as specified in the FCC Docket, suitable for modulating the BRC Type 202-E FM-AM Signal Generator or other FM signal generators with adequate modulation characteristics. Provision is made for Left (L) and Right (R) audio stereo channel inputs and/or subsidiary communications FM subcarriers in the 20 to 75 kc range. Preliminary specifications for the Type 219-A are given below.

Input Characteristics

ENTERTAINMENT STEREO

Source: Left (L) and Right (R)

Fidelity: 50 cps to 15 kc

Modulating Oscillator: an internal 1 kc oscillator is provided which, in conjunction with the Type 202-E internal modulating oscillator (50 cps to 10 kc) may be used to furnish stereo inputs.

Type 219-A

SUBSIDIARY COMMUNICATIONS

Source: FM sub-carriers

Frequency Range: 20 to 75 kc

Output Characteristics

ENTERTAINMENT STEREO

Pilot Carrier — Frequency: 19 kc
Accuracy: ±0.01%
Level: 9% of system deviation

Double Sideband Suppressed Carrier (L-R)
Frequency: 38 kc
Accuracy: ±0.01%
Fidelity: 50 cps to 15 kc
Carrier Suppression: <1% of system deviation
Sideband Level: 45% of main carrier modulation with either Left (L) or Right (R) signal
Distortion: <1% at a level corresponding to 45% of system deviation

Monaural Carrier (L + R)
Frequency: 50 cps to 15 kc

Preemphasis: Standard* preemphasis for main (L + R) and stereo (L—R) channels may be switched in or out of circuit

*per Section 3.322 h, FCC Docket 13506

SUBSIDIARY COMMUNICATIONS

Frequency: 20 to 75 kc

Metering — A meter is provided to read the multiplex output in terms of percent of system deviation* of the main RF carrier. The unit is factory adjusted to operate with the Type 202-E Signal Generator. (Alternative adjustment may be made for use with other signal generators.)

*100% = 75 kc deviation

Composite Output — A suitable linear adder is provided to permit summing the monaural channel, pilot carrier, stereo channel, and subsidiary communications FM sub-carriers.

Power Supply — 115 volts ±10%, 60 cps.

HANS SCHLOTT JOINS BRC SALES STAFF

The appointment of Hans Schlott as Regional Sales Manager for BRC was announced in March of this year. In this capacity Hans will direct the BRC sales operations along the east coast from the Metropolitan New York City area south to the Metropolitan Washington D.C. area. Beginning his association with the Company in March proved timely for Hans, as it afforded him the opportunity to serve in the BRC booth at the IRE show. Here he met scores of BRC customers and was able to hear, first hand, their problems concerning measurement instrumentation.

Hans, a native of Sweden, came to the United States in 1949. He graduated from the Charlottenburg Institute of Technology in Berlin, Germany and is an Applied Physicist. He also completed studies in Industrial Management at the Graduate Business School of St. Gall, Switzerland.

Prior to his association with BRC, Hans served with Curtiss Wright's Princeton Division as Senior Sales Engineer, Eastern Territory. Before that he was Eastern Sales Manager for the New Products Division of the Corning Glass Works in New York City.

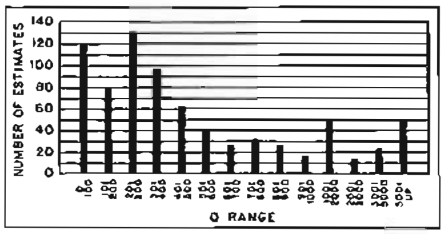
HANS SCHLOTT

EDITOR'S NOTE

The Q of the resonant circuit displayed at the IRE show is 524.2. Winner of the Q Meter, with an estimate of 523.5, is Mr. E. B. Sussman, an Engineering Consultant from Livingtson, N. J.

More than 1000 entries were submitted, with Q estimates ranging from 1 to 20,000. The bar graph below shows the distribution of these estimates. There were nine estimates in addition to the winning estimate which were very close to the actual measured Q and are certainly deserving of honorable mention.

Estimate	*Submitted By*
515	E. Queen, Stuyvesant High School, N.Y.C.
521	H. Korkes, CBS Radio, N.Y.C.
521	J. M. J. Madey, Student, Clark, N. J.
521.5	P. H. Daitch, Microwave Research Inst., Brooklyn, N. Y.
523	J. F. Isenberg, Jr., IBM, Poughkeepsie, N. Y.
523.5	E. B. Sussmann, Engineering Consultant, Livingston, N. J.
525	M. Gellu, FAA/NAFEC, Atlantic City, N. J.
527	J. Brady, Cooperative Ind., Inc., Chester, N. J.
527	P. Bahr, Schon Tool & Machine Co., Inc., Union, N. J.
528	A. Karr, Daystrom Central Research Lab., W. Caldwell, N. J.



Distribution of Q Estimates.

The resonant circuit in question was measured by means of the "in-circuit" technique on the UHF Q Meter Type 280-A at 500 megacycles. Six separate measurements were made on the most sensitive range on the instrument. The average of these measurements was 524.2.

Our congratulations to Mr. Sussmann and many thanks to our many friends who visited with us at the show.

NEELY ENTERPRISES APPOINTED NEW BRC SALES REPRESENTATIVE

The appointment, effective April 1st, of Neely Enterprises as Sales Representatives for Boonton Radio Corporation was recently announced. Neely maintains eight offices in the states of California, New Mexico, and Arizona. A list of these offices is given on page 8 under BRC Engineering Representatives. If you live in one of these states, there is a Neely office conveniently near you. All offices are fully staffed to help fill your electronic needs.

Engineering Representatives

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TWX: AQ-172

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3110 Maple Drive, N.E.
Tel. Atlanta, Georgia 233-1141
TWX: AT 987

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Vestal, New York
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TWX: ENDICOTT NY 84

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TWX: ARL MASS 253

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TWX: CG508

CLEVELAND 24, Ohio
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TWX: DL 411

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BOONTON RADIO CORPORATION

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

A MODULATOR FOR THE NEW FM STEREO SYSTEM

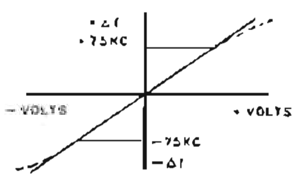
JOHN P. VAN DUYNE, *Engineering Manager*

For the past twenty-one years, the Boonton Radio Corporation has been recognized as a leader in the design and manufacture of frequency modulation signal generators. It is logical, therefore, that when FCC approved an FM Stereo Broadcasting System, as they did on April 20, 1961, that BRC should provide the market with a stable, easy to use, attractive, economical source of the multiplex signal for use with FM signal generators, or for direct use with receiving multiplex adapters. To this end, the 219-A has been designed.

THE FCC APPROVED SYSTEM

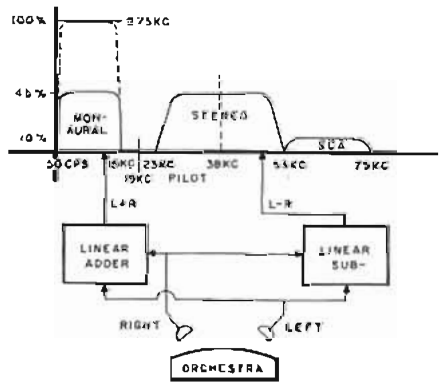
The system, approved by the FCC "Report and Order" dated April 20, 1961, and specified in FCC Docket 13506, provides for the simultaneous broadcasting of a main channel of monaural information, a separate subchannel for the transmission of stereo information, and provision for one or more channels for Subsidiary Communication purposes. This latter assignment may be used for program relaying, "mood music", broadcasting for industrial or commercial purposes, etc.

The FCC approval of the stereo system came after an extensive study of the matter in general, and specifically, the work of the National Stereophonic Radio Committee which provided the medium for field testing and analyzing many of the various stereo broadcasting systems which had been proposed. The author will not attempt to discuss the work of this committee, which has been



"A"

Modulation Characteristic



"B"

Baseband Multiplex Spectrum

Figure 1. New Stereo FM System

adequately documented, but will content himself with briefly describing the presently approved system.

Reference to Figure 1B will show the frequency spectrum of the baseband multiplex signal which is generated by the BRC 219A. The upper portion of this Figure (1A) shows a typical modulation characteristic of an FM signal generator. If the generator is reasonably linear, a single constant can be used to relate deviation of the radio frequency carrier to a voltage applied to the modulating terminals; i.e., $\Delta f = K\,E_{mod}$. (1) With this in mind, the system may be more readily understood by confining one's attention to the baseband multiplex spectrum (Figure 1B), which is fed to the modulating jack of the FM signal generator.

One requirement confronting the designers of this system, that was of key importance to the FCC, was the problem of providing "compatible reception" with receivers already in the hands of the FM listening public. This has been achieved by a matrixing technique which is shown in the block diagram in Figure 1B. The assumption is made here that if there were two parallel-connected microphones "listening to" the output of a program source, such as an orchestra, that the listener to a monaural receiver tuned to this broadcast would hear a rather arbitrary summation of the audio outputs of these two microphones. It can be seen that in the new stereo system, these outputs are indeed summed and applied to the monaural channel; thus providing compatible reception.

In nonstereo broadcasting, the peak deviation resulting from the sum of the left and right outputs is allowed to reach 100% of system deviation, or 75 kc, for the previously approved monaural broadcasting system. In the case of the stereo system, this maximum deviation is reduced to the level of 90%, for reasons that will be discussed shortly.

Since the summed output of the left

YOU WILL FIND . . .

IMPORTANT NOTICE

Boonton Radio Corporation has moved into its new plant and offices in Rockaway Township, New Jersey. Our new address and telephone number are as follows:

Mailing Address: Boonton Radio Corporation
P. O. Box 390
Boonton, New Jersey

Plant and Offices:
Boonton Radio Corporation
Green Pond Road
Rockaway Township, New Jersey

Telephone: OAkwood 7-6400

TWX: ROCKAWAY NJ 866

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and right microphones (L + R) are broadcast in this compatible manner, it simply remains to separately transmit information which is related to the instantaneous difference between the outputs of the Left and Right microphones. In order to minimize the bandwidth required for full fidelity (50 cps to 15,000 cps), this L — R information is transmitted by means of amplitude modulation of a subcarrier located at 38 kc. The subcarrier is suppressed to reduce crosstalk due to nonlinearity in the transmitter or receiver. Distortionless amplitude modulation of this suppressed carrier will, with signals not exceeding 15,000 cps in frequency, occupy the spectrum from 23 to 53 kc (38 ±15 kc). This leaves a portion of the spectrum from 53 to 75 kc for a small amplitude Subsidiary Communications signal.

It is obvious to the reader that the system, described to this point, provides L + R in the normal audio frequency portion of the spectrum (50 to 15,000 cps) and the sidebands of the amplitude modulated suppressed carrier (L — R) information in the 23 to 53 kc region. However, since the 38 kc carrier is suppressed to less than 1% of system deviation, it would be extremely difficult to demodulate this information in a distortionless, low-noise manner. Therefore, a pilot carrier at exactly one-half the suppressed carrier frequency is transmitted at reduced deviation of the main carrier.

The 19 kc pilot carrier is specified to produce a deviation of the main carrier between 8 and 10% of the system deviation of 75 kc. Since this 19 kc carrier is located with a guard band of 4 kc on either side, as shown in Figure 1B, it is evident that a practical filter can be used to extract the frequency and phase information carried by this carrier. If the filtered pilot is doubled in frequency, maintaining the proper phase relationship, it may be used to demodulate the L — R information on the subcarrier by means of simple amplitude modulation detectors in which the locally generated carrier is mixed with the L — R sidebands. The potential is also provided for the use of exalted carrier demodulation in the interests of good signal-to-noise ratio and low distortion.

Thus, we see that the multiplex stereo signal can be rather simply described by a plot of voltage generated versus the baseband frequency spectrum to 75 kc. The multiplex signal may also be described mathematically as follows:

$$E_{mx}(t) = \underbrace{(L+R)\sin(\omega_b t+\phi)}_{\substack{\text{monophonic}\\ \text{baseband}}} + \underbrace{\frac{L-R}{2}\left[\cos\left\{(\omega_c-\omega_b)t-\phi\right\}-\cos\left\{(\omega_c+\omega_b)t+\phi\right\}\right]}_{\substack{\text{stereophonic sidebands}\\ \text{of suppressed carrier } \omega_c}} + \underbrace{aE_p\sin\omega_p t}_{\substack{\text{pilot carrier}\\ 2\omega_p=\omega_c}} \quad (2)$$

A major feature of the L + R, L — R matrix system is that time-sharing between the monaural and stereo channels is automatically provided. Thus, if at an instant in time, the outputs of the left and right microphones were synchronous in frequency and identical in phase, a maximum output would exist in the L + R channel, and a zero output would exist in the L — R channel. Conversely, if the two microphone outputs were equal in amplitude and frequency, but opposite in phase, then the maximum output would exist in the L — R channel and zero in the L + R channel. If the maximum permissible deviation resulting from the L + R audio or from the sum of the L — R sidebands is limited to 45% (33.75 kc) with either left only or right only by FCC, then it can be seen that no combination of Left and Right signals can add to produce more than 90% system deviation. Thus, the 6 db signal-to-noise degradation

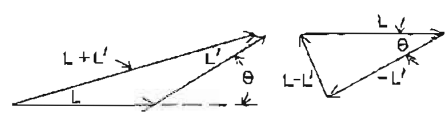
ASSUME:

1. Left (L) signal only being transmitted
2. (L+R) channel response is such that its output differs from (L−R) channel as though it were multiplied by a transmission fact $a\angle\theta = a = (\cos\theta + j\sin\theta)$
3. Stereophonic separation $\equiv R_s$

$$R_s = 20\log_{10}\left(\frac{\text{magnitude of output from Left channel}}{\text{magnitude of output from Right channel}}\right)$$

4. a. Let L $\equiv$ output from L − R channel (Left signal only)
   b. Let L′ $\equiv$ output from L + R channel (Left signal drive only)

VECTOR DIAGRAM:



$$R_s = 20\log\left(\frac{\overline{L+L}}{\overline{L-L}}\right)$$

$$L+L' = L + aL(\cos\theta + j\sin\theta) = L(1 + a\cos\theta) + jaL\sin\theta$$
$$L-L' = L - aL(\cos\theta + j\sin\theta) = L(1 - a\cos\theta) - jaL\sin\theta$$

$$R_s = 20\log_{10}\sqrt{\frac{1+2a\cos + a^2\cos^2\theta + a^2\sin^2\theta}{1-2a\cos\theta + a^2\cos^2\theta + a^2\sin^2\theta}} = 20\log_{10}\sqrt{\frac{a^2+2a\cos\theta+1}{a^2-2a\cos\theta+1}}$$

CHECK:

1. If $a = 1$, and $\theta = 0$; $a^2 - 2a\cos\theta + 1 = 2 - 2 = 0$, $\therefore R_s = \infty$
2. If $a = 1$, and $\theta =$ any value; $R_s = 20\log_{10}\sqrt{\dfrac{1+\cos\theta}{1-\cos\theta}}$
3. If $\theta = 0$, and $a =$ any real number; $R_s = 20\log_{10}\left(\dfrac{a+1}{a-1}\right)$

NOTE: These same equations may be used for the case where L − R channel output is multiplied by $a(\cos\theta + j\sin\theta)$ but reference vector is now output of L + R channel.

Figure 2. Stereo System with Identical Inputs (Assumed as Left Only) Fed Into the L+R and L−R Channels

which would result from transmitting Left and Right in the two channels is reduced to 1 db. The remaining 10% system deviation being reserved for the 19 kc pilot subcarrier.

There are several system transmission characteristics which must receive careful attention if the maximum capabilities of the system are to be reached. In any system that transmits stereo multiplex information, it is desirable that the maximum separation be retained between a signal which initiates output only in the Left channel or the Right channel. The ability of a system to produce the largest possible ratio of the signal in the left-hand loudspeaker at the receiving end, to the signal in the right-hand loudspeaker, when excited by a pure Left signal, is of interest (similarly for a pure Right signal). This ratio is termed "stereophonic separation".

It is a major endeavor of the newly approved system to maintain 30 db of stereo separation over the entire audio frequency range of 50 cps to 15,000 cps. If the system standards are adhered to, a broadcast transmitter can be assured of providing a signal capable of maintaining this separation. Considerable attention to detail is required to achieve this, however. Chief among these are the need for amplitude flatness of better than 3½% over the 50 to 15,000 cps frequency range. In addition, the time delay in the L — R channel must be so nearly equal to that in the L + R channel that not more than 3 degrees of differential phase shift exists between an audio frequency transmitted by one channel relative to the other channel.

This can be readily demonstrated if one will picture two parallel transmission systems driven with the same input and then take the outputs in these two systems and add them vectorily and then subtract them vectorily in two isolated circuits. Figure 2 illustrates the situation that exists in the stereo system when identical inputs (assumed as Left only) are fed into the L + R and L — R channels. If we consider that the transmission of the one channel differs from that of the other by a magnitude "a", and by an angle θ, we can then solve for the residual output which will exist in the difference circuit. If this residual output is defined as R_s, we find that:

$$R_s = 20 \log_{10} \sqrt{\frac{a^2 + 2a \cos\theta + 1}{a^2 - 2a \cos\theta + 1}} .$$

Figure 3 shows how stereo separation depends on values of "a" for any angle θ from 0 to 30 degrees.

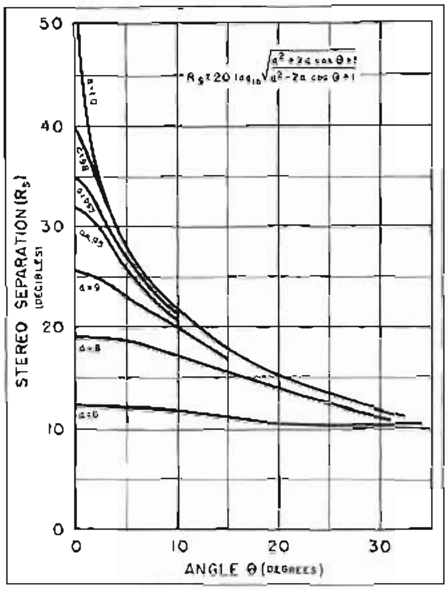


Figure 3. Stereo Separation (R_s) Versus Angle θ

TRANSMISSION OF THE MULTIPLEX STEREO SIGNAL

After the signal has been generated and exists in the form shown in Figure 1, or as described in equation 2, there are certain requirements placed on the components which transmit or process the multiplex signal if the stereo separation is to be maintained. These requirements are quite similar to those necessary to faithfully reproduce transient phenomena, in that the amplitude response must be quite flat and the time delay over the band must be of constant value. It is a well-known and unfortunate fact that most simple networks do not exhibit these characteristics over their entire passband.

A typical network that is of interest in connection with FM signal generators and receivers is shown in Figure 4. This figure illustrates a simple "constant k" low-pass filter section. The LC low pass is typical of the RF filters that exist between the modulation terminals and the reactive modulating element of most FM signal generators to prevent leakage of the carrier via the modulating leads. This filter is frequently one of the major limitations to the electric fidelity of the FM channel of such generators. It can be easily seen that only about 20% of the passband of the LC circuit meets the constant time delay criterion. Therefore, if constant time delay were a requirement to be maintained up to 50 kc, the circuit should actually have a bandwidth of 250 kc to provide negligible distortion of the multiplex signal.

It actually turns out that if the filter shown in Figure 4 were to contribute the maximum allowable delay error (3° at 15 kc, or 0.56 μsec.) that it is possible to operate with the upper sideband of the FM multiplex system at $0.29F_0$, where F_0 is the nominal cutoff frequency of the filter.

This fact is of considerable importance in selecting a suitable frequency modulation signal generator, since it shows that the actual modulation channel bandwidth will have to be 3 to 5 times 53 kc in order that phase-distortionless transmission can be relied upon. This is true of the BRC 202E Signal Generator, which is only 1 db down at 200 kc and exhibits essentially constant time delay up to 75 kc of the modulating frequency. There are many good FM signal generators of older design (for example, the BRC 202B) which have electric fidelity of the order of 15 to 20 kc. These are generally unusable with the stereo multiplex system, unless extensive work is done to predistort the signal for use with these narrower band generators. This can be done, but usually entails individual measurements and tests on a given generator.

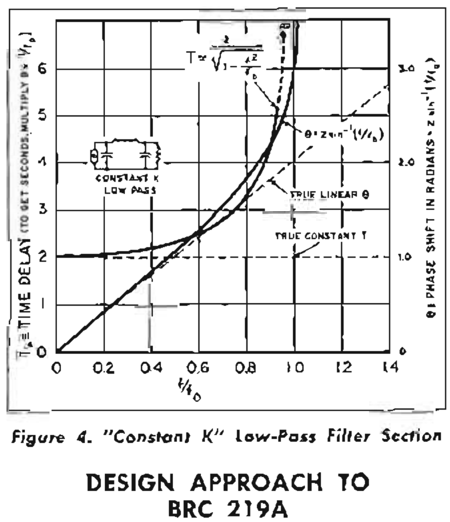


Figure 4. "Constant K" Low-Pass Filter Section

DESIGN APPROACH TO BRC 219A

With this background, we are ready to turn to the factors guiding the design of the 219A. Reference to the block diagram of Figure 5 will quickly show that the "classical" approach to generating the signal has been used in this instrument. This was done to provide a maximum of stability, ease of understanding, and flexibility in providing the specified and certain desired "out of specification" signals for the purpose of receiver and multiplex adapter testing. The major objectives of the design were to achieve simplicity of use, operational

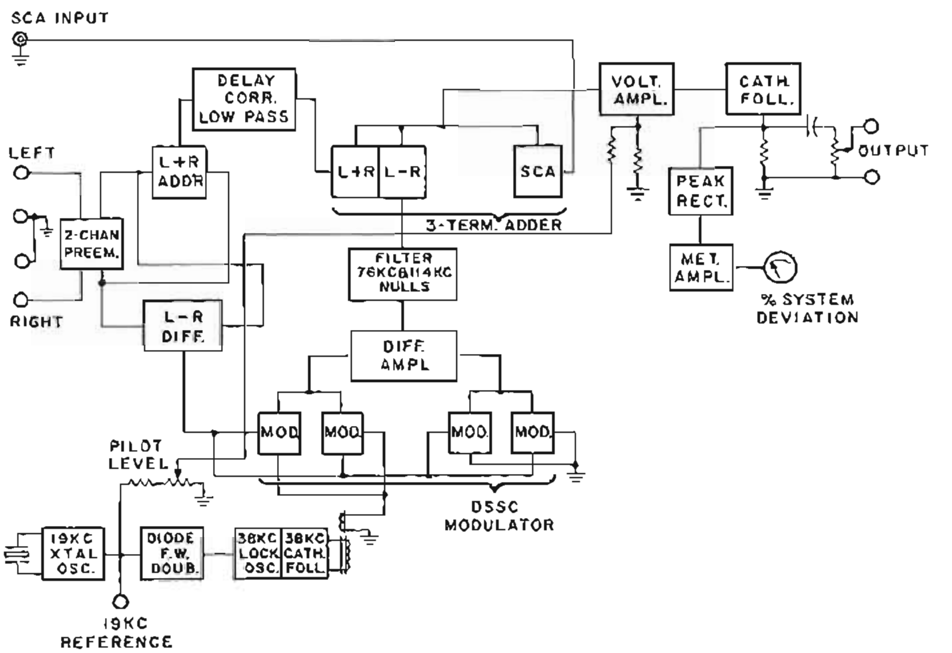


Figure 5. Block Diagram of the Type 219-A

stability, and self containment when used with the BRC 202E and comparable signal generators.

INPUTS PROVIDED

Separate left and right inputs of nearly identical phase and frequency response are provided. 1.7 volts rms typically is required in left only or right only for 45% of system deviation. The input impedance is 10 k ohms. In addition, an input is provided for an FM subcarrier generator to simulate the SCA signals with which receivers may have to deal. This requires 1.0 rms volts into 10 k ohms, for 10% of system deviation.

It should be noted that the 219A is particularly easy to use with a BRC 202E, since audio frequencies from 50 cps to 10 kc, in several steps, may be supplied from the internal modulating oscillator of the 202E. In addition, a stable, balanced source of 1 kc is available inside the 219A. A normal-reverse switch is provided on the 219A panel to permit either of these two oscillators to be connected to the Left and Right channels. Thus, different frequencies are supplied for testing the two channels, and the variable frequency oscillator in the 202E can be used for fidelity measurements from 50 cps to 10 kc. Obviously, an external audio oscillator of any suitable type may also be used.

CONTROLS

The chart shown in Figure 6 lists the controls provided and the functions which they perform. Particularly noteworthy are the matrix and mode switches. The matrix switch permits simple checking of the 219A for L — R null and L + R null, which permits a check on the internal alignment of the 219A. In addition, in these positions, outputs are provided using the internal 1 kc source, which permit crosstalk checking in a receiver or multiplex adapter between the L + R and L — R kc channels. This crosstalk is not to be confused with stereo separation which is measured with a signal in only the Left or the Right channel. The mode switch permits an orderly checkout of the makeup of the multiplex signal, using the internal peak reading meter.

For other than the most advanced measurements, the meter integral to the 219A is all that is needed to verify proper setup of the multiplex signal. In addition, it can be used in conjunction with various positions of the mode switch and panel adjustments, to set up a non-standard signal of the type which may be needed to simulate propagation effects on the multiplex signal. For example, the pilot carrier may be adjusted over a range of 0 to 30% to simulate the effects of multipath transmission, or non-flat transmission systems.

Since many receivers will use phase-locked sub-carrier oscillators, variation in level of the pilot carrier is necessary to test their performance. Knobs are provided to adjust the pilot carrier level and the absolute output level of the 219A. A reference phase 19 kc output is provided to facilitate this.

The output level adjustment permits the use of the 219A with most FM signal generators. In use, it is only necessary to set the mode switch to the "SET" position, and to set the internal 1 kc level so the meter in the 219A reads 100% with the output connected to the FM signal generator. The controls of the FM signal generator are then set

FRONT PANEL CONTROLS	FUNCTION
NORMAL-REVERSE	Interchanges input channels, including internal 1 kc oscillator.
PREEMPHASIS	Inserts 75 μsec. preemphasis in L and R channels.
1 KC	Connects and varies level of internal 1 kc oscillator to right channel.
MATRIX	Normal — See block diagram, Figure 6.
	L — R Null — Connects inputs of adder and subtracter to internal fixed 1 kc source.
	L + R Null — Connects adder and subtracter inputs to opposite polarity, equal amplitude internal 1 kc.
MODE	Set — Connects internal 1 kc source to output meter to set output for 75 kc deviation.
	L + R — Input signals go through adders to output; all else off; meter 100%.
	19 KC — Pilot on only; meter calibrated 6-10%, set pilot level.
	38 KC — Subcarrier on only, meter 0-10% to balance subcarrier.
	L — R — Pilot on; L + R off, meter 100%, inputs connected, set L — R gain.
	MULTIPLEX — All on; normal operation.
MULTIPLEX OUTPUT	Adjusts system output.
PILOT LEVEL	Adjusts level of pilot carrier.
SCREWDRIVER ADJUSTMENTS	**FUNCTION**
MODULATOR BALANCE	Fine balance adjustment for modulator.
L—R GAIN	Adjusts gain of L — R channel.
ZERO	Sets electrical zero of output peak voltmeter.

Figure 6. Type 219-A Control Functions

to provide 75 kc deviation. After that, if the output level control is not disturbed, the 219A will supply the proper relative levels of other signals which make up the multiplex spectrum.

Three screwdriver controls are provided: One permits adjustment of the suppressed carrier modulator amplitude balance, which although very stable, may experience changes from time to time as the tubes age differentially. Another, an L — R gain balance control, is provided for the same reason. The third control permits electrical zeroing of the output meter.

SPECIFICATIONS

Following are the detailed performance specifications of the 219A. When used with a BRC 202E, the combination accurately simulates an FM broadcasting station of superior performance to that specified in FCC Docket 13506.

modulator which transmits the L — R signal. The circuit used is shown in block diagram form in Figure 7. It uses four vacuum tubes biased to maximize the second order coefficient in the power series representing the plate current as a function of grid cathode voltage. In this circuit a balance of the 38 kc carrier is achieved which is more than 60 db below the fundamental components of plate current. In addition, excellent balance of the baseband audio signals from 50 cps to 15,000 cps is achieved. As the modulator generates the desired sidebands by means of the second order coefficient, it should be expected that fairly large second harmonics of both the baseband and the carrier would be generated. This is true for the basic modulator involving the pair of tubes shown in the left of the diagram. The subcarrier second harmonic at 76 kc is far enough removed from the desired output spectrum to be easily

baseband. This second harmonic is combined with the basic modulator output by means of the difference amplifier shown in the diagram, to produce a bucking baseband second harmonic output. The combined output baseband second harmonic is considerably more than 40 db below the desired sideband at 90% modulation.

As a result of this choice of modulator, an extremely stable circuit is provided which needs little output filtering. The baseband and carrier signals are supplied at low level and very low impedance, thus minimizing interaction and providing for extremely flat frequency and phase response, since no transformers need be involved in the broad band circuits. However, if the subcarrier signal source were rich in harmonics, unwanted outputs would result which could be only partially eliminated by a complex filter in the output of the modulator.

This problem has been greatly mini-

Input Characteristics

LEFT (AND RIGHT) INPUT

Frequency Range: 50 cps to 15 kc

Level: 1.7 volts rms Left (or Right) only gives 45% output; simultaneous inputs yield 90% system deviation.

Impedance: 10 k ohms

Preemphasis 75 μsec.: May be switched in or out.

SUBSIDIARY COMMUNICATIONS (SCA) INPUT

Frequency Range: 20-75 kc.

Level: 1.0 volts rms for 10% system deviation, typically.

Impedance: 10 k ohms

Modulating Oscillator Characteristics

Frequency: 1 kc

Accuracy: ±10%

Distortion: <1%

Connections: Switchable into Left or Right Inputs

Output Characteristics

Level: 0 to 7.5 volts peak of Multiplex Signal.

Load Impedance: Not less than 1500 ohms shunted with not more than 200 μμf

Residual Hum and Noise: 60 db or more below 100% output

Metering:
Range: 0-10%, 0-100%
Accuracy: ±2% of full scale

Matrix:
Normal: Output as selected by Output Mode Switch
L+R Null: Left input equals — (Right) input for internal 1 kc oscillator only.
L—R Null: Left input equals right input for internal 1 kc oscillator only.

Output Modes: Switchable to Set, L+R, 19 kc, 38 kc, L—R, and Multiplex

PILOT CARRIER

Frequency: 19 kc

Accuracy: ±.01%

Level: 0-30% of System Deviation.

MONAURAL (L+R)

Level: 0 to 100%

*Fidelity: ±1 db from 50 cps to 15 kc

*Distortion: <1%

*measured at 45% System Deviation

DOUBLE SIDEBAND SUPPRESSED CARRIER (L—R)

Carrier Suppression: <½% System Deviation

Level: 0 to 100%

*Distortion: <1%

*measured at 45% System Deviation

SUBSIDIARY COMMUNICATIONS (SCA)

Level: 0 to 20% System Deviation

Fidelity: 20 to 75 kc ±0.5 db

19 KC SYNCHRONIZING SIGNAL

Level: 0.5 volts rms Typical

PHYSICAL CHARACTERISTICS

Mounting: Cabinet for bench use; readily adaptable for 19" rack mounting

Dimensions: Height 5⁷⁄₃₂", Width 16¾", Depth 10⅜"

POWER REQUIREMENTS

105-125 volts, 60 cps, 130 watts

CIRCUIT DETAILS

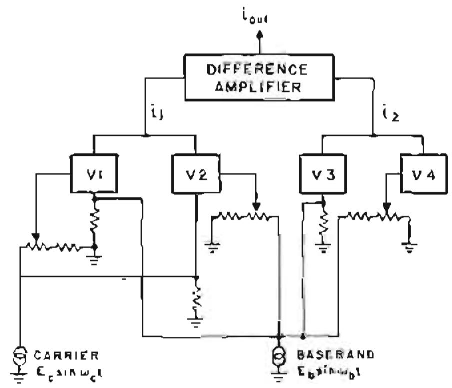
There are several novel circuits which permit the 219A to be stable and precise in its performance. A major factor in the design of such an instrument is the double sideband suppressed carrier

filtered out. However, the second harmonic of the baseband above 11.5 kc falls within the desired output spectrum of the modulator and therefore is not filterable. The second pair of tubes shown in Figure 7 corrects this problem by generating a second harmonic of the

mized by the use of a high purity 38 kc oscillator of the amplitude stabilized variety. This oscillator has high phase stability which is directly controlled by the second harmonic of a highly stable 19 kc crystal oscillator. A schematic of this circuit is shown in Figure 8. While amplitude stabilized oscillators have been well known in the art, this is a simple, effective circuit of high stability which fits in nicely with the modulator previously described.

To avoid the use of expensive broadband phase linear transformers in the adders and subtractors, sum and difference amplifier circuits were used. The difference amplifier shown in Figure 9 maintains its discrimination against common mode signals by 60 db or more for a wide range of voltages and tube characteristics. It is used in two places in the 219A, as may be seen by reference to the block diagram (Figure 5). One use is for the initial subtraction of the Left and Right signals in the early portion of the block diagram. In addition, it is used in place of a broadband phase linear transformer to subtract the second harmonic contribution of the balance modulator from the main modulated signal source as described above.

The use of the phase-locked oscillator provides a high degree of phase stability and at the same time has great freedom from unwanted phase modulation effects, thus providing a high degree of phase stability between the pilot carrier and the suppressed carrier.



i_1 contains the following terms:

$C_2E_cE_b\left[\cos(\omega_c-\omega_b)t - \cos(\omega_c+\omega_b)t\right]$ desired output

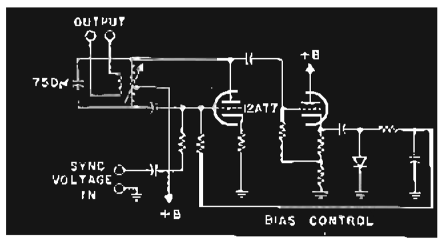
$\frac{C_2E_c^2}{2}\sin 2\omega_c t$ undesired carrier second harmonic

$\frac{C_2E_b^2}{2}\sin 2\omega_b t$ undesired baseband second harmonic

i_2 contains only $\frac{C_2E_b^2}{2}\sin 2\omega_b t$

$i_{out}=i_1-i_2$, thus cancelling undesired baseband second harmonic

Figure 7. Block Diagram — Double Sideband Suppressed Carrier Modulator



Typical 38 kc power output 0.05 mw
76 kc down > 45 db
114 kc down > 70 db
152 kc down > 75 db

Figure 8. Low Distortion Amplitude Stabilized Oscillator

TESTS WHICH CAN BE PERFORMED WITH THE BRC 219A

Reference to Figure 10 shows the 219A being used in conjunction with the BRC 202E FM Signal Generator for receiver testing. It has been pointed out that this combination represents an extremely versatile self-contained package for receiver testing. Obviously, the 219A may be used by itself for direct testing of a multiplex adapter as it supplies up to 7.5 volts peak of composite output into load impedances as low as 1500 ohms. It is obvious that the combination of these two instruments will enable many receiver test and alignment functions to be carried out expeditiously. While the following description will cover those tests which may be made on a completed FM stereo receiver, by deletion of the reference to the RF portion of the system, many of these tests may be performed directly on a multiplex adapter.

Stereo Separation and Matrix Adjustment

This is, of course, one of the major characteristics of a stereo system and will receive much attention from receiver designers and servicemen. It is a simple matter to make these measurements at 1 kc since no external audio signal generators are required, nor are any external connections needed, other than the connection of the 219A to the FM generator or to the multiplex adapter.

In order to make this test, the matrix switch is placed in the normal position, the 1 kc oscillator is turned on, and the mode switch set to L+R. The amplitude of the 1 kc output is adjusted for 45% on the meter, and the mode switch is moved to the multiplex position. Under these conditions, a right only signal is being fed through the system. If the receiver has not been previously aligned, the matrix adjustments may be made for maximum 1 kc in the Right channel and minimum 1 kc in the Left channel. When the normal-reverse switch on the 219A is set to the reverse position, the reverse situation will be true, and the stereo separation and receiver outputs may be measured at 1 kc.

If this information is desired at different frequencies, the internal 1 kc oscillator may be turned off and connection may be made between the AM terminals of the BRC 202E and the Left input terminals of the 219A using a special cable with a variable attenuator. Under these conditions, with the normal-reverse switch in the normal position, signal output will appear in the Left channel. The proper level may be set to correspond to 45% system deviation by again switching to the L+R position and adjusting the input signal level for 45%. Of course, this measurement may be made at other levels simply by adjusting the audio input level as required.

Electrical Fidelity

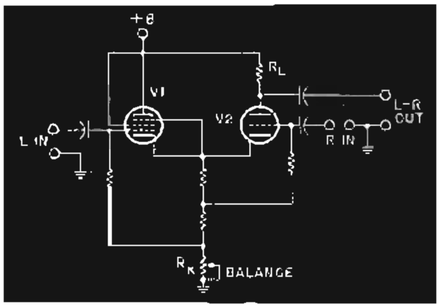
With the setup as described above, and by varying the frequency of the BRC 202E audio oscillator, it is possible to quickly determine the electrical fidelity between 50 and 10,000 cps of the receiver circuits. A simple output meter of adequate frequency response is the only additional equipment required. An external AF oscillator is required for measurement from 10 kc to 15 kc.

L+R — L—R Crosstalk

When the matrix switch is placed in the L—R null position, the Left and Right inputs are connected together to the internal 1 kc signal source. This should produce little or no signal to the input of the double sideband suppressed carrier modulator. Under these conditions, little output should come from the subcarrier detector in the multiplex device. Using this connection, and varying the RF signal level by adjusting the attenuator of the 202E, will show if overload effects from the receiver or the multiplex circuitry are occurring which would cause crosstalk prior to demodulation of the monaural and stereo signals.

NON-STANDARD SIGNAL MAKEUP

By manipulation of the pilot carrier level knob, non-standard levels of pilot carrier may be adjusted from 0 to about 30% of system deviation. The meter, in the 19 kc mode switch position, is calibrated from 6 to 10% only in the interest of maximum readability for the standard setting of the pilot carrier. Larger levels of pilot carrier may be metered by placing the mode switch in



Balance occurs when:

$$gm_1 = \mu_2 \left(\frac{1}{r_{p1}} + R_k\right)$$

$$\text{or } R_k = \frac{gm_1}{\mu_2} - \frac{1}{r_{p1}} = \frac{gm_1 r_{p1} - \mu_2}{\mu_2 r_{p1}}$$

$$\text{but } gm_1\ r_{p1} = \mu_1$$

$$\text{so } R_k = \frac{\mu_1 - \mu_2}{\mu_2 r_{p1}} \approx \frac{gm_1}{\mu_2}$$

$$\text{if } \mu_1 >> \mu_2$$

Figure 9. Inherently Balanced Difference Amplifier

the L—R position with no audio input. As the 38 kc subcarrier is well nulled, the pilot carrier only causes the meter to read. The pilot may then be set for the desired percentage of system deviation as read on the 0-100% system deviation meter scale. When the mode switch is returned to the multiplex position, the desired excess pilot carrier will be present, and tests may be run.

Figure 10. Type 219A and Type 202E Interconnected for Receiver Testing

Screwdriver adjustment for the L—R gain provides a range of ±10 db of L—R relative to L+R gain.

SUMMARY

From the preceding, it may be seen that the BRC 219A is a source of FM stereo multiplex baseband signals of considerable flexibility and involving some novel circuits in the interests of good performance and high stability. It should be re-emphasized that the combination of the 219A and the BRC 202E give the customer modulated RF stereo multiplex signals of a quality better than the FCC specification and in packages that are designed for stability of calibration, ease of control, and long life. The writer is indebted to W. N. Frick and R. W. Houskamp who contributed greatly to the design of this instrument.

BRC IN FULL SWING AT NEW PLANT

Boonton Radio Corporation is happy to announce that it is now situated in its new plant and offices, and that operation is again in full swing. The new facilities are located on a 70-acre site, near the recently completed Route 80 in Rockaway, New Jersey; approximately 7 miles from the old plant site. An announcement of the new plant address, telephone number, and mailing address is given on the first page of this issue. It should be noted that the Company is retaining the Boonton mailing address.

The administrative offices and engineering laboratory and the production sections of the building are interconnected at the upper level by a section which houses a cafeteria area and the model shop. This section acts as a buffer between the office and production areas.

The building is of ultra-modern design with exterior walls of pre-cast concrete and glass. The exterior walls of the office area are constructed almost entirely of tinted glass. These walls are recessed below the walls of the upper level, the overhanging upper level forming protection for a promenade which extends around the entire lower level. The recessed walls and the tinted glass provide protection against direct sunlight.

Our new plant is completely air conditioned and is equipped with the latest in production and laboratory equipment. Among some of the new facilities are a completely equipped plating room, a paint shop, and a cafeteria area. All of the other facilities have been enlarged and modernized.

The plant is designed on a modulator basis to allow for future expansion. The unit now completed provides 60,000 square feet, or more than twice the area available in the old plant. Ultimate expansion calls for four modular units which will provide 320,000 square feet of working and storage area.

A series of articles about our new plant will be published in future issues of the Notebook.

SERVICE NOTE

Adjustment of Q Dial Lock Tension on the Type 280-A

It is possible that, after prolonged use, the HIGH CIRCUIT Q and CIRCUIT Q dial locking mechanisms on the Type 280-A UHF Q Meter will require adjustment. To adjust the dial lock mechanisms proceed as follows.

1. Using a No. 8 Allen wrench, remove the two setscrews that fasten the Q dial control knob to the control shaft.

   NOTE: The procedure is the same for both the HIGH CIRCUIT Q and CIRCUIT Q dial lock mechanisms.

2. Lock the Q dial with the Q lock control.
3. Using a No. 4 Allen wrench, loosen the setscrew on the Q dial locking collar.
4. Turn the locking collar clockwise to remove all tension on the collar.
5. Turn the collar clockwise until it is just finger tight, then continue rotation for an additional 135° or about 3/8 of a complete revolution.
6. Tighten the setscrew on the locking collar.
7. Replace the Q dial control knob and tighten the two setscrews.
8. Check the operation of the Q dial wtih the dial locked and unlocked. Operation should be smooth and positive. There should be no slippage with rapid movements of the Q dial in either direction.

LAHANA & CO. APPOINTED BRC SALES REPRESENTATIVE

Boonton Radio Corporation is pleased to announce the appointment of Lahana & Company as exclusive sales representatives for BRC in Colorado, eastern Montana, Utah, and Wyoming.

EDITOR'S NOTE

BRC Expands Sales Engineering Staff

We are pleased to report that BRC has recently expanded its sales engineering staff. This is in step with BRC's overall expansion program, which in recent weeks has seen the completion of a new plant providing vastly improved engineering and production facilities. This increase in the sales engineering staff was made to improve service, not only to direct sales customers, but to all of our customers, through our sales engineering representatives around the world.

Most of our readers already know the BRC sales engineers: Charles W "Chuck" Quinn was introduced in Notebook 22; Willard J. "Will" Cerney's biography appeared in Notebook 25; and a story on Hans H. Schlott was published in Notebook 29. In addition to handling sales in our local area, Chuck, Will, and Hans are responsible for aiding in the development of new

applications; handling the introduction and initial evaluation of new products; participating in sales exhibitions, seminars, and meetings; and contributing to the BRC Notebook, BRC Bulletin, and other publications.

To better organize our sales service, we have split our local area into three territories: Chuck handles New Jersey and Eastern Pennsylvania; Hans covers the Metropolitan New York area; and Will has been assigned to the Metropolitan Philadelphia, Baltimore, and Washington, D.C. areas. While our sales engineers still make their headquarters at the plant, they regularly tour their respective territories and, in case of an emergency, can generally be at a customer's door within hours. Old friends or new are encouraged to drop them a line or give them a call for application engineering assistance.

Since most of our readers will be in touch with the BRC sales staff, at one time or another, we would like to take this opportunity to introduce the entire group. In the photograph, from left to right, are: front row; Marion A. Derrico, Domestic Order Processing; Evelyn D. LaHart, Export Order Processing; second row, Eleanor D. Matschke, Literature Requests; Grace L. Stone, Secretary to Sales Manager; Willard J. Cerney, Sales Engineer; back row, Frank P. Montesion, Technical Editor; Charles W. Quinn, Sales Engineer; Harry J. Lang, Sales Manager; Hans H. Schlott, Regional Sales Manager; Harry A. Schmidt, Technical Writer; and Bruce A. Barnes, Sales Coordinator.

All of these people are anxious to serve you. They may be reached by telephone at OAkwood 7-6400, or by TWX at Rockaway NJ 866.

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DESIGN OF AN IMPROVED FM-AM SIGNAL GENERATOR

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The age of missiles and satellites has suddenly created a need for communications and data transmission systems which impose tight requirements on FM receivers. For example, FM telemetering systems have squeezed more and more carrier frequencies into their band, and each carrier is required to contain more information per unit time than has been needed in the past. These considerations alone imply that FM receivers for telemetering purposes be capable of handling higher modulation frequencies than they have in the past and, at the same time, provide for isolation of the more closely spaced subcarriers. It is important that the modulated carriers in such systems be free of extraneous sidebands in order to minimize crosstalk between adjacent subcarriers. Hence, the transmitters used in these systems are apt to have a highly linear FM characteristic (frequency vs. voltage characteristic). In the world of FM signals for home entertainment too, the importance of FM linearity is stronger than ever, now that we are faced with the closely spaced (L+R) and (L—R) stereo channels.[1] The 202H and 202J FM signal generators are intended to assist engineers and technicians associated with the function and/or development of such communication systems.

In view of the above considerations, BRC decided to concentrate its design effort on obtaining a linear modulation characteristic. Readers familiar with the 202E and 202G, which these new instruments replace, are aware that these instruments have an exceptionally stable FM characteristic. Since the 202 line has been much admired for this characteristic by communications engineers for nearly fifteen years, it was decided to build on the same basic design rather than to embark on a new idea. Hence, the RF portion of the instrument, in block form, remains unchanged. Figures 2 and 3 show the 202H and 202J, respectively, in block form.

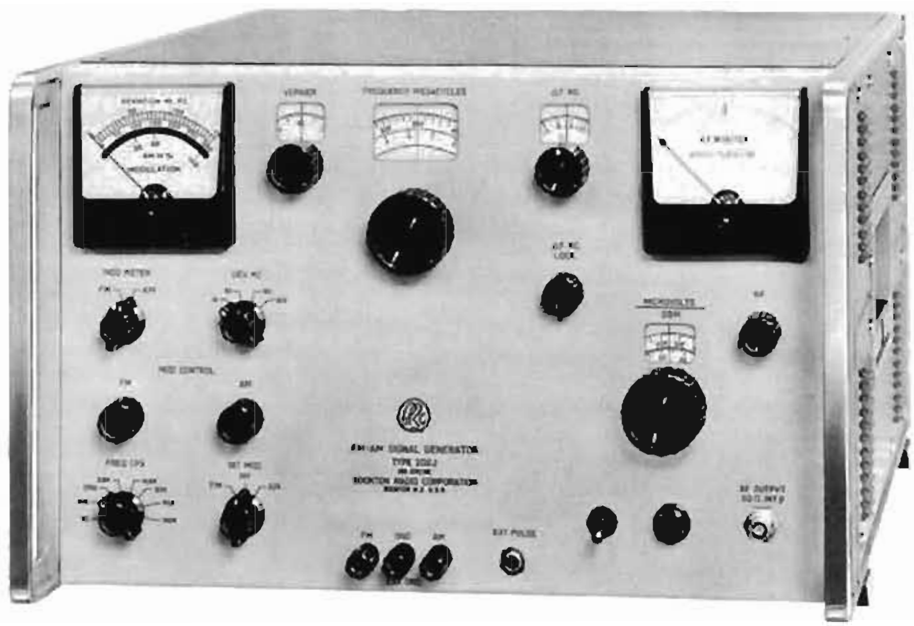

Figure 1. Type 202J Signal Generator

FM Linearity

By giving routine, but careful, attention to the oscillator and reactance tube circuits (described under design Considerations for the Oscillator and Modulator), a linearity of 1½% (equivalent to ¾% total harmonic distortion) at 150 kc deviation for the 202J was achieved. At 300 kc deviation, the FM linearity is 5%. The 202H FM characteristic yields a demodulated output with less than 1% THD at 75 kc deviation. At 100 Mc carrier and 75 kc deviation, the 202H introduces ½% THD.

All of the above numbers are specified limits, and typical performance is consistently better.

Electronic Vernier Tuning

A new system of electronic vernier tuning has been incorporated into both

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the new 202H and 202J to permit relatively small calibrated changes in output frequency. This system operates by applying dc voltage to the grid of the reactance tube through a precision potentiometer. The total range of the control is ±40kc. The calibrated dial covers the range of ±30kc in 1kc increments and may be slipped against the potentiometer shaft by operating a dial lock mechanism. A jack is also provided for inserting external dc voltage for use in connection with an external X-Y plotter or frequency control system.

Microphonics and Vibration

Potentially, vibration and sound are two of the main sources of disturbance in low deviation measurements. Design of the 202H and J is aimed at alleviating this problem. Figure 4 shows the mounting of an RF unit in the 202J and H. The unit is supported by four vibration absorbing mounts. Bellows-type, flexible couplings isolate the shafts of the main tuning capacitor and of the attenuator's piston from the front panel. The net result is that the FM sensitivity of the 202H and J to acoustic or mechanical impulses, applied to the front panel or to the mounting hardware, is five times less than that of the 202E and G. Thus, the short-term frequency stability is much improved in all but the most quiet environments.

Automatic Level Set

The modern tubes used in the amplifier and doublers of the RF unit provide enough reserve RF power output to operate a level control circuit. The output meter remains within ±2% of the "red line" setting (for 0.2 volts maximum output) across the frequency band.

Power Supply

The new generators contain regulated dc power supplies (both plate and fila-

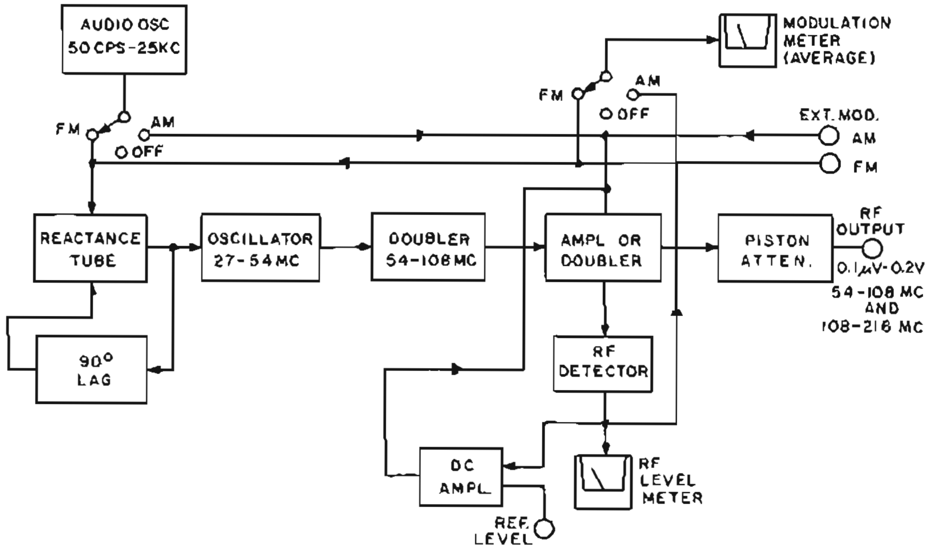

Figure 2. Block Diagram of Type 202H

ment) for all tubes in the RF unit. As a result of these regulated supplies the residual FM at line frequency in the 202J is less than half as much as it is in the 202G.

In order that the entire instrument be contained in a single package without excessive heating of the RF unit by the power supplies, all active components of the power supplies are semiconductors.

There are three basic improvements in the 202H over the 202E supply.

1. The power supply and signal generator are contained in a single cabinet.
2. Both supplies are more stable (by a factor of 5), with respect to line voltage fluctuations, than are their counterparts in the 202E.
3. The 202H power supply contains neither a ballast tube nor a VR tube, both of which components are vulnerable.

FM Bandwidth of 202J

A need for wide bandwidth arises particularly in PCM telemetering systems. The maximum IRIG Telemetering standard bit rate is 330kc in the VHF band, and the receiver IF bandwidth requirement for this bit rate is about 500 kc. Hence, the bandwidth of an FM Signal Generator for checking a receiver's ability to handle such signals should be more than 500 kc. The 202J has a 1 mc FM bandwidth.

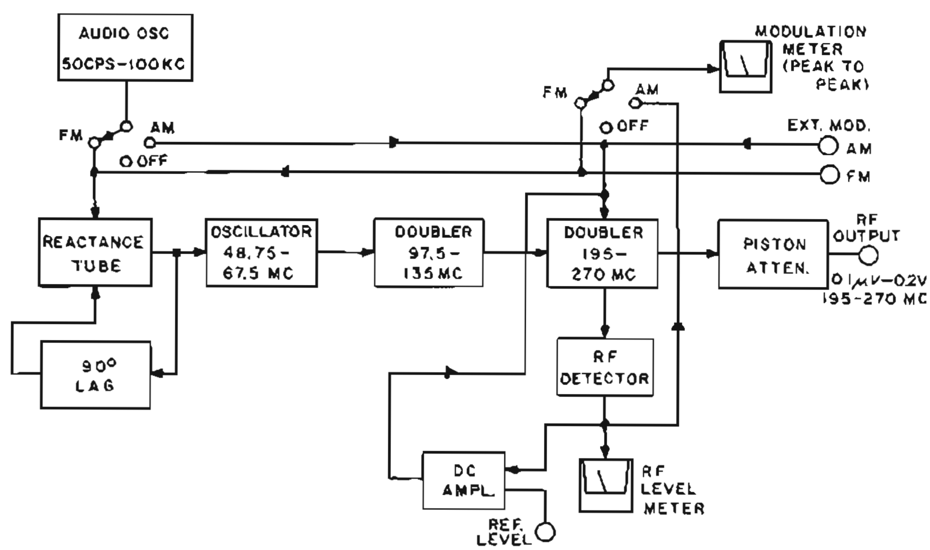

Figure 3. Block Diagram of Type 202J

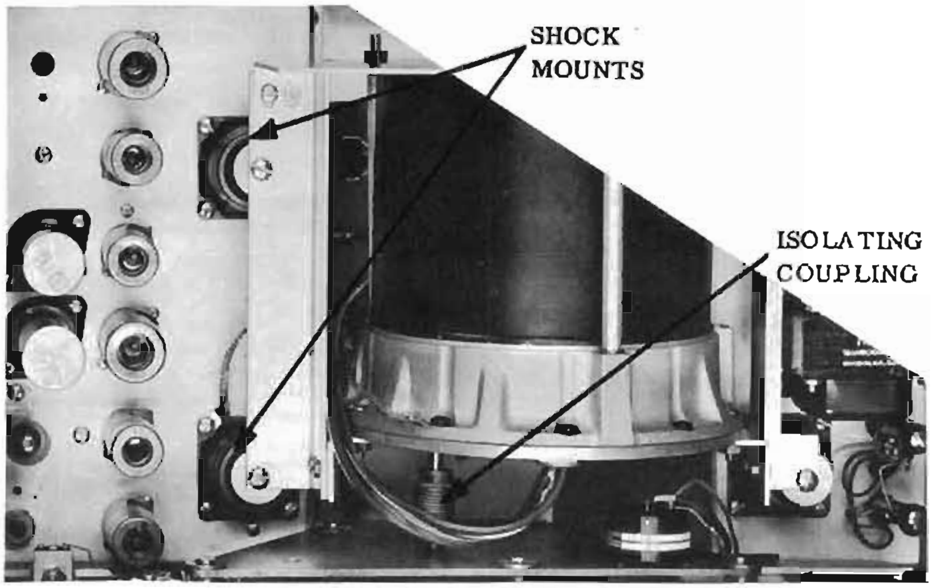

Figure 4. Type 202J — Top View

Peak Reading Modulation Meter Circuit for 202J

The modulation meter circuit is entirely new for the 202J. It reads the peak-to-peak deviation for all modulation waveforms which do not have important frequency components outside the pass band of 10 cps to 1 mc. The circuit consists of a two-stage, feedback amplifier, followed by a peak-to-peak voltmeter circuit.

The 202H has the same kind of averaging meter circuit as the 202E it replaces, with two exceptions: the diodes in the 202H meter circuit are more stable, and the dependence of the meter reading stability on diode stability has been reduced by amplifying the signals to be measured before metering them.

In order to ensure a net gain in overall meter accuracy, the amplifier has been heavily stabilized by negative feedback.

Design Considerations for the Oscillator and Modulator

The basic BRC 202 frequency modulator is now, as it always has been, a reactance tube with a bridged-T phase shift network. One of the merits of this type of network is that it allows relatively little variation of deviation sensitivity with carrier frequency. The small variation in FM sensitivity is diminished by ganging one of the passive elements in the phase shift network to the tuning capacitor in the oscillator. An analysis of the action of the bridged-T and other phase shift networks used with reactance tube frequency modulators has been given by M. Crosby and D. Hill.[3] Within this framework, the improved FM linearity in the 202H and J has been achieved with no loss of constancy of FM sensitivity or carrier frequency stability.

In the process of developing an FM oscillator, one must be sure that, at every frequency in the tuning band, the oscillator operates at a frequency fairly close to resonance of the LC circuit. If, at some frequency in the band, the oscillator departs from this frequency, then the impedance of the LC circuit contains an equivalent parallel reactance at that frequency. Of course, this kind of action does not prevent us from frequency modulating the oscillator, but it does result in rather violent changes in FM sensitivity at some points in the band.

Occasionally one finds a carrier frequency close enough to a point of discontinuity to be traversed in the modulation cycle. We still modulate the inductance of the LC circuit as planned, but the departure of operating frequency from resonance is now varying during the cycle. This is not the kind of behavior one would care to make allowances for. Therefore, great pains were taken to design the oscillator-reactance tube combintaion such that these effects were minimized. The oscillator was loaded as lightly as practicable by the phase shift network, for example.

In the interest of FM linearity, the oscillator level is rather low so that its phase shifted output cannot drive the reactance tube outside of its linear region.

Tube type 6688 was chosen for the reactance tube and doubler stages, primarily because of its high stability and high trans-conductance. It has the added merit of low input conductance in the VHF band.

New Packaging

Both of these new signal generators are packaged in a restyled cabinet which can be readily rack-mounted. The front panel RF output jack can also be readily installed in a mounting hole, provided in the rear cabinet panel, for rack-mounted applications. Convenient carrying handles are provided on the sides of the cabinet in addition to the standard rack handles integral with the side frame castings.

Specifications

The following specifications apply to both the Types 202J and 202H, unless otherwise indicated.

RADIO FREQUENCY CHARACTERISTICS

RF Range: 195-270 MC (202J) 54-216 MC (202H)
No. Bands: 1 (202J) 2 (202H)
Band Ranges: 195-270 MC (202J) 54-108 MC, 108-216 MC (202H)
RF Accuracy:
Main Dial: ±0.5%*
Electronic Vernier: ±(10% +1 KC)*
*after one hour warm-up
RF Calibration:
Main Dial: Increments of 0.5 MC (202J and 54-108 MC on 202H)
Increments of 1.0 MC (108-216 MC on 202H)
Mechanical Vernier: 2200 divisions through range (202J)
2300 divisions through range (202H)
Electronic Vernier: Increments of 1 KC over ±30 KC range*
*total range ±40 KC; provision for slipping dial to place "0" at a specific frequency
RF Stability:
0.02% per hour* (202J)
0.01% per hour* (202H)
*after two hour warm-up
RF Output:
Range: 0.1 µv to 0.2 volts*
*across external 50 ohm load at panel jack
Accuracy: ±10%, 0.1 µv to 50 K µv
±20%, 50 K µv to 0.2 volts
Auto Level Set: holds RF monitor meter to "red line" over band
Impedance: 50 ohms
VSWR: 1.2
Spurious Output: All spurious RF output voltages are at least 25 db below desired fundamental on 202J (30 db on 202H)
RF Leakage: Sufficiently low to permit measurements at 0.1 µv

AMPLITUDE MODULATION CHARACTERISTICS

AM Range:
Internal: 0-50%
External: 0-100%
AM Accuracy: ±10% at 30% and 50% AM
AM Calibration: 30, 50, 100%
AM Distortion: 5% at 30%
8% at 50%
20% at 100%
AM Fidelity: ±1 db, 30 cps to 200 KC

FREQUENCY MODULATION CHARACTERISTICS

FM Range:
Internal: 0-300 KC in 4 ranges (202J)
0-250 KC in 4 ranges (202H)
External: 0-300 KC in 4 ranges (202J)
0-250 KC in 4 ranges (202H)
FM Accuracy: ±5% of full-scale*
*indication proportional to peak-to-peak (202J) and sine-wave (202H) of modulating waveform

FM Calibration:

202J
- 0-15 KC in increments of 0.5 KC
- 0-30 KC in increments of 1 KC
- 0-150 KC in increments of 5 KC
- 0-300 KC in increments of 10 KC

202H
- 0-7.5 KC in increments of 0.5 KC
- 0-25 KC in increments of 1 KC
- 0-75 KC in increments of 5 KC
- 0-250 KC in increments of 10 KC

FM Non-Linearity*: 1.5% at 150 KC, 5% at 300 KC (202J)
*"least squares" departure from straight line passing through origin

FM Distortion: 0.5% at 75 KC (100 MC and 400 cps modulation only)
(202H) 1% at 75 KC (54-216 MC)
10% at 240 KC (54-216 MC)

FM Bandwidth: ±3 db, 3 cps to 1 MC (202J)

FM Fidelity: ±1 db, 5 cps to 500 KC (202J)
±1 db 5 cps to 200 KC (202H)

Spurious FM: Total RMS spurious FM from 60 cps power source is at least 60 db below 150 KC (202J)

Signal-to-noise Ratio: 60 db below 10 KC (202H)

Microphonism: Extremely low; shock-mounted RF unit

External FM Requirements: 1 volt RMS into 100K ohms for 150 KC deviation

PULSE MODULATION CHARACTERISTICS

PM Source: External
PM Rise Time: 0.25 μsec
PM Fall Time: 0.8 μsec

MODULATING OSCILLATOR CHARACTERISTICS

OSC Frequency:

202J		202H	
50 cps	10.5 KC	50 cps	7.5 KC
400 cps	30 KC	400 cps	10 KC
1730 cps	70 KC	1000 cps	15 KC
3900 cps	100 KC	3000 cps	25 KC

OSC Accuracy: ±5%
OSC Distortion: 0.5%
OSC External Output:
30 volts approx. at external FM posts
30 volts approx. at external AM posts

ACCESSORIES

Furnished: Type 502-B Patching Cable
Available: Type 207-G Univerter (202J)
Type 207-E Univerter (202H)
Type 501-B Output Cable
Type 504-A Adapter
Type 505-B Attenuator
Type 506-B Patching Cable
Type 507-B Adapter
Type 508-B Adapter
Type 509-B Attenuator
Type 510-B Attenuator
Type 514-B Output Cable
Type 517-B Output Cable

TUBE COMPLEMENT

Tubes	Transistors	Diodes	Zener Diodes
3—6688	1—2N1008	2—1N660	6—G31A-7H
1—6AF4	1—2N1136B	2—1N1763	1—G31H-56L
1—6AW8	1—2N1136	4—1N1764	1—G31G-7H
3—6AU6	2—2N1379	2—1N1581	2—G31A-12H
1—6BK7		1—51029	1—G31G-47L
1—6AQ5			
1—12AU7			
1—6DJ8			

PHYSICAL CHARACTERISTICS

Mounting: Cabinet for bench use; readily adoptable for 19" rack mounting
Finish: Gray engraved panel; green cabinet (other finishes available on special order)
Dimensions: Height: 10⅜" Width: 16¾" Depth, 18⅜"
Weight: Net: 45 lbs.

POWER REQUIREMENTS

105-125/210-250 volts, 50-60 cps, 100 watts

References

1. John P. Van Duyne, "A Modulator for the New FM Stereo System," BRC Notebook Number 30.
2. M. G. Crosby and D. M. Hill, "Design of FM Signal Generator," Electronics, Nov., 1946, pages 96-101.

EXTERNAL OR "IN CIRCUIT" MEASUREMENTS ON THE UHF Q METER

CHARLES W. QUINN, *Sales Engineer*

Detailed information about the design and theory of operation of the UHF Q Meter Type 280-A is given in Notebook Number 27. Conventional applications are covered in detail in Notebook Number 28. An article concerning calibration of the instrument appears in Notebook Number 29. This article will deal specifically with the "unconventional" external or "in circuit" applications.

Basic Theory of Measurement

Briefly, the UHF Q Meter utilizes the bandwidth relationship $Q = \frac{f_0}{\Delta f}$ (for $Q \geqq 10$) to determine Q, as shown in Figure 1. A resonance indicating meter is used to determine the peak of the resonance curve and to resolve the half-power (.707V) points. This means that the UHF Q Meter reads Q in terms of frequency; the frequency being determined by the ability of the instrument to measure a *relative* amplitude change of 3db. The absolute value of V in Figure 1 is of no consequence and may

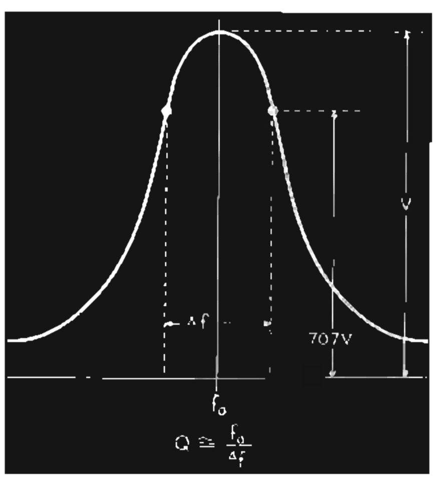

Figure 1. Q Resonance Curve

vary over a few decades (depending on the coupling of the probes and gain) without affecting the ability of the instrument to read *circuit Q*.

Type 580-A Probe Kit

In order to provide a convenient means of coupling into the external resonators or circuits, BRC has designed a probe kit (Type 580-A) which is suitable for many external measurements. The kit consists essentially of an injection probe and a detection probe, designed for coupling the external circuits to the output of the oscillator and the input to the chopper amplifier in the 280-A. For external measurements, the Q capacitor and associated coupling circuits in the UHF Q Meter are disconnected at the rear of the instrument and the injection and detection probes, furnished with the probe kit, are connected in their place. A basic block diagram of the UHF Q Meter, connected for external measurements, is shown in Figure 2.

Methods of Coupling to Test Circuits

Before discussing some of the techniques for "in circuit" measurements, it would be well to point out two basic requirements which should be met by the external circuit to be measured. First, the circuit must be resonant in the fre-

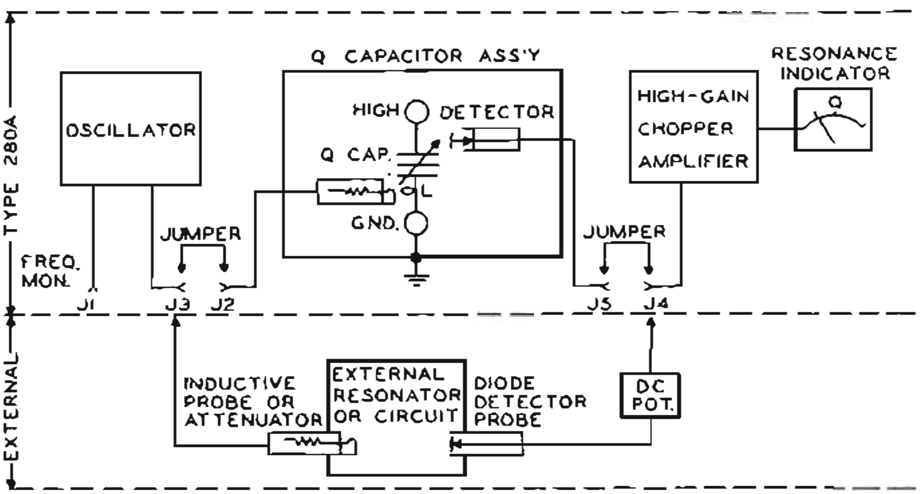

Figure 2. Block Diagram of UHF Q Meter Showing Connections For External Measurements

quency range of the 280-A (210 to 610 Mc). Secondly, the Q of the circuit must range from 10 to 25,000.

Figure 3A shows a typical amplifier configurtaion with a tuned coaxial resonator used as a plate load. This circuit may be measured using both the injection and detection probes in the Type 580-A Probe Kit, since it has access holes large enough for insertion of both probes. A preselector resonator could be similarly measured.

Figure 3B demonstrates another technique where an existing or "built in" loop is use for injection into the circuit to be measured and the 580-A detection probe is inserted into an access hole. In this case, care must be taken that the 280-A oscillator load does not exceed a 1.2 VSWR, referred to 50 ohms.

In Figure 3C the 580-A Probe Kit is not used. Existing or "built in" injection and detection circuits are connected directly to the 280-A oscillator and chopper amplifier circuits. When using this technique, the law of the detector used must be evaluated and taken into account. This may be accomplished by connecting a signal generator to the detector and using the variable attenuator in the signal generator to set up the 3db point on the 280-A resonance indicating meter. A precision 3db attenuator, connected in series with the signal generator, may be used for a more precise check of the detector.

Figure 3D illustrates still another technique for measuring an external circuit. The 280-A oscillator is connected to a tube input circuit, which may be "hot" or "cold", and the detection probe is connected at the output connector. This type connection is especially de-

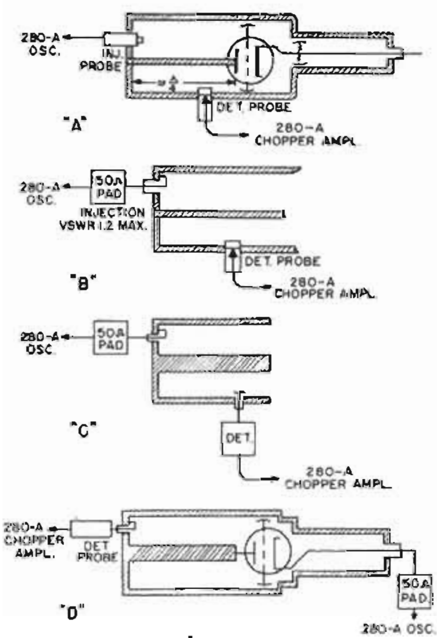

Figure 3. Coupling Techniques

sirable where it is necessary to evaluate the effects of dynamic loading. For this type measurement the input circuit should be relatively broad.

From the foregoing, it can be seen that there are many techniques which may be used to couple into the external test circuits for performing "in circuit" measurements on the 280-A; serving an extremely broad range of applications.

Resonator Measurements

The configurations of resonators in the frequency range of the 280-A are varied. A few of these configurations are shown in Figure 4. When performing resonator measurements it should be remembered that the resonators should be coupled to the 280-A at points which provide optimum coupling, wtih minimum loading by the 280-A. Basically, magnetic coupling is optimum at the Voltage Node point, or point of maximum current; i.e., the low Z point. Detection is optimum at the Current Node point or the point of maximum voltage; i.e., the high Z point (Figure 4).

Test Circuit Loading

Minimum loading is accomplished with the detector coupled as "loosely" as practical. The injection circuit, on the other hand, may be coupled much "tighter" without loading. The extent of loading can be evaluated by making a series of Q measurements at different sensitivity levels and probe spacings. Higher Q readings are indicative of negligible loading. The measurement should, therefore, be made at the minimum sensitivity level at which negligible loading occurs.

Extension of L Range

It is interesting to note that external measurements permit the inductance range of the 280-A to be extended beyond the specified 2.5 to 146 mμh range of the internal resonating capacitor. Referring to Figure 4G, if C were known, and could be adjusted to a value less than 4 pf (the minimum capacitance of the 280-A Q capacitor), the inductance range of the instrument could be extended to as much as 2 μh. This may be accomplished using a high quality, small variable capacitor with a range of approximately 0.2 to 3 pf and a Q of 200 or more. The 580-A probes are connected as shown in Figure 2 for external measurements. Either of the following techniques may then be used. The capacitance required to resonate the inductor could be estimated and the variable capacitor set for this value, using the calibrated Q capacitor. C_1 and Q_1 would then be measured. Then, with the coil and capacitor placed approximately as shown in Figure 4G, the resonant frequency and circuit Q (Q_c) would be determined. The inductance of the coil would then be computed for the measurement frequency using the following equations:

$$L = \frac{1}{\omega^2 C_1} \qquad Q = \frac{Q_1 Q_c}{Q_1 - Q_c}$$

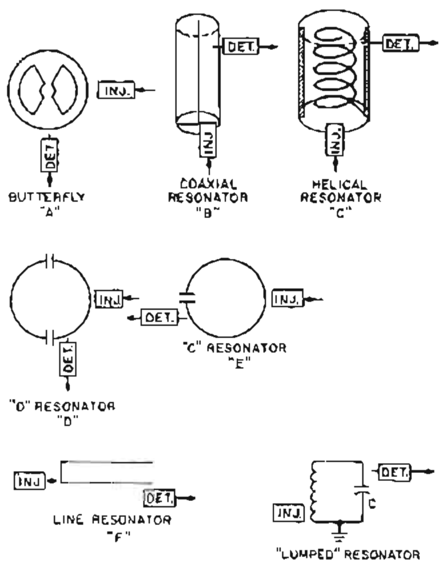

Figure 4. Typical Resonator Configurations

Another technique for extending the inductance range of the 280-A involves adjusting the variable capacitor until the desired resonant frequency is indicated by the 280-A. Circuit Q (Q_c) is then measured. The variable capacitance (C) and Q are then measured on the 280-A and the above equations are used to determine L and Q.

Test Circuit Shielding

When small or unshielded resonators are to be measured, it is often desirable to make use of shielding to minimize "hand capacitance" and radiation effects. This shielding may be in the form of a "work box" with built in supports for the 580-A probes, as shown in Figure 5. A setup of this type would be useful for evaluating the inductor mentioned in the previous paragraph.

External Resonators as Jigs and Fixtures

Resonators, in various forms, have applications in many fields. The "D" resonator (Figure 4D), because of its peculiar magnetic field, is used in the investigation of molecular resonances in the field of basic research to improve our understanding of materials. "C" type resonators could be used for the measurement of dielectrics where the specimen is inserted into the gap that forms the resonating capacitance. Helical resonators are used to investigate the effects of ionization of gases in the ion propulsion field. The helical resonator will also be useful in the evaluation of micro-

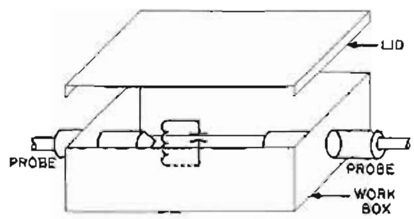

Figure 5. "Work Box" used as Shield for Resonator Measurements

wave varactors, diodes, and other semiconductors, where capacitance values are low and Q values are high.

The reentrant cavity or coaxial resonator is probably one of the most versatile resonators, since its performance can be readily calculated. It can serve as a fixture for evaluating dielectrics (Figure 6A) or magnetics (Figure 6B) in the frequency range of the 280-A (210 to 610 Mc).

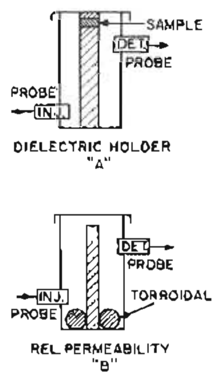

Figure 6.

Measurements Under Simulated Environmental Conditions

The coaxial and other type resonators may be used with the 580-A Probe Kit to provide thermal isolation between the specimen and the instrument. This would be of special interest to research and development people involved with temperature measurements.

Extension Probes for Small or Limited Access Resonators

The 580-A Probe Kit utilizes a telescoping sleeve principle. The outside diameter of the inner sleeve is 0.430 inch. The outside diameter of the outer sleeve is 0.500 inch. These diameters may be too large for some special applications, either because of the physical size or the effect this size would have on the circuit under test. For high Q circuits (100 or more), this situation can be remedied by fabricating extensions for the probes, similar to the samples shown in Figure 7A. This would permit measurements where the diameter of the access holes would be limited only by the diameter of the coaxial cable used.

Note that the length of these probe extensions must be such that their resonant frequency is well above the resonant frequency of the resonator being measured. Resonance of the probes may be checked by bringing them close to each other, with the external resonator removed, and sweeping the frequency through the point of measurement. If the probe extensions are functioning properly, no sharp slope in output indications should be observed.

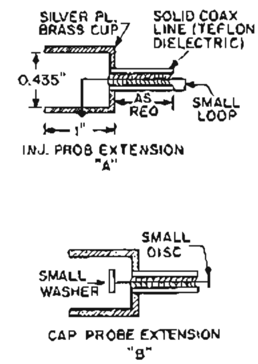

Figure 7. Examples of Probe Extensions

Conclusion

An attempt has been made here to point up some of the techniques which could be used for performing external measurements of circuits and resonators with the Type 280-A UHF Q Meter. Doubtless there are many applications which have not been touched upon or that should be expanded upon. It is our intension to delve more into dielectric measurements, low and high temperature techniques, semi-conductor measurements, and magnetics, in future Notebook articles. In the meantime, we here at BRC would appreciate hearing from any of our customers who have measurement problems in this area or who have evolved new measurement techniques which could be applied to this versatile instrument.

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SERVICE NOTE

Modification of Type 250-A for Reduced Signal Level and Increased Sensitivity

Reducing Signal Level

An improved method has been devised for reducing the signal level at the terminals of the Type 250-A RX Meter for special applications where a low signal level is required. In the past, the signal level was lowered by inserting a 100K ohm potentiometer in the signal oscillator plate supply, as described on pages 16 through 18 of the 250-A instruction manual. This provided a means for varying the oscillator plate voltage, thus reducing the signal oscillator output level. This system works satisfactorily for the most part, but occasionally a plate voltage will be selected which will "shut-off" the oscillator.

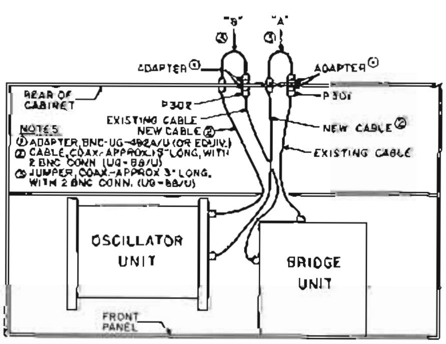


Figure 1. Modification for Reduced Signal Level

The new method provides for the connection of a fixed or variable external attenuator (or attenuators) in series with the RF signal source to the bridge. In order to provide a convenient means for connecting the attenuators in the signal circuits, the bridge and oscillator connections are made accessible at the rear of the instrument cabinet. These connections are jumpered for normal operation. For special applications, the jumpers are removed and replaced with fixed or variable attenuators. In most cases, the attenuator is connected in place of jumper "A" (Figure 1). In some instances, because of mixing in the test component, it would be desirable to attenuate both signals.

The modification, together with a list of the parts required, is shown in Figure 1.

Increasing Sensitivity

When the signal level to the bridge is reduced for special applications, as described above, increased detector sensitivity is often desirable. The sensitivity of the 250-A may be significantly increased by modifying the instrument as shown in Figure 2.[1] The signal from IF amplifier V202(6AG5) is connected to a jack at the rear of the 250-A cabinet. A VTVM, connected to this jack, amplifies the signal and serves as a null indicator which has improved sensitivity over the null indicator on the front panel of the 250-A. This improves the resolution of the C_p and R_p dials and also results in improved resolution of the R_p or X_p parameter which is occasionally the minor impedance in a measurement. (Minor impedance is defined as that impedance which contributes least to the amplitude and phase of the current in an RF circuit.)

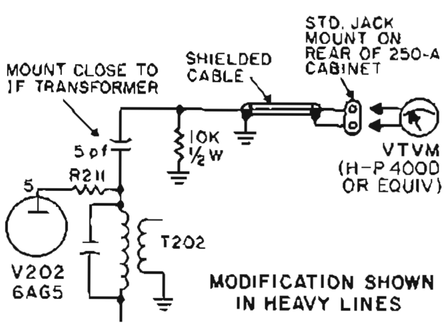


Figure 2. Modification for Increased Sensitivity

For best results, and to obtain optimum sensitivity, it is necessary to select mixer tube V101 (6AB4) for minimum noise deflection (preferably less than one division on the VTVM).

[1]This modification was suggested by H. Thanos of R.C.A., Somerville, N. J.

BRC DEDICATES NEW PLANT

Friday, October 20, 1961 was a perfect day for the dedication of the new BRC plant. Nestled in the Rockaway Valley, against a backdrop of high hills ablaze with autumn-painted trees, the new plant was the center of scenic splendor.

On hand for the dedication ceremony were BRC employees; friends and business associates of BRC; members of the local, county, and state government; and executives from the Hewlett-Packard (BRC's parent company) family.

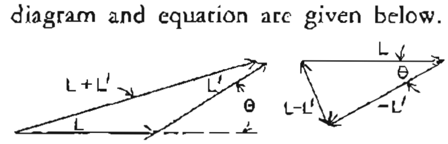
The dedication ceremony was presided over by Dr George Downsbrough. President of BRC. Guest speaker was Mr. David Packard, President of Hewlett-Packard. Mr. A. R. Post, Chief, Bureau of Commerce for the State of New Jersey, delivered a message of congratulations from the Governor of New Jersey.

A "ribbon-cutting" ceremony was held at the main entrance to the new plant, with cutting honors going to Mr. John Vandermark, Mayor of Rockaway Township.

Mr. John Vandermark, Mayor of Rockaway Township, N. J., cuts ribbon to officially open the new BRC plant as Dr. George A. Downsbrough and Mr. David Packard look on.

CORRECTION

The vector diagram and accompanying equation in Figure 2 of the article entitled, "A Modulator for the New FM Stereo System," published in Notebook Number 30, are not correct. The correct diagram and equation are given below.



$$R_s = 20 \log \left(\frac{\overline{L+L'}}{\overline{L-L'}} \right)$$

EDITOR'S NOTE

BRC to Show Four New Instruments at IRE

(VISIT BOOTHS 3101-3102)

This year, at the IRE show, BRC will show four new instruments: the new Types 202-H and 202-J FM-AM Signal Generators, the new Type 230-A Signal Generator Power Amplifier, and the new Type 219-A FM Stereo Modulator.

The Type 202-H signal generator covers the frequency range of 54 to 216 Mc and replaces the Type 202-E. The 202-J blankets the 195 to 270 Mc telemetering range and replaces the 202-G. Improvements in these instruments include: improved FM linearity, automatic RF level set, electronic vernier tuning, increased FM modulation bandwidth, improved FM deviation metering, reduced FM microphonism, a completely solid-state power supply, and a completely redesigned cabinet.

The 230-A Signal Generator Power Amplifier provides a means of increasing the RF power output of conventional signal generators up to 4 watts or +6 dbw (14 volts rms into 50 ohms), in the frequency range of 10 to 500 Mc.

The 219-A FM Stereo Modulator provides the stereo modulation outputs, as specified in FCC Docket 13506, suitable for modulating FM signal generators, such as the BRC Type 202-E or 202-H.

A "Guess the Q" contest, which seems to have become traditional with BRC, will be a feature at the BRC booth again this year. Our engineers have been hard at work devising a "guess-defying" Q circuit that should be a delight and a challenge to booth visitors.

Visit booths numbers 3101 and 3102. See and hear more about our new instruments, and have the BRC engineers on duty help you with your test instrument application problems.

Photograph taken at a previous IRE show demonstrates that the BRC "mystery coils" put all of the Q contest entrants in a pensive mood.

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The NOTEBOOK

BOONTON RADIO CORPORATION · BOONTON, NEW JERSEY

APPLICATIONS OF THE SIGNAL GENERATOR POWER AMPLIFIER

CHARLES W. QUINN, *Sales Engineer*

INTRODUCTION

At first glance the application of an RF Power Amplifier appears limited or obvious. While there are obvious uses, there are many applications which are not apparent. It is the purpose of this article to enumerate and discuss both application categories. The specific requirements for the tests to be described are as many and as varied as the systems involved. For this reason, this article will be limited to a general description and/or example of each test. Detailed information on radio receiver tests may be found in References 1 and 2, at the end of the article. Procedures for radio frequency interference (RFI) testing are given in References 3 and 4. Before dealing directly with applications, let us look at the specifications of the Type 230A Signal Generator Power Amplifier (Figure 1) and clarify them where necessary.

Figure 1. Type 230A Signal Generator Power Amplifier

SPECIFICATIONS

RADIO FREQUENCY CHARACTERISTICS

RF RANGE

Total Range: 10 to 500 mc

No. Bands: 6

Band Ranges: 10-18.5 mc, 18.5-35 mc, 35-65 mc, 65-125 mc, 125-250 mc, 250-500 mc

RF Calibration: Increments of approximately 10%, accurate to ±10%

RF OUTPUT

Range: Up to 15 volts*. *Across external 50 ohm load

Range: Up to 15 volts*
*Across external 50 ohm load

Calibration: 0.2 to 3 volts f.s; increments of approx 5%
1.0 to 10 volts f.s; increments of approx 5%
2.0 to 30 volts f.s; increments of approx 5%

Accuracy: ±1.0db of f.s. (10-250mc)
±1.5db of f.s. (250-500mc)

Leakage: Sufficiently low to permit measurements at 0.1 volts.

RF Bandwidth:* > 700kc (10-150mc)
> 1.4mc (150-500mc)
*Frequency interval between points 3db down from max. response.

RF INPUT

Level: ≦ 0.316 volts* (30db gain) (10-125mc)
≦ 0.446 volts* (27db gain) (125-250mc)
≦ 0.630 volts* (24db gain) (250-500mc)
*for 10 volts output into 50 ohms

Impedance: 50 ohms

AMPLITUDE MODULATION CHARACTERISTICS

AM RANGE: Reproduces modulation of driving signal generator 0-100%*

AM DISTORTION: <10% added to distortion of driving Signal Generators*
*Up to 5 volt max. carrier output for up to 100% AM

FREQUENCY MODULATION CHARACTERISTICS

FM RANGE: Reproduces modulation of driving Signal Generator except as limited by the RF bandwidth.

INCIDENTAL AM: <10%* added to modulation of driving Signal Generator
*At 150kc deviation.

FM DISTORTION: Negligible distortion added to distortion of driving Signal Generator for deviations and modulation frequencies <150kc.

PHYSICAL CHARACTERISTICS

MOUNTING: Cabinet for bench use; by removal of extruded strips suitable for 19-inch rack mounting.

FINISH: Gray wrinkle, engraved panel (other finishes available on special order).

DIMENSIONS: Height — 7-3/16"
Width — 19-1/2"
Depth — 17-11/16"

WEIGHT: Net: 37 lbs.
Gross Export: 75 lbs.
Gross Domestic: 45 lbs.
Legal Export: 43 lbs.

POWER REQUIREMENTS

230-A: 105-125/210-250 volts, 50—60 cps, 150 watts.

YOU WILL FIND . . .

SIGNAL SOURCES

Virtually any signal source, within the frequency range of the 230A, may be used. The obvious sources are signal generators such as the BRC 225A general purpose signal generator, the 202 series FM-AM signal generators, and the 211A and 232A Navigation Aid signal generators. Not so obvious, but nevertheless convenient signal sources, are the 260A and 280A Q Meters, the 250A RX Meter, crystal oscillators, etc.

AMPLIFIER, RECEIVER, AND SYSTEM TESTING

Amplifier, receiver, or system testing may take many forms. A few of these tests are:

1. Overload tests.
2. AGC characteristics.
3. Skirt selectivity.
4. Adjacent channel desensitization.
5. Cross modulation and intermodulation.
6. Image and IF rejection.

Connections for tests 1, 2, 3, and 6 are shown in Figure 2.

Overload Tests

Overload tests are made to determine the input level at which the output of

Figure 2. Setup for Receiver Testing

the unit under test departs from a specified characteristic; i.e., linear, log-linear, etc., by a specified tolerance. Overload tests are usually made on circuits with active elements and are not restricted to the conventional superheterodyne receiver. Single frequency and broadband amplifiers are also tested for overload.

The output of the unit under test may be the input signal amplified, or the demodulated output (AM, FM pulse, etc.), or a voltage or current proportional to input, (analog digital, dc, etc.).

AGC Characteristics

AGC (automatic gain control) characteristics are measured or determined by measuring the relationship between RF input voltage and the dc voltage bias developed by the AGC detector. It is often desirable to determine the RF level which will override the AGC and cause blocking and/or distortion. This level is often much higher than 500,000 microvolts in well-designed systems. See Reference 1 for measurement details.

Skirt Selectivity

Skirt selectivity testing of a communications system requires that the performance of the frequency selective circuits be determined at a frequency considerably removed from the desired frequency, or on the "skirts" of the resonance curve, where attenuation is at a very high value. Typical values are 2 to 5 volts for attenuation figures of 80 to 120 db. In this test, one must be ever cautious of the possibility of overload occurring before the desired point on the skirt is reached. In AM systems, an increase in distortion indicates that overloading has taken place and limits the extent of the "skirt" measurement.[2]

Adjacent Channel Desensitization Test

Most communication centers transmit on many channels simultaneously. Usually, a given receiver will be in contact with signals of less than 100 microvolts in strength, while one or more transmitters in the same room are operating at a frequency only a few channels from the receiver frequency. The receiver must not, therefore, be affected by strong signals in adjacent channels. It is for this reason that desensitization characteristics are specified by communication system designers, and that desensitization tests are made.

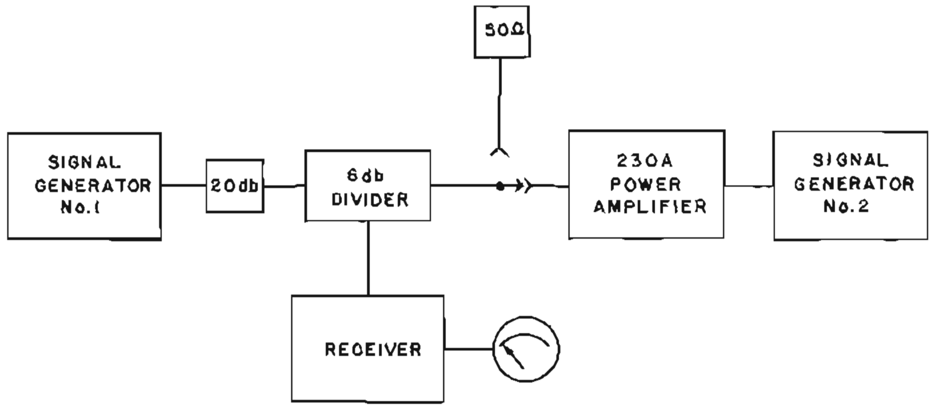

Figure 3. Connections for Desensitization Test

Desensitization tests are made by connecting the equipment as shown in Figure 3. Signal generator #1 is set to give a convenient metered detector level (sometimes specified for a given system). This is the "desired signal" on channel. Using signal generator #2 in conjunction with the 230A Power Amplifier, the adjacent channel level is raised until the detector level is reduced by a specific amount (usually 3 db). The reading on the 230A voltmeter is twice the voltage required for "desensitization."

Cross Modulation and Intermodulation

Cross modulation and intermodulation tests are made on systems which are inherently very linear. Intermodulation tests are performed by supplying two or more signals to a system and measuring the resultant spurious products. For example, two 15-Mc signals, spaced 1 kc apart, will produce a spectrum (Figure 4) which can be analyzed to determine the amount of intermodulation.

There are two approaches to this test depending upon the amount of intermodulation expected. If the expected intermodulation is greater than 2%, a single 230A amplifier may be used; connected as shown in Figure 5.

Typical intermodulation performance data, taken with the 230A connected as in Figure 5, is given in the table in Figure 6. The test unit was replaced by a 50-ohm termination and measurements were made at 15 Mc.

The data in Figure 6 is indicative of the intermodulation present in the 230A Power Amplifier and expresses its limits for given output levels. Column "V_1/V_2 shows the value of the rms amplitude of each signal. Column "V_T" shows the meter indication when both signals are applied. The "db" column indicates the level of the highest spurious signal produced. It is the "db" column which is most significant. For instance, a figure of 46 db is typical of a 0.5% product; consequently, system intermodulation products of less than this figure will have little or no significance. Actually, any change in the spurious output detected, when using a passive linear termination, indicates a departure from linearity or phase shift. However, evaluation of absolute value is impractical.

There is another approach to the measurement of small amounts of intermodulation, which, while not tested to date, theoretically should extend this measurement to another order of magnitude. The connections for this technique are shown in Figure 7. (The meter switch is set to the "off" position for this application.)

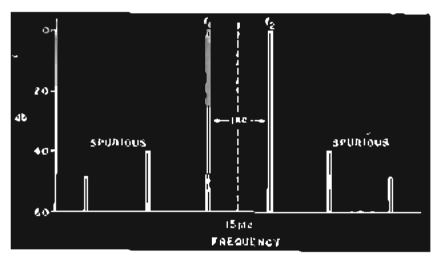

Figure 4. Spectrum for Analyzing Amount of Intermodulation

This system virtually eliminates the intermodulation products present or generated in the driver stages of the 230A Power Amplifier. It also reduces the nonlinear effects of the dynamic plate resistance of the output stage by loading it with considerable linear, passive resistance. It is estimated that a 10 db

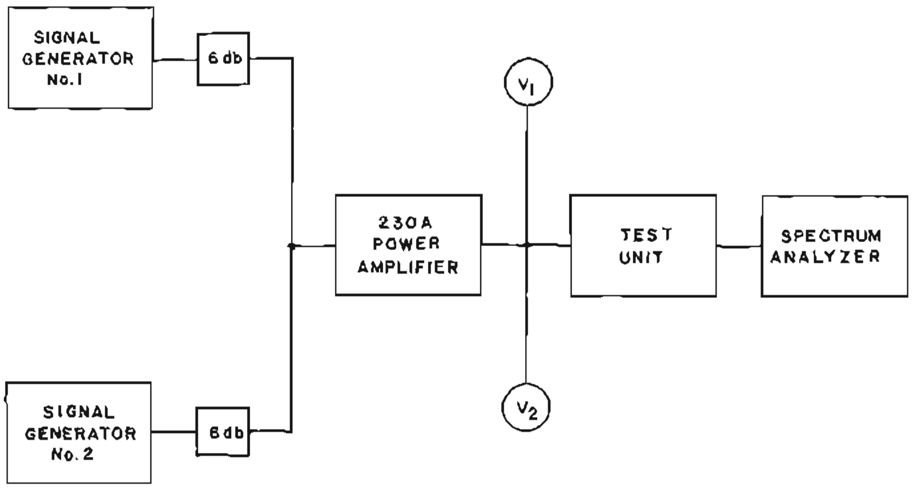

Figure 5. Connections for Checking More than 2% Intermodulation

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coupling attenuator will be a good factor for V_1/V_2 values of 1 to 3 volts. Intermodulation of 50 to 60 db should be obtainable using this technique.

For cross modulation tests, two signals are required; connected as shown in Figure 3. Signal generator #1 is set to a prescribed RF level (E_1 and percent modulation) depending upon the system being tested. The demodulated output is noted. Signal generator #2 and the 230A Power Amplifier are then connected and set on an adjacent channel in accordance with the specific test to be made. Signal generator #2 is modulated in the same manner as signal generator #1, which is now set for CW or unmodulated operation. The output (E_2) from the 230A is increased until the demodulated output equals that noted previously.

The cross modulation performance may then be calculated as follows: $C_m = 20 \log E_2/E_1$.

It should be observed that the demodulated output falls off when E_1 only is removed.

Image and IF Rejection Tests

Using the 230A Power Amplifier, IF rejection tests are made on receivers where this rejection is extremely high (in the order of 100 db or more). The image frequency (F_I) is that frequency which is twice the intermediate frequency (IF) away from the desired signal frequency (F_0), in the same direction as the local oscillator (F_{lo}). See Figure 8.

IF rejection is made at the intermediate frequency by driving into the receiver front end. This attenuation level is usually much higher than image rejection; so that even less sophisticated receivers may require use of the 230A Power Amplifier.

V_1/V_2	V_T	db
1 v	1.7 v	-48
2 v	3.5 v	-46
3 v	5.6 v	-33
5 v	8.7 v	-24
7 v	11.6 v	-22

Figure 6. Typical Intermodulation Present In Type 230A

In AM systems, an increase in distortion indicates that overloading has taken place.

RF Wattmeter Calibration

RF wattmeter calibration is accomplished by using a standard signal generator in conjunction with the 230A Power Amplifier as a power source. Power is then connected to the standard wattmeter, and then to the wattmeter to be calibrated. Power up to 9 watts is available for short periods for this application. Higher than specified input levels are necessary, however.

RF Voltmeter Calibration

The procedure for RF voltmeter calibration is somewhat different than for wattmeter calibration, since the voltmeter is usually a relatively high impedance device. The National Bureau of Standards has developed an A-T (Attenuator-Thermocouple) type standard RF voltmeter which may be used for this application. This instrument is a standard for RF voltages from 1 to 300 volts at 10 to 1000 Mc. The output voltage of the 230A amplifier is a function of loading and can be increased many fold over the 50-ohm value. Experiments to date, using line stretchers, stub tuners, and resonant transformers, indicate that voltages from 60 to 100 volts may be developed for voltmeter calibration and other applications requiring large signal levels.

COMPONENT TESTING

Component testing usually takes the form of a breakdown or parameter change which can be checked after sub-

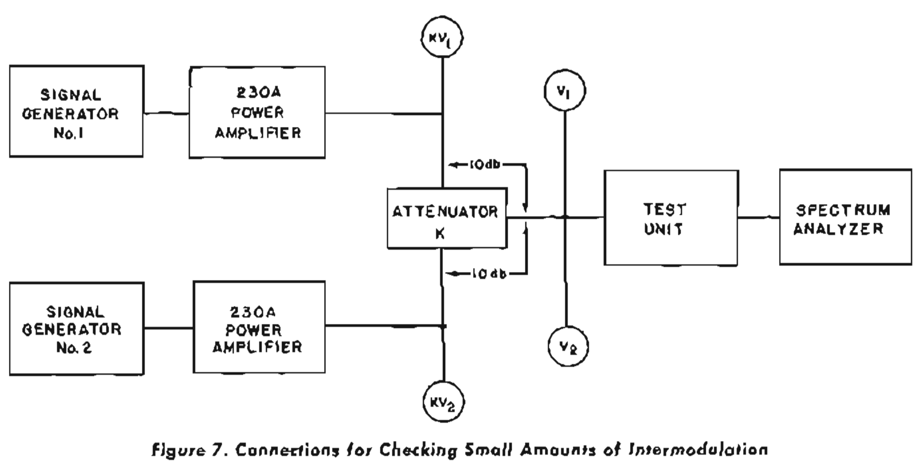

Figure 7. Connections for Checking Small Amounts of Intermodulation

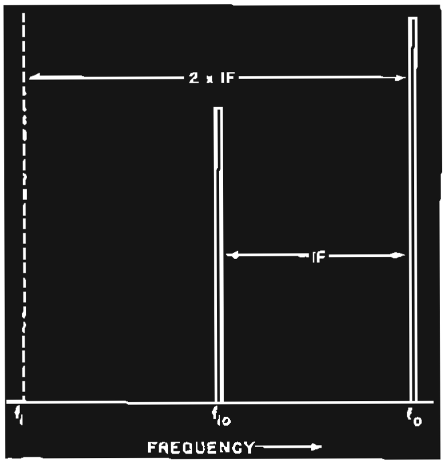

Figure 8. Image Frequency

jecting the component to higher voltage or power stresses than are normally encountered in standard tests. Chokes, resistors, and capacitors are examples of such passive components. For example, in an actual test, a small 7.5μh choke subjected to 100 volts RF at 20 Mc, exhibited no change in Q or L after testing. It would be safe to conclude, therefore, that these chokes could be used up to this level. This application could also be extended to active components.

Diode rectification efficiency can be determined with the setup shown in Figure 9. The gain of power transistor circuits can be determined in a similar fashion, since the voltages are measureable with existing RF voltmeters. Circuits requiring high voltages; such as discriminators, limiters, power amplifiers, etc., may be supplied by the 230A. This is sometimes called "down-stage" testing.

HIGH LEVEL DRIVER

As a high level driver, the 230A can be used to power bridges and slotted lines to improve the resolution and accuracy of these measurements. Computers that require high level signals sources for synchronizing purposes at moderately high frequencies may also be driven by the 230A Power Amplifier.

ANTENNA TESTING

The 230A is capable of supplying moderate power for antenna measurements, while, at the same time, providing relatively small leakage from the Power Amplifier itself. This feature permits two antennas to be closer together, thereby shortening the range required.

ATTENUATION MEASUREMENTS

Using the 230A Power Amplifier and an RF millivolt meter, attenuation measurements can be made in the order of 80 db. The 230A provides an additional 28 to 40 db of gain or signal level (assuming the circuit being measured will permit the high voltage) to add to the existing measuring system in the field of attenuation measurements. Filters, long transmission lines, etc., can be tested in this manner.

LOW LEVEL AMPLIFIER

As a low level amplifier, the 230A can be used to amplify small signals, such as a crystal spectrum at a given frequency, for frequency drift measurements.

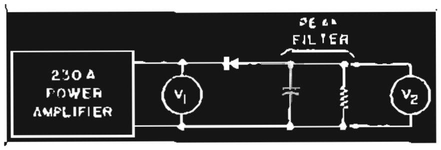

Figure 9. Connections for Checking Diode Rectification Efficiency

FREQUENCY MULTIPLYING

A number of approaches to this application are possible. First, it is possible to amplify the harmonics present in the input signal. The output under these conditions is in the order of 0.2 to 0.5 volts, with 0.2 volts of fundamental input. Another approach is to use a semiconductor harmonic generator to augment the harmonics present in the input signal. This technique yields several volts output, depending upon the input levels available. If sufficient input is available, the 230A input stage may be overdriven and the attendant distortion will produce higher harmonic levels. Approximately 1 to 2 volts may be expected for inputs of the order of 1 volt. Crystal frequency synthesizer output may be multiplied as many as ten times, extending the usefulness of these units to the UHF range.

RADIO INTERFERENCE TESTING

Other applications of the 230A Power Amplifier are found in the Radio Frequency Interference (RFI) field of measurements.

Screen Room Testing

Screen rooms are used to reduce RFI in cases where equipment being tested, or operated, is capable of causing RFI, or is sensitive to RFI. The screen room, in either case, must provide a prescribed amount of attenuation; usually in the order of 100 db or more.

A method for testing screen rooms is described in Military Specification, MIL-E-4957-A (ASG). This method has become general practice. The specification includes a test at 400 Mc; a frequency that has proved to be rather critical, regardless of room size.

In general, the procedure for making this 400 Mc test is as follows (Figure 10). First, a clear channel at approximately 400 Mc should be selected by listening with the field intensity measuring or receiving equipment antenna, outside the shield room. The antenna is placed a few inches from the outside of the screen room to be tested, several feet from the transmitting antenna. If the receiving equipment has a calibrated attenuator system, the signal generator and Power Amplifier may be operated at full output, and the attenuator set to give a convenient meter reading on the receiving equipment. Alternatively, the receiving equipment can be set to high sensitivity and the signal generator level adjusted to produce a convenient meter reading. The receiving antenna is then moved inside the screen room and placed within a few inches of the wall being tested. With full output from the 230A, the receiving antenna is used as a probe, along the seams, etc., and the point of maximum leakage is determined. The appropriate attenuator setting is read as the shielding attenuation figure. The procedure is repeated for the other walls of the screen room.

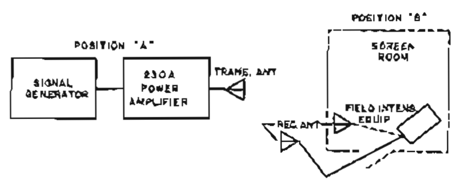

Figure 10. Setup for Screen Room Testing

A paper[5] was given at the 1961 IRE Convention in New York which described another approach for measuring shielding performance at critical frequencies. One of these frequencies is the frequency (f_o) at which the screen room is resonant; usually between 50 and 200 MC. The frequency (f_o) can be readily detected using a grid-dip meter technique. The procedure, described in the IRE paper, for making the attenuation measurement is basically the same as previously described except that the measurement is made at the center of the screen room.

Other RFI Applications

Other RFI applications include powering of probes, loops, etc., for the testing of filters, shielded cables[8], and small compartments. It is also possible to conduct "Standard Susceptibility to Radiation Tests."[4] The output of the 230A Power Amplifier is sufficient to set up standard field intensities of greater than 1 volt per meter throughout most of the frequency range.

Additional Performance Data

The following performance data has been taken on the 230A, and, while not incorporated into the specifications, is considered typical of production units.

Noise Figure — Approximately 8 db, or about 4 microvolts per Kc of bandwidth.

Power Output — It has been found that if enough drive is available, the 230A may produce as much as 16 watts for short periods of time without damage at some frequencies.

Conclusion

We have presented here many of the applications of the 230A Power Amplifier. Even at this writing, additional applications are in the making. These will be taken up in subsequent issues of the Notebook, as the details become known.

The author wishes to thank Mr. Fritz Popper of Shielding, Inc., Messrs. Sidney White and Guy Johnson of USASRDL, and the BRC Engineering Staff for their comments and assistance given during the preparation of this article.

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4. MIL-E-4957-A(ASG), Enclosure, Electromagnetic-shielding, Demountable, Prefabricated for Electronics Test Purposes.
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Using the FM Stereo Modulator

Willard J. Cerney, *Sales Engineer*

Figure 1. Type 219A FM Stereo Modulator

INTRODUCTION

With the approval of an FM Stereo broadcast system by the FCC on April 20, 1961, as specified in FCC Docket 13506, BRC designed and made available the Type 219A FM Stereo Modulator, a stable, easy to use, compact, economical source of the multiplex signal for use with FM Signal Generators, or for direct use with receiving multiplex adapters. Complete details of the basic design and circuitry of the Modulator may be found in Notebook No. 30. The purpose of this article is to describe the various tests which may be performed with this new instrument.

AREAS OF APPLICATION

The Type 219A FM Stereo Modulator finds application in the design, production testing, and servicing of FM multiplex receivers and adapters. The extreme versatility of the instrument in providing both standard and non-standard signal outputs make it ideal for laboratory use. The reliable, compact design, and the relatively high available output, permitting the operation of several test stations from a single modulator, make it well suited for production line applications. The functional controls and convenient output meter fulfill the requirements of service applications.

METHODS OF OPERATION

The Type 219A FM Stereo Modulator generates the complete multiplex signal, consisting of (L + R), (L − R), and 19 KC pilot and may be externally fed with an SCA sub-carrier. Employing either the internal 1 KC oscillator or an external tone or program source, the output may be used directly for the testing of multiplex adapters or other base-band circuitry in the 50 cps — 70 KC range.

Alternatively, the output of the modulator may be used to modulate a suitable FM Signal Generator, such as the BRC 202E or 202H, to provide a complete multiplex signal at RF, simulating transmissions in the 88-108 MC broadcast band. The Type 519A Adapter provides a convenient means of interconnecting these instruments and permits use of the modulating oscillator in the 202E/H as an audio tone source to accomplish fidelity and distortion measurements from 50 cps to 15 KC without an external audio oscillator. The "set" position on the 219A output mode switch permits convenient setting of FM deviation on the signal generator to equate 100% multiplex output to 75 KC deviation.

DEFINITION OF TESTS

Essentially, there are four basic tests which may be performed with the Type 219A FM Stereo Modulator alone or in combination with a suitable FM Signal Generator.

1. Stereophonic or Channel Separation

Stereophonic or channel separation, usually expressed in db, is the ratio of the signal output from an excited channel (L or R) to the residual or undesired output present on the non-excited channel (R or L).

2. (L + R) — (L − R) Crosstalk

(L + R) — (L − R) Crosstalk is the residual or undesired (L + R) or (L − R) output when either L = R or L = −R, respectively, in the multiplex signal.

3. Electrical Fidelity and Distortion

Electrical fidelity and distortion are measured conventionally on both (L) and (R) channels over the audio frequency range from 50 cps to 15 KC.

4. Non-Standard Signal Makeup

Non-standard signal makeup involves the setting of (L + R), (L − R), and/or 19 KC pilot levels over a range simulating the effects of propagation; eg., multipath transmission.

TEST SETUPS AND INTERCONNECTIONS

Typical equipment setup and interconnections for the tests, listed above, are shown in block diagram form in Figures 2 through 4. Figure 2 shows the connections for measurement of a multiplex adapter or base-band circuitry of a receiver. Figure 3 shows the connections for measurement of an FM receiver and multiplex adapter (or multiplex receiver). Figure 4 is essentially identical to Figure 3 except that a distortion analyzer, output meter, or other suitable instrument is connected in place of the oscilloscope for testing recovered audio.

AUXILIARY TEST EQUIPMENT REQUIREMENTS

The approved stereo broadcasting system employs base-band frequencies

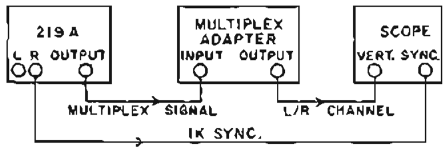


Figure 2. Connections for Checking Multiplex Adapter

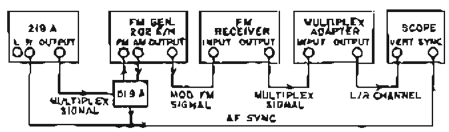


Figure 3. Connections for Complete System Test

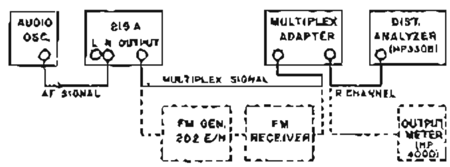


Figure 4. Connections for Fidelity and Distortion

over the range from 50 cps to 75 KC and it is essential that all auxiliary equipment have adequate bandpass capabilities to handle the complex waveforms with negligible time delay and amplitude variation over this range. A typical FM Signal Generator must have adequate FM modulation bandwidth in terms of sufficiently constant amplitude response and time delay in order to maintain stereo separation and should, of course, introduce minimum amplitude distortion.

While the levels of the various components of the multiplex signal can be readily and precisely adjusted using the peak reading output meter on the 219A, an external oscilloscope is desirable to measure and interpret the performance of the receiver or adapter under test. The oscilloscope used must possess minimum variation in amplitude response and linearity of phase vs. frequency within ± ½° from 50 cps to 70 KC. Figure 5A shows a typical oscilloscope pattern for a properly balanced multiplex signal (without 19 KC pilot). The flat base line indicates amplitude identity between (L + R) and (L — R); a small and tolerable amount of phase shift is indicated by the non-coincidence of the zero crossings on the base line. Figure 5B indicates amplitude unbalance (excess L — R) with approximately the same phase shift present in Figure 5A. It is important to note that this amplitude and phase shift, viewed on the oscilloscope, may be due to either improper adjustment of the 219A (readily checked on the output meter) or inadequate response characteristics of the oscilloscope. The peak voltmeter in the 219A may be relied on to indicate equality of L + R and L — R signal peak amplitudes to better than ±1%.

OSCILLOSCOPE SYNCHRONIZATION

The oscilloscope may be synchronized with either the audio tone input signal or the 19 KC pilot carrier, generated in the 219A. The audio sync signal should be obtained from the external tone source or the right (R) input terminals of the 219A if the internal 1 KC tone oscillator is used. The 19 KC sync signal is available directly from a jack on the rear of the 219A. When audio tone synchronization is employed, the oscilloscope pattern for the complete multiplex signal, including 19 KC pilot carrier, is shown in Figure 5 C.

CHANNEL IDENTIFICATION

It is often necessary to identify left (L) and right R) channels. While such identification can be made by observing the multiplex signal presentation with audio sync on an oscilloscope and ap-

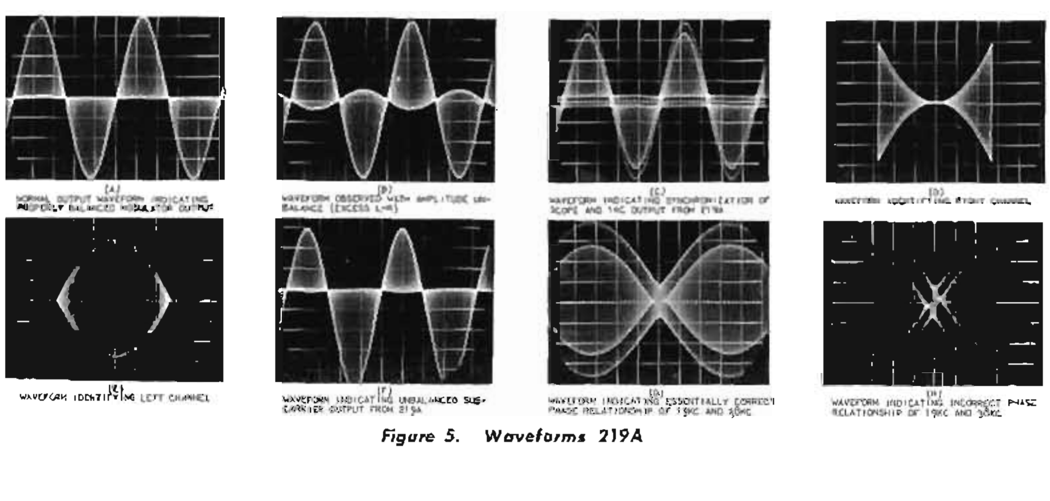


Figure 5. Waveforms 219A

plying the basic definitions in FCC Docket 13506, a far simpler and easier method is available. In this method, the multiplex signal is observed using the 19 KC sync output of the 219A, phase-shifted 45°, on the external input of the oscilloscope as shown in Figure 6. The phase-shift network, including typical component values, is shown in Figure 7. The oscilloscope patterns that will result from audio tone signals on either the right (R) or left (L) channel are shown in Figures 5Dand 5E.

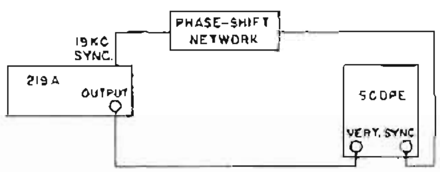


Figure 6. Connections for Channel Identification

MODULATOR BALANCE

The 38 KC carrier, generated by the balanced modulator, must, of course, be properly suppressed and the 219A is specified to produce less than 0.5% of 38 KC, when properly adjusted, compared to the level of the composite output. Proper balance will produce the waveform shown in Figure 5A; the waveform resulting from a grossly unbalanced sub-carrier output is shown in Figure 5F.

PHASING OF 19 KC AND 38 KC CARRIERS

Proper phasing of the 19 KC pilot carrier and the 38 KC carrier, modulated by (L — R), is essential for proper demodulation in the receiver. This phasing in the 219A may be readily verified by observing oscilloscope patterns of (L — R) and 19 KC pilot, employing the setup shown in Figure 6, except that the sync input of the scope is fed from the audio tone signal. A typical waveform with nearly correct phasing is shown in Figure 5G; the "diamond" pattern, in the center of the display, should be perfectly symetrical. Incorrect phasing is shown in Figure 5H

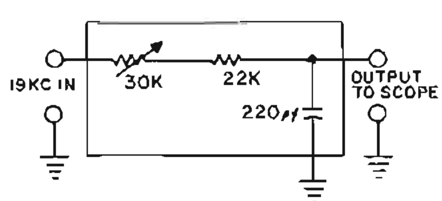


Figure 7. Phase Shift Network

TEST PROCEDURES

Stereophonic or Channel Separation

1. Adjust the 219A for standard multiplex output with an audio tone signal of the desired frequency applied to either the left (L) or right (R) channel.

2. Adjust frequency, modulation, and output of signal generator; also frequency and output of receiver. (This step is omitted if measurements are to be made directly at base-band, eliminating the signal generator).

3. Adjust receiver stereophonic controls for maximum audio output from channel on which audio tone signal has been applied and minimum output from undesired channel. If receiver or adapter has not been previously aligned, also adjust matrix to optimize these conditions.

4. Measure the audio output levels of the desired and undesired channels. Alternatively, the 219A audio tone modulating signal may be reversed to the opposite channel, thereby permitting reading of audio output to be made on either channel individually.

NOTES: (a) The receiver or adapter matrix and stereophonic balance controls are usually adjusted for optimum separation at 1 KC; measurements at other audio tone frequencies are then made without disturbing these settings.

(b) Since the audio output of the receiver or adapter under test may include 19 KC and/or 38 KC components, either low-pass filters (15 KC cut-off) should be inserted between the audio outputs and the indicating voltmeters or readout should be made on an oscilloscope, ignoring the 19 and 38 KC components in the measurement.

(L + R) — (L — R) Crosstalk

1. Repeat steps 1 thru 3 under separation.

2. Select L + R or L — R by means of the function switch and measure the undesired L + R or L — R output in the device under test. If a 19 KC pilot is required, cross-connect (parallel) the 219A left (L) and right (R) inputs to obtain (L — R) null or, conversely, provide left (L) and right (R) inputs so that L = —R, providing (L + R) null. (If the 219A internal 1 KC oscillator is employed, switching may be conveniently performed by operating the matrix switch).

3. Measure the undesired (L + R) or (L — R) output, respectively, in the receiver or adapter under test.

NOTE: If an FM Signal Generator is employed for measurements through the RF section of a receiver, crosstalk, prior to demodulation due to overload in the receiver, may be detected by varying the RF output level of the signal generator.

Electrical Fidelity and Distortion

1. Repeat steps 1 thru 3 under separation.

2. Measure the audio output level from the receiver or adapter at selected frequencies in the 50 cps to 15 KC range.

3. Measure the distortion on the receiver or adapter audio output at selected frequencies in the 50 cps to 15 KC range.

NOTE: The audio tone input to the 219A must be readjusted for standard level at each test frequency.

Non-Standard Signal Makeup

Repeat steps 1 thru 3 under separation, modifying the levels of 19 KC pilot carrier, (L + R), and (L — R) to simulate the desired test condition.

NOTE: Receivers and/or adapters, especially those employing phase-locked sub-carrier oscillators, should be tested at various levels of 19 KC pilot carrier.

CONCLUSION

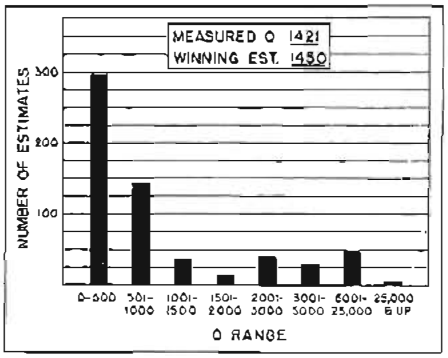
The Type 219A FM Stereo Modulator is an excellent source of FM stereo multiplex baseband signals for use with FM signal generators or for the direct use with receiving multiplex adapters. When used with the BRC Types 202E or 202H signal generators, the instrument provides modulated RF stereo multiplex signals which are of a quality better than that specified by FCC. The versatility and reliability of the instrument, together with its compactness of design, relatively high available output, functional controls, and convenient output meter, make it a valuable tool for use in the design, production testing, and servicing of FM multiplex receivers and adapters.

EDITOR'S NOTE

Q Meter Winner

The circuit displayed at the IRE show has been carefully measured on the Type 280A UHF Q Meter and the Q is **1421**. The winning estimate of **1450** was submitted by Mr. Jan Solomon, Management Engineer with Federal Electric Corp., Paramus, N. J. Second, with an estimate of 1480, is Mr. B. Nohre, an Engineer from Stockholm, Sweden.

Q estimates in the contest ranged from 1 to more than 25,000, with 40 persons guessing within 2% and 21 persons guessing within 1% of the measured Q. The distribution of estimates over the entire range of estimates is shown in the bar graph.



Distribution of Q Estimates

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The NOTEBOOK

BOONTON RADIO COMPANY · ROCKAWAY, NEW JERSEY

A Division of Hewlett-Packard Company

New Techniques In FM Linearity Measurement

JOHN P. VAN DUYNE, *Engineering Manager*

INTRODUCTION

It has been characteristic of the development of electronic systems and terminal equipment that effort has been continually devoted to methods of making the output of the equipment or system more faithfully reproduce the important characteristics of the input. A major consideration of this nature has been improvement in the linearity of the transfer characteristics of devices, equipment, or systems. While there are equipment and devices in which some function, other than a straight line, is wanted between the input and output, the straight line or linear relationship is by far the most common. There are several detailed motivations for this activity which have slightly different connotations in various fields of endeavor. However, it is possible to generalize on some of the factors that push development in the direction of improved linearity of transfer characteristic. In the fields of communication equipment development, extreme linearity did not become of major importance until these systems were directed toward more efficient use of the frequency spectrum. This frequently results in multiple channel transmission in which nonlinearity of the transfer characteristic causes unwanted crosstalk between the various channels.

In recent times, the use of communications type systems for the transmission of scientific, commercial, and other forms of data has become common. Since the information being transmitted has other than a subjective end result, the accuracy of data transmission is related to the linearity of the transfer

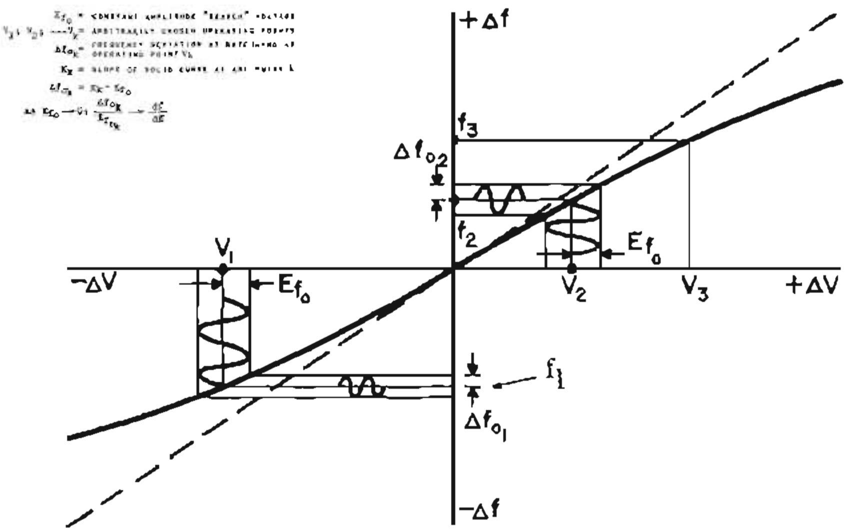

Figure 1. Transfer Characteristic with Typical Nonlinearity

characteristic. Even in the field of entertainment electronics, pressure for more faithful reproduction has been continuous. A recent development in this field, that of a system of FM Stereo Multiplex transmission,[1, 2] has forced the manufacturers of FM transmitters and receivers for the entertainment field to improve the linearity of their equipment to minimize crosstalk between the Monaural and Stereo channels of the system. One fact that should be noted is that, because of the different objectives and the different technologies which have developed around these objectives, the effects of nonlinearity in the transfer characteristic are described in many different terms.

GENERAL COMMENTS ON NONLINEAR DISTORTION

Before discussing the detailed techniques of measurement, it might be well to consider what we really mean by "nonlinearity", and the distortion which results therefrom. Nonlinearity, as used in this discussion, refers to that characteristic of a circuit which causes the output to be related to the input by other than a straight line function. As a result, the circuit may not be specified by linear differential equations with time as the independent variable.[3]

Many terms are used to describe distortion due to this concept of nonlinearity. In order that we may clearly understand the problems, a few definitions will be resorted to. *Nonlinear Distortion* has been defined by the IRE Standards on Circuits as "Distortion caused by a deviation from a desired linear relationship between specified measures of the output and input of a system."

NOTE: The related measures need not be output and input values of the same quantity; e.g., in a linear detector, the desired relation is between the output signal voltage and the input modulation envelope."

Other names used to describe the effects of nonlinear distortion are *Amplitude Distortion, Waveform-Amplitude Distortion*, and *Harmonic Distortion*. In addition to these names, there are specific means of measuring nonlinearity among which are *Total Percent Harmonic Distortion* and *Inter-*

YOU WILL FIND . . .

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modulation Distortion. It should be stressed, that while there may be different names for the effects of nonlinearity and many different standard ways of expressing the effects numerically, all of these related types of distortion stem from the same nonlinearity of the transfer characteristic as discussed above.

The casual reader of papers on distortion in the field of sound reproduction, may be left with the feeling that, somehow, there is a large difference between intermodulation distortion (IM) and total percent harmonic distortion (HD). While there may be advantages in one system of measurement or the other, when it is desired to relate the numerical result of the measurement to the subjective response of the listener, it should be remembered that finite results in either an IM or HD measurement occur due to nonlinearity of the transfer characteristic, and measurement by either method may be analytically related for any specific shape of curve. Further discussion of this may be found in Reference 4.

A related case occurs in the measurement of deviation from a frequency modulated source by zero beating another RF source with the peak excursions of the FM carrier. Then, the frequency of this source, for zero beat with the upper and lower FM excursions, is measured and compared with the unmodulated FM carrier. Equality of these differences has sometimes been taken as evidence of good linearity. That this is not so, can be seen in Figure 1. Since voltages V_1 and V_3 have been chosen equal and opposite, and the resulting frequencies, f_1 and f_3, are equidistant from the origin, the condition of the beat frequency measurement is described. However, the transfer characteristic shown is anything but linear. It is true, though, that the symmetry about the unmdoulated carrier, so measured, indicates a predominance of odd-order harmonic distortion. Asymmetry about the unmodulated carrier is a strong indication of even-order distortion.

Many observers have commented on this so called "FM carrier shift" or difference in frequency between the average carrier frequency (when modulated) and the unmodulated carrier. This is due to the "DC" or constant term which appears in a Fourier expansion of the polynomial representation of a transfer function with even-order curvature (second harmonic or "squared" terms), and is a reliable indication of the presence of even-order harmonic distortion; usually second harmonic.

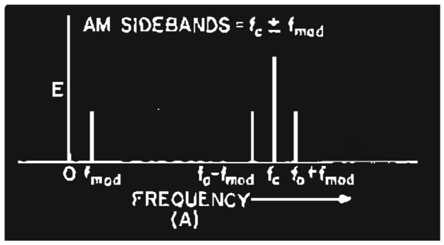

Figure 2A. Undistorted AM Spectrum

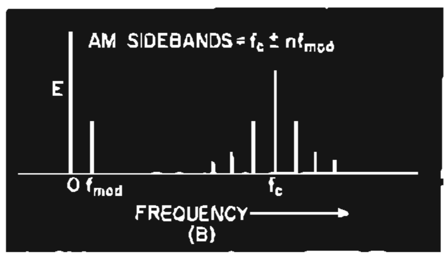

Figure 2B. Distorted AM Spectrum

FM SIGNAL GENERATOR MEASUREMENT

In the development of the Type 202J and 202H FM Signal Generators, Boonton Radio Company faced a specific aspect of the nonlinearity problem. Since we were designing a linear frequency modulator, it was necessary that we be able to measure the transfer characteristic nonlinearity of this modulator to a high degree of precision. An obvious way to accomplish this would be to measure the results of frequency modulation through a perfectly linear demodulator. However, it is somewhat difficult to produce the perfect demodulator without having a perfect modulator with which to measure it. Therefore, another method was needed. Let us examine some common techniques.

In the simple case of amplitude modulation, it is possible, by means of a suitable wave analyzer, to measure the frequency spectrum resulting from the modulation process. If the modulating signal is free of distortion and the modulation process is perfectly linear, the resulting spectrum in the vicinity of the carrier will contain only the carrier and the two sidebands, separated from the carrier by the modulation frequency as seen in Figure 2A. If nonlinearity of the transfer characteristic exists, the results may be as shown in Figure 2B. Here, the appearance of sidebands, separated from the carrier by integral multiples of the modulating frequency, is not only qualitative evidence of nonlinear distortion, but affords a measure of this distortion if the amplitudes of all the sidebands relative to the carrier are determined.

In the FM case, for deviations and modulation frequencies for which the modulations indices, defined as

$$m = \frac{\text{peak deviation}}{\text{modulating frequency}},$$

are appreciably greater than unity (the usual case), the resulting frequency spectrum becomes exceedingly complex, even for an undistorted modulation process.[5] In the presence of nonlinearity of the frequency modulator, the spectrum becomes even more complex; thus ruling out spectrum analysis as a simple means of measuring nonlinearity.

A second method for measuring a frequency modulator transfer characteristic is possible, if the modulator is capable of operation at dc. This method consists of introducing a known increment of modulating voltage and measuring the resulting frequency increment. Its precision is limited by the stability of the carrier frequency in the absence of modulation and the ability to read small increments in voltage and frequency precisely. Modern measuring equipment has solved the latter two problems, but not the former.

In addition to the matter of precision, it was desired by BRC to be able to make adjustments on the frequency modulator for minimum nonlinearity as a result of an instantaneous display of nonlinearity of the frequency modulator. This same need is felt by designers and manufacturers of FM transmitters and receivers.

DEVELOPMENT OF THE METHOD

In order to best understand the method to be described, let us avoid specific numbers or circuits, and resort to a general discussion of a circuit which relates a change in voltage to a change in frequency or vice versa. This covers the general case of frequency modulators and demodulators. Reference to Figure 1 will show the transfer characteristic of such a circuit. The horizontal axis represents changes in voltage and the vertical axis changes in frequency. Either may be the independent variable for purposes of our discussion. Shown dotted is a reference straight line which passes through the origin, as does the arbitrary nonlinear curve. Our problem is to measure the departure of the actual curve from the ideal straight line. A corollary to this problem is to define the initial relationship of the straight line to the actual curve. This may be done in several ways. It has been common for people in the telemetry and data transmission field to be interested in the instantaneous departure of their actual transfer characteristic from the ideal straight line. This stems from their historical use of such characteristics to transmit simple analog information. Several ways have been used to express this, but a common one is illustrated in the performance specification of the BRC 202J Telemetry Signal Generator. Here, the reference straight line chosen is that line for which the rms sum of the differences between the actual curve ordinates and the straight line ordinates is a minimum. Another method would be to have the average departure a minimum, or the peak departure a minimum. For small, low order, nonlinearities, there is little difference between these cases.

Figure 1 shows that one way to describe the departure of the curve from the straight line is to measure a large number of individual ordinates of the actual curve, subtract them from the ordinates of the straight line, and report the differences in tabular or curve form. This type of data would result from a point-by-point measurement of the characteristic, using an electronic counter to measure the frequency and a precision incremental voltmeter for the voltage axis. This technique is not instantaneous, even when automated, and yields more data than can be conveniently digested quickly.

It is also evident in Figure 1, that if the slope of the actual curve could be compared, at any point, with that of the straight line, a relationship covering the nonlinearity could be readily developed. Pursuing this line of reasoning, it should be possible to make an incremental measure of the slope of the curve by introducing a small amplitude

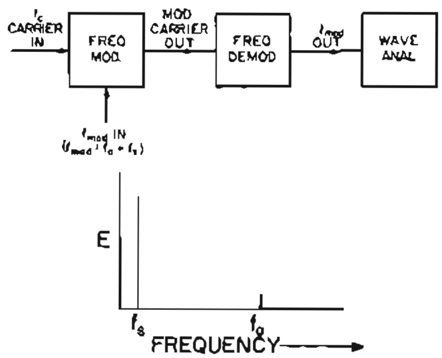

Figure 3A. Linear AM and FM Demodulator with Output Spectrum

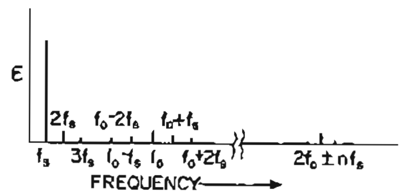

Figure 3B. Nonlinear FM Demodulator — Linear FM Modulator Spectrum

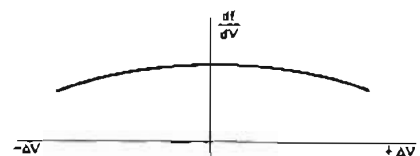

Figure 3C. Plot of df/dV of Figure 1

sine wave of voltage, E_{f_0}, (if voltage were the desired independent variable) superimposed on a direct voltage. The direct voltage may then be varied. If the resulting frequency variation, due to the small constant amplitude search voltage E_{f_0} were measured at each value of V, and plotted as a function of V over the entire range of interest of the voltage axis, a plot of the varying slope of the actual transfer characteristic would result. The extent to which this departs from the slope of the straight line, is a measure of the nonlinearity of the transfer characteristic. This method sounds relatively laborious, until it is realized that this can be done dynamically by allowing V to vary over the desired dynamic range at a sinusoidal rate, f_s. The slope of the transfer characteristic can be plotted on an oscilloscope by driving the vertical axis with an amplitude proportional to Δf and the horizontal axis with a voltage proportional to V, and varying at rate f_s.

Let us now consider the case of a linear frequency modulator and frequency demodulator, to which we will apply the two test voltages, E_{f_0} and V_{f_s} with the understanding that V_{f_s} is of such magnitude as to sweep through the entire dynamic range of interest and that E_{f_0} is very much smaller than V_{f_s}, typically one-tenth as great. This situation is shown in Figure 3.

As might be expected, the output spectrum of the linear demodulator driven by the linear modulator is simply E_{f_0} and V_{f_s}, with the amplitude relationship established by the input. If we now further assume that the frequency modulator is linear, but that the frequency demodulator is nonlinear (curve of the nature shown in Figure 1), we realize that if E_{f_0} were to be plotted as a function of V_{f_s}, we would approximately plot the slope of the transfer characteristic being studied, namely, that of the assumed nonlinear demodulator. Therefore, we may write that $\Delta f = \frac{df}{dV}(E_{f_0})$, where df/dV is recognized as the slope of the transfer characteristic in Figure 1.

Another view of the same phenomena is shown in Figure 3B. This shows the output spectrum of the nonlinear demodulator driven by the linear modulator, and with the input as assumed previously. The output spectrum is now far more complicated than in the linear case (Figure 3A), consisting not only of V_{f_s} and E_{f_0}, but also of integral multiples with decreasing amplitude of both f_s and f_0, and sum and difference frequency terms resulting from the nonlinearity of the transfer characteristic.

Figure 3C shows the df/dV curve as determined by plotting the amplitude of Δf from the demodulated output as a function of V_{f_s}. This would be an accurate plot of the transfer characteristic slope if the magnitude of E_{f_0} in the input were made vanishingly small compared to V_{f_s}. For the cases studied by the writer, this approximation is entirely satisfactory if $V_{f_s}/E_{f_0} \approx 10$. It should be emphasized that Figure 3C and the spectrum plot of Figure 3B contain the same information concerning the transfer characteristic nonlinearity, but displayed in a different manner. The question now is, how may we simply achieve such a display?

The block diagram of Figure 4 shows the introduction of a bandpass filter which accepts f_0 and the significant sidebands about it due to f_s, and the display of the resulting carrier envelope on the vertical axis of an oscilloscope. The horizontal axis is driven by the initiating V_{f_s} from the input of the frequency modulator. Figure 5 shows a

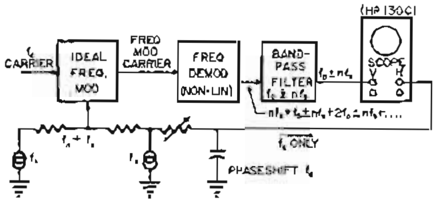

Figure 4. Block Diagram of Circuit Used to Provide Instantaneous Display of Transfer Characteristic Change of Slope

typical result. Since we have passed our demodulated signal through a bandpass filter, the upper and lower section of the envelope presented to the oscilloscope must be symmetrical; therefore, we may select either for consideration. (The final equipment shown in Figures 10 and 11 chose the negative half of the envelope for display.) Where the amplitude of the f_s sidebands about f_o is very low, as would be true of a highly linear circuit, the resulting departure of the oscilloscope trace from a rectangle is small. Since we may ignore half the envelope, it is possible to offset the centering of the oscilloscope and greatly expand either half of the display. If an oscilloscope, such as the Hewlett-Packard 130C, is used, linearity of the display will be maintained for an expansion of half the envelope by 10 times. This permits measuring changes in slope down to about 0.1% of one-half the envelope.

The departure of the envelope in this display from a reference value (chosen as the slope at the origin of Figure 1, and appearing on the vertical center line of the oscilloscope display

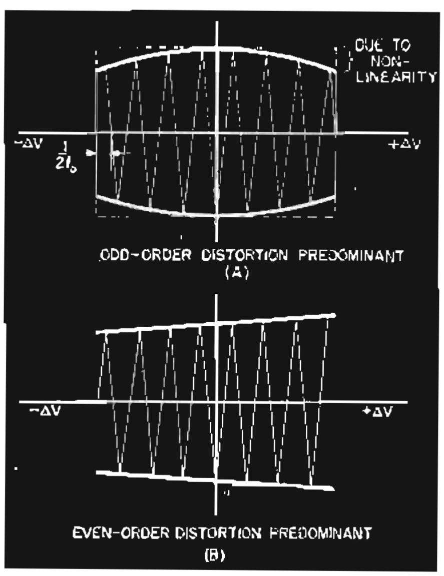

Figure 5. Typical Display from Circuit in Figure 4.

in Figure 5), is a direct measure of the change in slope of the transfer characteristic. Therefore, the slope of this curve is approximately the second derivative of the transfer characteristic. If the envelope of Figure 5 shows symmetry about the vertical center line of the trace (Figure 5A), it can be shown that this is a result of odd-order (3rd, 5th, etc.) harmonic distortion. If the envelope is asymmetrical about the vertical center line and has mirror image symmetry about the horizontal center line (Figure 5B), the distortion is largely even-order (2nd, 4th, etc.) Thus, it can be seen, that much information is contained in the displays of Figure 5, produced by the circuit of Figure 4.

Up to this point in the discussion, we have been assuming a linear modulator. The reader may well comment, "This is all fine, but where do I get the perfectly linear modulator?" The technique illustrated in the block diagram of Figure 6 shows a method of "synthesizing" an FM signal, modulated simultaneously by two frequencies (f_o and f_s), in which sidebands of f_s around f_o are not produced, even though the individual signal generators have some nonlinear distortion. The technique used is borrowed directly from the receiver practice of mixing an incoming signal with a local oscillator and amplifying at a fixed intermediate frequency before frequency demodulation is performed. Let us then consider the linearly demodulated spectrum produced by modulating either signal generator by a single frequency, remembering that one generator is modulated by f_s and the other by f_o.

The output spectrum from a perfect demodulator driven by a nonlinear modulator with single frequency input would consist of the original modulating frequency and integral multiples of this frequency, generally in decreasing magnitude. Since the two frequencies f_s and f_o do not simultaneously exist in either frequency modulator, it is not possible to produce sum and difference frequencies in the linearly demodulated output. (See Figure 7.) When the carrier frequencies (f_{c_1} and f_{c_2}) generated by the two FM sources (one modulated by f_s and the other by f_o) are mixed at RF (Figure 6), and the difference carrier frequency is amplified by a suitable selective amplifier, the instantaneous frequency in the selective amplifier is the simple arithmetic difference

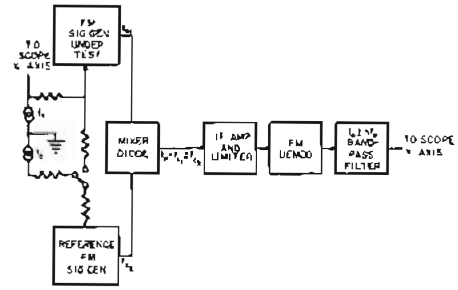

Figure 6. Block Diagram of Circuit for Synthesizing Perfectly Linear Modulation

of the instantaneous frequency modulated sources. Thus, no sum and difference frequencies of the modulating signals f_o and f_s appear in the FM signal at average frequency, f_d.

Now, following the technique discussed previously, let us introduce a bandpass filter centered on f_o in the output of the demodulator. The bandwidth of this filter must be wide enough to accept the significant sidebands of f_s about f_o. For the cases studied by the writer, acceptance of sidebands up to the 3rd is sufficient. This bandpass is shown by dotted lines in Figure 7. It will be seen that this filter eliminates f_s and all of the distortion products due to the nonlinearity of the frequency

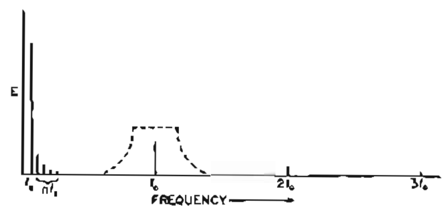

Figure 7. Demodulator Output Spectrum of nf_s and nf_o with Superimposed Bandpass Filter — Linear Case

modulators in the individual signal generators which produce the difference frequency, f_d.

Now, let us consider the effect of nonlinearity in the frequency *demodulator* when driven by a signal generated by the circuit of Figure 6. It is evident from the previous discussion, that the spectrum will consist not only of the components shown in Figure 7, but of sidebands about f_o and its harmonics spaced by nf_s. (See Figure 8.) If the output of the bandpass filter is connected to an oscilloscope, as discussed previously, an envelope, similar to that of Figure 5, will be seen. Thus, we have a technique for using two signal generators of finite modulator nonlinearity to synthesize an apparently distortionless signal which may then be

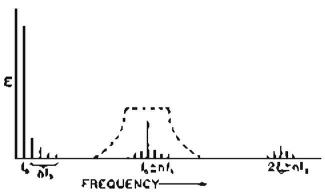

Figure 8. Demodulator Output Spectrum of nf_s and nf_o Including nf_s Sidebands About f_o — Nonlinear Case

used to measure the nonlinearity in a frequency demodulator.

To avoid the necessity of offsetting the axis of the oscilloscope display and expanding the scale to look at small deviations of the envelope from rectalinearity, it is possible to eliminate the carrier with a linear amplitude demodulator. The composite block diagram of Figure 9 shows the application of a simple linear AM detector followed by a low-pass filter to eliminate the carrier, f_o, from the display. The output from the linear detector will be a direct current component proportional to the peak amplitude of f_o, together with the envelope produced by the nf_s sidebands beating with the carrier. The percentage change in slope, compared to a straight line, may be read by expressing the peak value of the demodulated envelope as a percentage of the dc component (which is proportional to the peak value of f_o).

If the output of the AM detector and nf_s low-pass filter is read on an averaging meter, an average value of the percent change in slope will be read. For the cases studied by the writer, the amplitude of the significant harmonics (2nd and 3rd) of f_s has been relatively low, so that the value read by an rms calibrated, average reading meter (such as the Hewlett-Packard 400D) closely approximates the rms value of the departure of the slope of the transfer characteristic from the constant slope of the ideal straight line. Since the departure of the transfer characteristic of Figure 1 from a straight line necessitates a change in the slope of the transfer characteristic, it is seen that these are related effects. Calculations have shown that the numerical values for percent change in slope, as described above, are always less than the values for percent departure of the transfer characteristic from a straight line. The specific relationship depends on the actual shape of the curve. However, for transfer characteristics experienced in FM modulators and demodulators, the ratio of percent departure from a straight line to percent change in slope from that of a straight line varies from 1/2 to 1/3. Thus, a measurement of 1% change in slope insures a departure of the transfer characteristic from a straight line of less than 1%.

Figure 9 shows a composite block diagram of the entire setup with a few additional features. Instead of separately modulating the Signal Generator under test and the reference Signal Generator with f_s and f_o respectively, if the Signal Generator under test is simultaneously modulated by f_s and f_o (by switching S_1) the change in slope of the frequency modulator of the Signal Generator under test will be superimposed upon that of the FM demodulator. Thus, if the individual ordinates of the display produced by the block diagram in Figure 9 are noted for the case of separate modulation, and then determined for the case of simultaneous modulation of the Signal Generator under test, the algebraic difference of these ordinates will produce a curve which is due to the nonlinearity of the Signal Generator under test only, and does not depend upon that of the frequency demodulator. Note that when switching f_s from one generator to the other, the horizontal polarity of the display is reversed.

Thus, we have developed a circuit and measuring technique which will measure the change in slope of a frequency modulator largely independently of the demodulator, as well as the change in slope of the demodulator independently of the frequency modulator. Many additional refinements are possible to the basic circuit of Figure 9 to speed up the measurement. Some of these features are:

1. An ac voltmeter, calibrated directly in kc deviation may be connected to the output of the discriminator. Methods of calibration are covered in Reference 5.
2. A dc meter may be used to indicate the level of the demodulated magnitude of f_o.
3. If the dc meter indication is kept constant, an ac meter on the output of the nf_s bandpass filter may be calibrated directly in percent change in slope.
4. Switching may be added to facilitate the alternate connection of f_s and f_o to separate or to a common signal generator (S_1 in Figure 9).

A photograph of such equipment, used in the production test of BRC 202H and 202J Signal Generators, is shown in Figure 10. It should, of course, be noted that the FM demodulator used must be preceded by an excellent amplitude limiter, or spurious effects will occur which are not measures of the true nonlinearity of the frequency modulator or demodulator. In the equipment shown, normal circuits have proven satisfactory. The particular discriminator used was centered at 21 mc and is approximately 5 mc wide, so that the resulting change in slope for deviations of ±150 kc is less than 0.2% peak. This high degree of linearity permits direct measurement of signal generator nonlinearity without the need for calculation. In addition, it permits direct measurement of total percent harmonic distortion of a signal generator by the connection of a good distortion analyzer (such as the Hewlett-Packard 330B) to the output of the discriminator.

Figure 11 shows several cases of nonlinearity displayed by the BRC production test equipment shown in Figure 10. The scale calibration is such

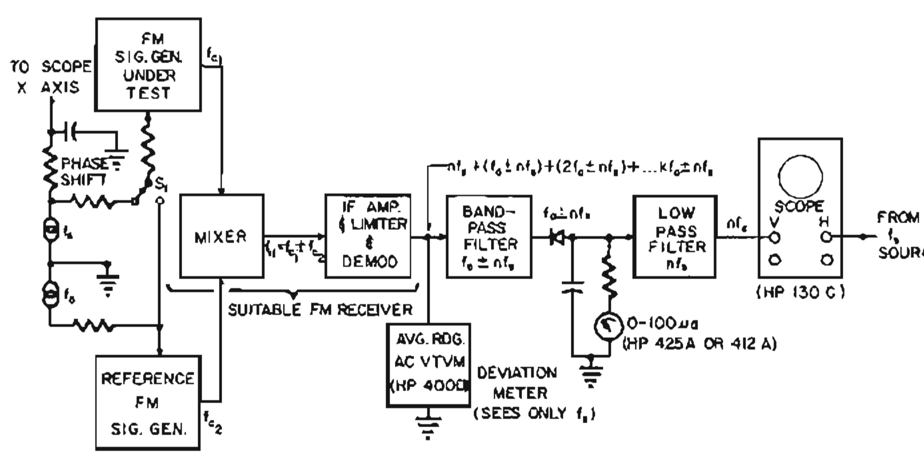

Figure 9. Block Diagram of Circuit Used to Obtain Change of Slope Display

Figure 10. BRC Production Test Station

that a peak-to-peak amplitude of 1.0 cm represents 1% total change in slope.

In this particular equipment, $f_x = 50$ cps and V_{f_x} is set for 150 kc deviation. A frequency of 100 kc was chosen for f_o and E_{f_o} is set for about 15 kc deviation. Almost any other choice of frequencies may be made, subject to the availability of suitable filters; however, f_o should not exceed 100 kc.

Figure 11A shows the change in slope of the BRC production test discriminator alone. This discriminator, plus a properly adjusted 202J Signal Generator, is shown in Figure 11B. Figure 11C shows a typical case of incorrect setting of the operating point of the reactance tube in a 202J Signal Generator. With this display, it is a simple matter to readjust the bias for minimum nonlinearity of the frequency modulator. Resulting changes in carrier frequency must be corrected. A properly set up 202J Signal Generator will show a percent change in slope of less than $\pm 1\frac{1}{2}\%$.

EQUIVALENT FIELD TECHNIQUE

It is possible to utilize the method described above without the complex, specialized equipment shown in Figure 10. This equipment was designed for a particular production test operation and has many features unnecessary to the person who must occasionally adjust a frequency modulator or demodulator for maximum linearity. The circuit of Figure 12 permits the application of our method with readily available commercial instruments. The only special piece of apparatus required is the $f_o \pm nf_x$ bandpass filter and the nf_x low-pass filter which may be readily constructed from the values shown in Figure 12. Should a suitable oscilloscope (such as the Hewlett-Packard 130B or 130C) be available, the refinement of the amplitude demodulator and the nf_x low-pass filter may not be needed.

For FM demodulators of high output impedance, a good electronic voltmeter (HP 400D, etc.) may be used

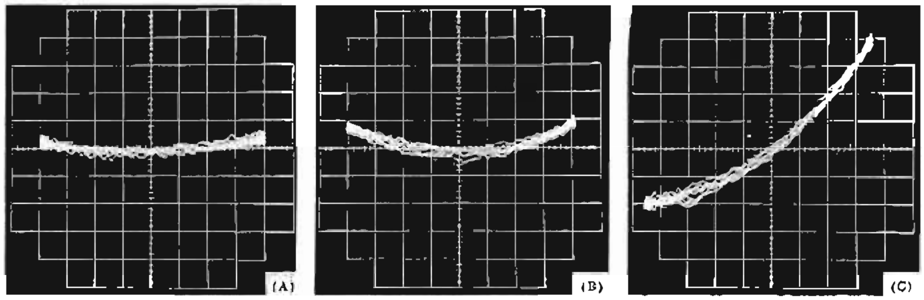

Figure 11. Typical Change of Slope Displays Obtained from Setup in Figure 10.

to provide high input and low output impedance and some attenuation (in 10 db steps). Many telemetry receivers have cathode follower or other low impedance outputs which will drive the 4K ohm bandpass filter impedance adequately. To avoid nonlinearity caused by the voltmeter diodes, the diodes should be shorted out.

The nf_x low-pass filter includes a resonant impedance transformer ($Q \leq 5$) to give a high enough level to the AM demodulator to insure good linearity of the change of slope display. Its output should be checked by the oscilloscope envelope method to insure accuracy.

While it is desirable that the frequency demodulator used be sufficiently linear to permit measuring the signal generator nonlinearity directly, this may be difficult to realize in the field. Therefore, it is essential that the two signal generator method (separately modulated) of Figures 6 and 9 be employed in order to measure or minimize the nonlinearity of the frequency demodulator. After adjustment for minimum nonlinearity, a "grease pencil" trace may be made on the oscilloscope reticule of the nonlinearity (expressed as percent change in slope) of the demodulator. Switching to simultaneous modulation of the signal generator being adjusted will alter the pattern, due to the nonlinearity of this generator. Allowance must be made for the pattern reversal on the horizontal axis, or the polarity of f_x applied to the generator under test may be reversed. It is then necessary to readjust the reactance tube bias of the signal generator to most nearly equal the change of slope curve of the demodulator traced on the oscilloscope reticule. It is obvious that the method becomes inaccurate when the nonlinearity of the demodulator is large compared to that of the signal generator.

RECEIVER ADJUSTMENT

The equipment setup of Figure 12, with one significant addition, can be

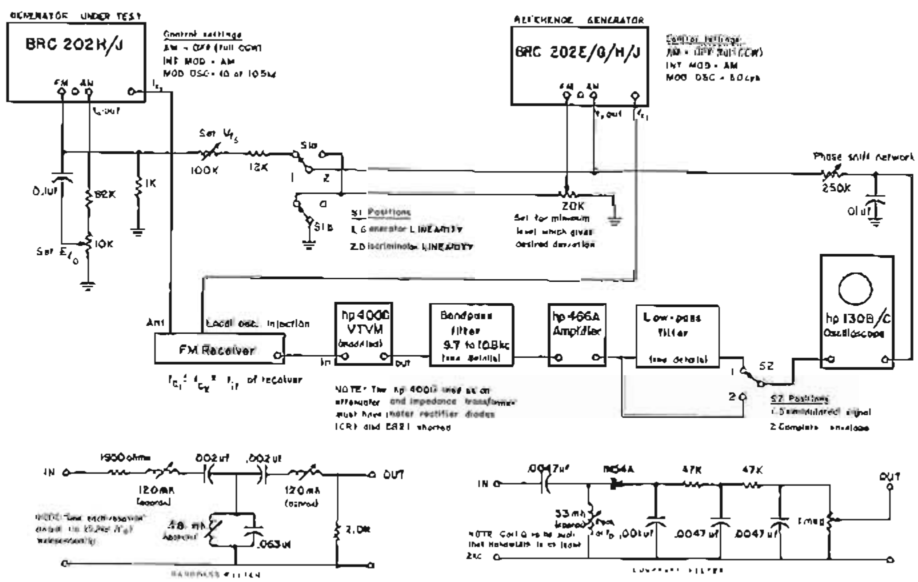

Figure 12. Block Diagram of Setup for Adjusting FM Modulator or Demodulator for Maximum Linearity Using Commercially Available Equipment

used to adjust an FM demodulator for maximum linearity, even in the absence of the complete FM receiver. Substitution of a suitable commercial RF mixer for the receiver front end, followed by one or two of the BRC 230A Signal Generator Power Amplifiers (which provide 24 to 30 db gain) will frequently permit proper drive levels to a limiter and demodulator circuit under development. This use of the 230A facilitates the detailed experimental evalution of an FM limiter and demodulator prior to the availability of the rest of the receiver. Having substituted the 230A and a commercial mixer for the receiver front end, it is only necessary to proceed as previously outlined. In order to minimize the effects of residual amplitude modulation in the FM signal generator, it is desirable to operate the level of the beating signal generator 6 to 10 db lower than that of the frequency-modulated generator. This allows the diode mixer to minimize the unwanted amplitude envelope due to the tuned carrier bandpass circuits in the signal generator.

When displaying the nonlinearity of the demodulator, it is desirable to vary the carrier level of the generator under test with its attenuator to show that proper amplitude limiting is provided ahead of the frequency demodulator.

In setting up the equipment of Figure 12, the use of 202H and J Signal Generators reduces the needed external instruments over that required with other signal generators. Since the 202H and J (and also 202E, G, etc.) include an AF oscillator, it is possible to use this oscillator for the search voltage (E_{f_n}) and the sweep voltage V_{f_s}. The connections of the AM and FM terminals to accomplish this are shown in Figure 12. It is extremely important, though, to keep the AM controls full counterclockwise to avoid unwanted AM of either carrier.

The filter bandwidth is adequate to be used with both the 10 kc AF of the 202H and the 10.5 kc of the 202J.

When aligning a 202H, after following the procedure to achieve minimum nonlinearity, if the demodulator is appreciably more linear than the 202H, a total percent harmonic distortion reading may be meaningfully made.

CONCLUSION

In conclusion, let us summarize the following points:

1. The general definition of distortion resulting from transfer characteristic nonlinearity has been given, together with qualitative relationships between various types of nonlinear distortion and methods of expressing them.

2. A technique for producing an essentially linear frequency-modulation signal from two relatively imperfect signal generators has been described. This technique permits the accurate measurement of demodulator nonlinearity in the presence of practical levels of modulator nonlinearity in terms of the departure of the slope of the transfer characteristic from that of a reference straight line. Having achieved this measurement, it is then possible to measure the nonlinearity of frequency modulators by calculating out, or eliminating, the nonlinearity of the demodulator in the composite display. This method results in an instantaneous dynamic display which facilitates adjustment of either the modulator or demodulator.

3. An equipment setup has been shown which permits the application of the technique in the field with commercially available test equipment, presuming a suitable frequency modulation receiver or demodulator is available.

Similarity of the techniques discussed here to the Intermodulation Distortion method of C. J. LeBel[7] will be seen. However, the numerical results of LeBel's method depend critically on the amplitude ratio of the two signals, while, in the change of slope method, it is only necessary that E_{f_n} be sufficiently small to truly measure slope. V_{f_s} is chosen for the desired deviation being studied.

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2. Appendix to FCC Docket 13506, "Amendment of Part 3 of the Commission's Rules and Regulations to Permit FM Broadcast Stations to Transmit Stereophonic Programs on a Multiplex Basis."
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BRC APPOINTS FOUR NEW ENGINEERING REPRESENTATIVES

Boonton Radio Company recently announced the appointment of four new east-coast engineering representatives: Horman Associates, Inc., RMC Sales Division and Robinson Sales Division of Hewlett-Packard Company, and Yewell Associates, Inc.

Horman Associates, Inc. has its headquarters in Rockville, Maryland, a suburb of Washington, D.C., and a branch office in Baltimore. RMC has two offices: the main office in New York City, and a branch office in Englewood, New Jersey. Robinson Sales Division has three offices, with headquarters located near Philadelphia in West Conshocken, Pennsylvania, and branch offices in Camp Hill, Pennsylvania, and Asbury Park, New Jersey. Yewell Associates, Inc. has its main office in Burlington, Massachusetts, with branches in Middletown, Connecticut, and Poughkeepsie, New York.

With the appointment of RMC, Robinson, Horman, and Yewell, BRC expects to further improve customer services to its many customers. "Chuck" Quinn, who previously had been handling a portion of this territory from the factory, will still be on hand to assist customers with their special application problems.

The complete story on our new representatives, including an introduction to their key personnel, will be presented in the "Meet Our Representatives" series which will be resumed in the next issue of the Notebook. Meantime, we urge our customers in the areas served by these new representatives to call or write them for information about BRC equipment.

A complete list of addresses and telephone numbers for all BRC Engineering Representatives appears on page 8 of this issue.

EDITOR'S NOTE

BRC Assumes Divisional Status

Boonton Radio Corporation, a subsidiary of the Hewlett-Packard Company since 1959, assumed divisional status November 1, 1962. At that time, BRC's name was changed to Boonton Radio Company.

The conversion of BRC to a division of the Hewlett-Packard Co. is a part of an over-all program to achieve greater flexibility of the entire HP organization and to improve operating efficiency. This change will in no way affect BRC policies or product line, but will permit us to offer improved and expanded services to our customers.

Mr. William D. Myers, formerly manufacturing manager of HP's Microwave Division in Palo Alto, has been named general manager of the new Boonton Radio Division. Mr. Myers joined HP in 1944 as a development engineer and has since held a number of executive positions, including manager of quality-control engineering.

A native of San Jose, California, Mr. Myers is an electrical engineering graduate of Stanford University and a senior member of the Institute of Radio Engineers.

William D. Myers

BRC Wins "New Good Neighbor" Award

Boonton Radio Company was one of ten winners in the third annual "New Good Neighbor" contest, a state-wide contest sponsored by New Jersey Business Magazine, a service of the New Jersey Manufacturers Association. Nominations for the contest are made by the mayors of the communities in which new buildings are located. Winners are selected by a panel of noted business and civic leaders and architects. Points considered in the judging are: the general attractiveness of the building and its economic value to the community, and the company's overall community relations approach and effectiveness.

BRC is indeed happy to have been selected a winner in the contest and is proud of its "good neighbor" role in the community.

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BOONTON RADIO COMPANY

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JUN 7 1963

A New Unity Gain Frequency Converter

RICHARD H. BLACKWELL, *Development Engineer*

With the development of the 202H and 202J VHF Signal Generators, the design of an improved version of the 207 Univerter was undertaken. Designated the 207H, this instrument retains the basic concept of earlier models while offering the following additional features:

1. An improved wideband mixer to permit operation with input frequencies above and below the local oscillator frequency.
2. A redesigned local oscillator affording better stability and low residual FM.
3. A built-in 40 db attenuator to aid in making low-level measurements.
4. An -hp- modular cabinet to complement the 202H and 202J.

The 207H Univerter is basically a unity gain frequency converter covering an output frequency range of 100 kc to 55 mc. Figure 2 is a block diagram of the principal functions. A local oscillator frequency of 200 mc was chosen because it falls within the range of both the 202H and 202J Signal Generators, and because it is a convenient figure to use in determining the Univerter output frequency for a known input frequency. The output frequency of the 207H, F_o, is related to the signal generator input frequency, F_i, as follows:

$$F_o = \left|200 - F_i\right| + \left\{\begin{matrix}\text{Frequency Increment}\\ \text{Dial Reading}\end{matrix}\right\} \times 10^{-1}$$

$$145 \leq F_i \leq 199.9 \ (202H)$$
$$200.1 \leq F_i \leq 255 \ (202J)$$

F_o = 207H output frequency in mc.
F_i = Signal generator input frequency in mc.

If the Frequency Increment Dial is set at zero, the output frequency of the 207H is simply the difference between the signal generator input frequency and the 200 mc local oscillator frequency.

When loaded with 50 ohms, the unity gain or X1 output level is within

Figure 1. Type 207H Univerter

± 1 db of the signal generator input level in the operating ranges previously defined. The X.01 output provides a signal level 40 db below that obtained at the X1 output and the High Output provides a minimum level of one volt for 0.1 volt input. The input, X1 output, and X.01 output have a nominal impedance of 50 ohms and the High Output has a nominal impedance of 300 ohms.

MIXER

Previous models of the 207 Univerter were designed to operate only with input signals of a higher frequency than the local oscillator frequency.

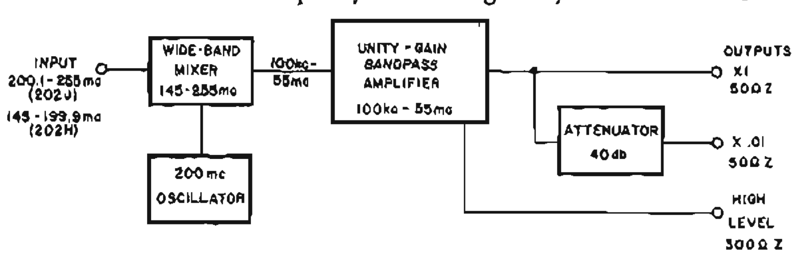

Figure 2. Functional Block Diagram — 207H

Small variations in mixer response over the input frequency range could be compensated for by adjusting the response of the wideband amplifier. In the 207H Univerter, however, any departure from flat mixer response appears as an asymmetry in the upper and lower halves of the overall response curve. The amplifier can be used to correct for mixer non-flatness for either input frequency range, but not for both ranges simultaneously.

Two possible solutions for this problem are: (1) a switching circuit to provide the proper compensation for each range, or (2) a mixer circuit which is sufficiently flat over the input frequency range. The latter approach was chosen.

Figure 3(a) is a schematic diagram of the mixer circuit employed in all previous models of the 207 Univerter. The signal generator input signal is fed to the cathode of a triode mixer and the oscillator signal is applied to the grid by means of a coil coupled

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to the oscillator tank circuit.

Figure 3(b) is a simplified Norton equivalent circuit with the grid leak bias components and all but one of the interelectrode capacitances omitted. The output of the mixer is determined by the conversion transconductance g_c, the signal frequency load impedance Z_L and the grid to cathode voltage V_{gk}. V_{gk}, however, is not always equal to the signal generator input voltage V_{in}, and the relationship between them is frequency dependent. At some frequency F_r, series resonance occurs between C_{gk} and L_c. For input frequencies near F_r, V_{gk} will be larger than V_{in} by a factor which depends, in part, on the Q of the resonant circuit. Because the series resonant circuit forms a feedback circuit between the grid and cathode of the mixer, the actual resonant frequency is slightly higher than the natural resonant frequency. For the circuit of Figure 3(a), F_r is 230 mc. This undesirable resonance effect can theoretically be eliminated by either shifting F_r to a higher frequency (by reducing C_{gk} and L_c), or by reducing the effective Q of the circuit to unity. These possibilities suffer from practical circuit limitations which are difficult to overcome.

A much better solution to this problem would be non-resonant coupling from the mixer to the oscillator. Rather than add a buffer or cathode follower stage, an attempt was made to take the oscillator signal directly from the cathode of the oscillator tube. A small value of resistance in the cathode circuit would provide the RF voltage necessary to saturate the mixer and also discourage mixer grid circuit resonance effects because of the low Q it presents. In addition, the mixer would be operating essentially in the grounded-grid configuration with respect to input signals, resulting in a lower input VSWR.

The result of this development is shown in Figure 4. A 6ER5 VHF triode is used as a mixer because its transconductance can be made to swing over a large range with a small change in grid voltage, thus providing a large conversion transconductance with small drive voltages. Approximately 3 volts rms of 200 mc signal is developed across a 15-ohm resistor in the oscillator cathode. The Q of the resistor varies from 0.6 at 150 mc to unity at 250 mc. The 100-ohm resistor in the cathode of the mixer provides a 50-ohm nominal input impedance.

$$Z_{in} = \frac{R_k}{1 + g_m R_k} = \frac{100}{1 + (.01)(100)} = 50 \text{ ohms}$$

This formula neglects interelectrode capacitance and assumes grounded-grid operation. In prototype models, the maximum input VSWR is 1.7 at 255 mc and the flatness over an input frequency range of 145 mc to 255 mc is approximately ½ db total variation.

200 Mc OSCILLATOR

The Colpitts oscillator circuit of previous Univerters has been improved for use in the 207H. The modifications are:

1. Substitution of a Type 6AF4A tube for the Type 6C4.
2. Use of a metalized glass tank coil.
3. Temperature compensation for improved stability.

The 6AF4A, designed to operate as a UHF oscillator, is more stable and has a higher transconductance than the 6C4. The metalized glass coil is far superior to conventional wire-wound coils in this application. It is unaffected by vibration and humidity and has a maximum temperature coefficent of only 20 parts per million per degree Centigrade. Since the warmup drift is related to temperature rise inside the oscillator compartment, temperature compensation can be employed. The frequency decreases with increasing temperature, thus, requiring a negative temperature coefficient capacitor. Best results are obtained using a grid blocking capacitor whose coefficient is —330 parts per million per degree Centigrade. All other fixed capacitors in the oscillator are of the NPO type. Figure 5 is a table of the drift characteristics of 207H prototype models. The specification allows a maximum drift of 10 kc (.005%) in any one-hour period or 2 kc (.001%) in any five-minute period after a one-hour warmup.

The 207H oscillator circuit has two frequency adjustments. An uncalibrated Frequency Adjust trimmer gives a tuning range of approximately four megacycles to permit zero beating the oscillator with an external standard. This control is a recessed screwdriver type on the front panel. The Frequency Increment capacitor is controlled by the large knob on the front panel and permits a change in frequency of 300 kc either side of the center frequency. The dial is calibrated in 5 kc increments and the accuracy is ± (3% of the dial reading + 1 kc).

The residual FM of the oscillator due to the 60-cycle power line frequency and 120-cycle power supply ripple is 65 db below 10 kc or 6 cps deviation typically.

WIDEBAND AMPLIFIER

If the Univerter is to have unity gain at 50 ohms output impedance, an amplifier must be used to restore the power insertion loss of the mixer. A two-stage, low-pass filter coupled amplifier with an output cathode follower is used. An additional low-pass filter section couples the mixer to the first amplifier.

Each filter consists of a constant K pi section followed by an m-derived half section (m = 0.6) terminated in a resistance which is approximately equal to the characteristic impedance of the filter. The input and output capacitance of the amplifier tubes become the capacitance elements of the filter as shown in Figure 6. Small trimmer capacitors of the glass piston type

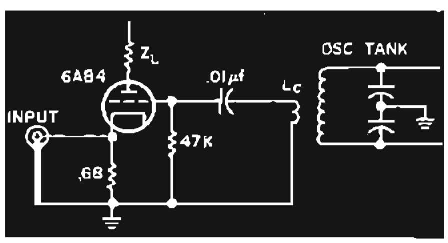


Figure 3(a). Mixer Circuit Used In Previous Univerters

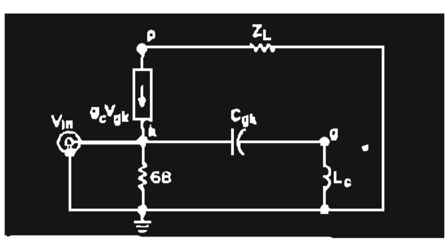


Figure 3(b). Simplified Equivalent Circuit of Figure 3(a) parallel the tube capacitances and compensate for tube variations and component tolerances. The gain that can be obtained for a given bandwidth is limited by the input and output capacitance of the tubes and the tube transconductance. In order to be useful in this circuit, a tube must have a large transconductance combined with small input and output capacitance. Both the 6AK5 and 6688 used in this circuit are suited for wideband amplifier use. Variable resistors for gain control are placed in the cathode circuits of both amplifiers. One of these controls is a recessed screwdriver type on the front panel; the other is a locking potentiometer located at the rear of the casting. The 6AK5 amplifier stage produces a maximum gain of 1.5, and the 6688 has a maximum gain of 8.4. Although the 6AK5 gain seems quite small, it serves to isolate the relatively high input capacitance of the 6688 from the mixer output, and thus permits an additional 6 db of gain in that stage. The final 6AK5 provides two outputs. It acts as a cathode follower to supply the 50-ohm unity gain output and as an additional stage of amplification to supply the high level, 300-ohm output from the plate circuit. The High Output must be loaded externally with a 10 pf capacitance if maximum output flatness is desired. (The resistance loading is not critical). Although the low-frequency limit of the 207H is specified as 100 kc, the response extends down into the audio range to facilitate zero beating the oscillator with a signal generator oscillator, using headphones or a VTVM as a null indicator. The premium quality 6688 frame grid pentode not only produces a large gain, but offers reliable operation and long life. The mixer and amplifier together provide flat response within ± 0.7 db over the entire operating frequency range.

ATTENUATOR

A mixer, followed by a wideband amplifier, has an inherently high output noise level. The noise power output is proportional to bandwidth within the passband of the amplifier. For the 207H Univerter, the noise level at the unity gain output is a maximum of eight microvolts over a one-megacycle bandwidth. This corresponds to a noise figure of approximately 25 db. This noise level can be troublesome when making measurements with sensitive,

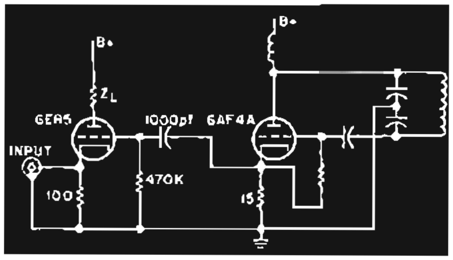


Figure 4. 207H Univerter Mixer Circuit

TIME INTERVAL (T = 0, Cold Start)	T = 60 Min To T = 65 Min	T = 60 Min To T = 120 Min.	T = 5 Min. To T = 60 Min.
Published Specification	< .001%	< .005%	None
Best Unit	.0001%	.001%	.0077%
Worst Unit	.00035%	.007%	.004%

Figure 5. Local Oscillator Stability — Typical Performance (Nominal Oscillator Frequency is 200mc)

wideband devices. A simple solution to this problem is the use of an attenuator between the wideband amplifier and the device being tested at low signal levels. A 40-db attenuator for this purpose has been incorporated in the design of the 207H.

Figure 7 shows the multisection frequency compensated attenuator which serves a dual purpose. It provides 3.5 db attenuation between the X1 output and the cathode follower output of the amplifier in order to obtain a low VSWR. It also provides a X.01 output which gives a level 40 db below the X1 output level. Output levels are specified across a 50-ohm load connected to the output in use. Both outputs should never be loaded at the same time or a serious error in attenuation will result.

The resistors used are half watt, one percent, carbon film types (MIL RN 20X). The maximum possible error in attenuation due to the resistance tolerance is ± 0.4 db. Change in attenuation with frequency, due to the rising impedance of the 10-ohm resistors at high frequencies, is compensated with shunt capacitors. The total maximum error in attenuation, due to both effects, is approximately 1 db. Typical attenuators have errors of less than ½ db. The attenuator exhibits a rise in attenuation above 55 mc. This is desirable because the transmission of spurious signals above 55 mc is reduced. For instance, at 200 mc the attenuation is 60 db. The X1 output has a maximum VSWR of 1.22 and the X.01 output has a maximum VSWR of 1.17.

The attenuated output should be used for making measurements at levels below 1000 microvolts. The attenuated output noise power is less than the noise produced by a 50-ohm resistor at room temperature. Therefore, the X.01 output noise power is essentially only that associated with the 50-ohm internal resistance.

POWER SUPPLY

Two power supplies are available with the 207H Univerter, a 95-130 volt, 60-cycle model and a 95-130 volt or 190-260 volt, 50-cycle model. The dual voltage supply has a voltage changeover switch mounted on the power supply chassis. Both supplies employ resonant stabilizers for ± 1% voltage stabilization over the indicated range of input line voltages. The B+ is developed by a conventional voltage doubler circuit, using selenium rectifiers and a two-section, choke-capacitor filter. As a result of voltage stabilization, the local oscillator frequency change, due to a 1-volt change in line voltage, is less than 400 cycles.

PHYSICAL CHARACTERISTICS

The oscillator, mixer and amplifier are constructed on a silver plated brass plate mounted on an aluminum casting with a silver plated brass cover plate. The shielded attenuator subassembly is mounted to the side of the casting, while the regulated power supply is a separate chassis. The entire unit is housed in the new Hewlett-Packard Modular Cabinet. This cabinet matches

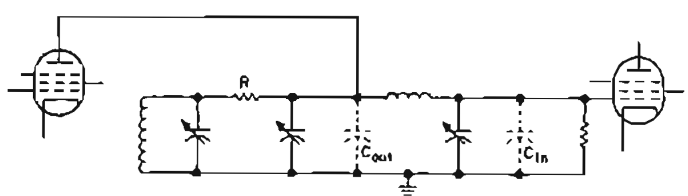


Figure 6. Wideband Amplifier Interstage Filters (Bypass and Coupling Capacitors Omitted for Simplicity)

the appearance of the 202H and 202J cabinets and permits stacking the 207H with either generator.. The simple addition of flanges permits rack-mounting. The front panel layout of the 207H is designed to complement the appearance of the 202H and 202J front panels when the units are stacked. A short accessory cable, Type 524A, is used to connect the signal generator output to the Univerter input.

It is desirable to have the Frequency Increment Dial indicate the sense of output frequency change as well as magnitude. The sense is opposite for input frequencies above and below 200 mc. To avoid possible confusion, the input frequency ranges are color coded to correspond to the appropriate Frequency Increment Dial calibrations.

OPERATION WITH 202H AND 202J

The major advantage of the Univerter principle is the extension of the superior modulation characteristics and precision piston attenuator of the VHF signal generator to the lower frequency range.

The 207H will reproduce the modulation of the 202H or 202J, with negligible distortion, provided the following precautions are observed. Care should be taken when using low carrier frequencies that significant modulation sidebands do not fall below 100 kc, otherwise severe distortion may result. The following simple rules will avoid this condition:

Modulation—Lowest Permissible Output Carrier Frequency
AM—100 kc plus Modulation Frequency
FM—100 kc Modulation Frequency plus Deviation Frequency

In addition, input amplitude modulated signal levels should be kept below .05 volts for minimum envelope distortion.

The X1 output level of the 207H Univerter can be read directly from the 202H or 202J attenuator dial with an accuracy of ± 1 db plus the accuracy of the signal generator attenuator itself. In this way the 207H effectively extends the range of the 202H or 202J precision piston attenuator to cover frequencies of 100 kc to 55 mc.

The stability of the output signal of the 207H depends upon the stability of the 207H local oscillator and the stability of the signal generator with which it is used. Much effort has been

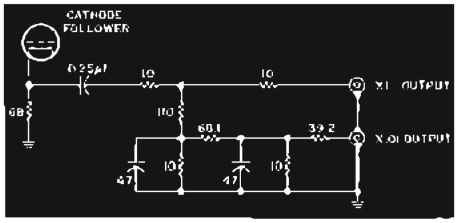


Figure 7. 40 db Attenuator

put into stabilizing the 207H local oscillator in order that the output frequency stability of the Univerter will be controlled almost entirely by the stability of the signal generator. Very little could be gained in output frequency stability with a crystal controlled local oscillator. In addition a crystal controlled oscillator would preclude the use of the Frequency Increment capacitor for calibrated frequency deviations of ± 300 kc. The drift specifications for the 207H refer only to the local oscillator and not to the output frequency.

$$(\%\ \text{change in } F_o) = \frac{(\%\ \text{increase in } F_I)\ F_I - (\%\ \text{increase in } F_L)\ F_L}{F_o}$$

F_I = sig. gen. input freq. in mc.
F_L = local oscillator freq. in mc.
F_o = output freq. in mc.

The percent change in output frequency is a function of the magnitude of input and local oscillator drift, the direction of the drift and the output frequency itself. It is possible, for the output frequency drift to be zero while both the input and local oscillator frequencies are changing.

For some applications, especially at low frequencies, the output frequency drift may be larger than desirable. It is possible to lock the output frequency of the 207H to an external discriminator using a simple AFC arrangement. The dc output of the discriminator must be amplified for best results and aplied to the DC FM INPUT of the 202H or 202J. Care must be taken to use the proper polarity of feedback signal. Figure 8 shows the recommended setup. The dc amplifier should have a high input impedance, a low output impedance, a polarity reversing switch and a gain of at least 15. Both the discriminator and amplifier should be as stable as possible. The time constant of the dc amplifier must be short enough to prevent "hunting" and long enough to prevent carrier demodulation and the introduction of FM hum modulation. Experimentation may be necessary to determine the best value for a given application.

Spurious output frequencies from the 207H Univerter result mainly from signal generator spurious outputs which are converted to lower frequencies along with the desired signal and appear in the output.

The total harmonic distortion of the 207H is less than 2.5% at a level of 0.1 volts. The second and third harmonics are at least 30 db below the

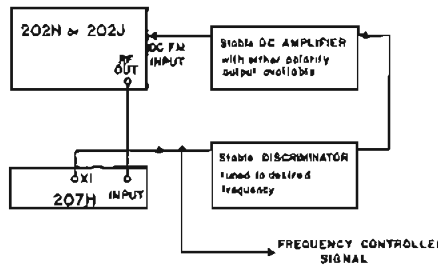


Figure 8. AFC Circuit Using External Discriminator

fundamental for levels up to 0.1 volt. Approximately 200 microvolts of 200 mc local oscillator signal appears at the unity gain output. Output frequency signal leakage is at least 60 db below the unity gain output level in the vicinity of the output panel connectors.

The results of a series of environmental tests indicate that the 207H is capable of withstanding any shock, temperature and humidity conditions likely to be encountered in normal laboratory use.

Figure 9. Typical 207H and 202H Setup

SUMMARY

The 207H is a valuable accessory to the 202H and 202J Signal Generators. The three instruments together offer calibrated output levels with both AM and FM modulation over a frequency range of 100 kc to 270 mc.

CORRECTION

The block diagram shown in Figure 9, Page 5 of Notebook Number 33 is not correct as shown. The blocks designated "FM SIG. GEN. UNDER TEST" and "REFERENCE FM SIG. GEN.", together with the output designations "f_{c_1}" and f_{c_2}", should be interchanged. The Notebook is indebted to Mr. K. E. Farr of Jerrold Electronics Corp. for pointing out this error.

New Techniques in FM Fidelity Measurements

RICHARD N. SCHULTE, *Production Engineer*

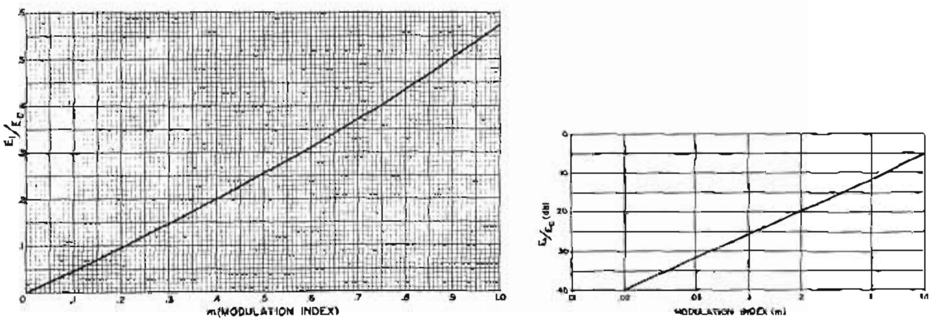
INTRODUCTION

Common methods of measuring a signal source FM fidelity involve the use of a receiver or detector with known fidelity characteristics. This article describes a method of determining fidelity by measuring the deviation of an FM source as a function of a constant amplitude modulating signal, without dependence on the receiver's fidelity characteristic. Any relatively good narrow band AM receiver will suffice for the measurement since the fidelity requirement is not critical.

In the absence of AM, the frequency spectrum of an FM signal shows the amplitude relationship between the carrier and the various sidebands. The modulation index, m, is defined as $\Delta f/f_{mod}$; where Δf is the peak frequency deviation of the carrier from its center frequency, and f_{mod} is the modulation frequency.

The various carrier and sideband amplitudes that result from values of m are related to the Bessel Functions of the first-kind, J_n (m), with order equal to n where nf_{mod} equals the separation between the carrier frequency and the sideband or spectrum component of order n.

The carrier amplitude is $E_c = E_oJ_o$ (m), the first-order sideband is $E_1 = E_oJ_1$ (m), and the n_{th} order sideband is $E_n = E_oJ_n$ (m); where E_o is the amplitude of the unmodulated carrier.



(a) E_1/E_c in Absolute Values (b) E_1/E_c in Decibels

Figure 1. Amplitude Ratio — m vs. E_1/E_c

BESSEL ZERO METHOD

For years, frequency deviation has been measured by using the fact that the carrier amplitude, related to the Bessel Function J_o (m), goes to zero at certain values of modulation index (m= 2.405, 5.520, 8.653, 11.79, etc.). For this type of measurement all that is needed is an accurately known modulation frequency source, and a receiver selective enough to precisely indicate the carrier null in the presence of a first-order sideband. The procedure is simply to tune to the carrier with no modulation present, and then increase the amplitude of the modulating signal until the desired n_{th} order null is reached as indicated by the disappearance of the carrier, for the n_{th} time[1]. For example, the deviation of an FM signal source can be set to 150 kc using the second-order Bessel Zero modulation index (m) of 5.520. In this case the modulation frequency used is

$$f_{mod} = \frac{\Delta f}{m} = \frac{150 \text{ kc}}{5.520} = 27{,}173 \text{ cps.}$$

Tuning to the carrier, without modulation, and then increasing the amplitude of a 27,173 cps modulating signal until the second null is reached, will set the deviation of the signal source to 150 kc. Table 1 contains frequently used Bessel Zero frequencies and resulting deviations.

The minimum receiver bandwidth that can be used with a VHF FM signal source of good stability limits the modulation frequency to a minimum of about 5 kc. This puts a limit of approximately 12 kc on the minimum deviation that can be measured by Bessel carrier zeroes. On the other hand, nulls above the 4th order become difficult to identify and precisely locate. This limitation is not serious, however, because most FM telemetry and entertainment receivers are quite flat, before de-emphasis, from 50 cps to 15 kc. Most FM Signal Generators are also quite flat in this frequency range.

DEVIATION	NULL ORDER		
(kc)	1	2	3
50	20,792	9,058	5,778
75	31,188	13,587	8,667
150	62,375	27,173	17,334
250	103,959	45,289	28,889
300	124,750	54,347	34,667

Table 1. Bessel Zero Modulating Freq.

The first-order Bessel Zero appears at a modulation index of 2.405, which is the minimum value useable with the Bessel Zero method. Obviously, the Bessel Zero method fails, and hence a problem arises at modulation indices less than 2.405. For instance, a 50 kc deviation at a modulating frequency of 500 kc gives a modulation index of 0.1 and is not measurable by the Bessel Zero method.

SIDEBAND AMPLITUDE METHOD

Analyzing the Spectrum

A ratio measurement of the first-order sideband to carrier amplitudes fills this gap. The ratio under consideration, E_1/E_c, equals $J_1(m)/J_0(m)$, because $E_1 = E_0J_1(m)$ and $E_c = E_0J_0(m)$ and therefore, E_1/E_c is a function of the modulation index ($m = \frac{\Delta f}{f_{mod}}$). Therefore, the actual deviation can be determinded since f_{mod} is known and m can be calculated from E_1/E_c, which equals $J_1(m)/J_0(m)$. If $J_1(m)/J_0(m)$ is known, m can be found in any table of Bessel functions of the first kind. When m is less than 0.5, a good approximation is $\frac{J_1(m)}{J_0(m)} = \frac{m}{2}$. For convenience, m vs. $J_1(m)/J_0(m)$ is plotted in Figure 1. The values of $J_1(m)/J_0(m)$ or E_1/E_c, for some of the more common modulation indices, are listed in Table 2.

m	Δf (kc)	f_{mod} (kc)	E_1/E_c RATIO	E_1/E_c db
.5	50	100	.258	—11.7
.25	50	200	.125	—18
.1	50	500	.050	—26
.05	50	1000 (1 mc)	.025	—32

Table 2. Modulation Indices —

$$m = \frac{\Delta f}{f_{mod}}$$

Effects of AM Distortion

For the sideband method to give accurate results, the FM spectrum can not be distorted by AM. Residual AM on an FM spectrum usually increases the amplitude of one sideband and decreases the amplitude of the other, as shown in Figure 2. If the FM signal is distorted by incidental AM, the IF signal obtained by beating it with a local oscillator will also have AM distortion. This distortion can be minimized by adjusting the relative RF levels of the AM distorted FM signal and local oscillator. If the distorted signal is made large enough to operate the diode in its saturated region, clipping by the diode will reduce the AM distortion. Reducing the AM content, while maintaining a constant IF amplitude, can be obtained by increasing the level of the distorted signal and decreasing the level of the local oscillator.

Receiver Requirements

The receiver to be used must be selective enough to locate the carrier in

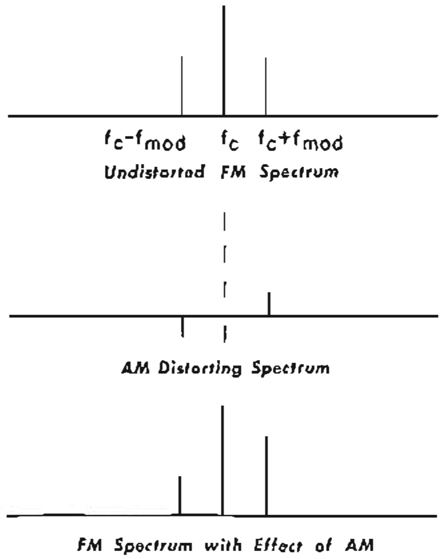


Figure 2.

the presence of first-order sidebands for modulation frequencies down to 9 kc. Beating the RF signals to an IF carrier allows the use of a receiver with average selectivity. A receiver with adjustable selectivity, such as the Hammarlund SP-600, makes it possible to adjust the bandwidth as the modulation frequency varies. For example, with a 20 mc carrier and a 10 kc modulation frequency, the receiver must be able to distinguish the carrier (20 mc) from the first-order sidebands (19.99 mc and 20.01 mc). At higher modulation frequencies, the spectrum components have wider spacing and the receiver bandwidth can be increased for easier tuning to the carrier and sidebands.

The RF amplitude of the sideband carrier is indicated on a VTVM connected to the second detector output of the receiver, which should vary in a somewhat linear manner with the signal. The meter does not measure the absolute amplitudes of the carrier and sidebands, but it is used to set and match equal levels at the receiver input; therefore, its nonlinearity is not too important to the measurement.

Sideband-to-Carrier Ratio

There are many ways to measure the sideband-to-carrier ratio, two of which will be discussed. The first, and probably the faster, shown in Figure 3, uses step attenuators to determine the amount of attenuation needed to reduce the carrier amplitude down to the sideband amplitude. This gives the ratio E_1/E_c directly in decibels.

The second method (Figure 4) uses a reference generator operating at the Intermediate Frequency, matching its output level (as indicated on the VTVM), to the levels of the carrier and sidebands.

Using either method, the modulation signal applied to the FM generator must be held at a constant amplitude over the range of modulation frequencies used.

Setting Reference Deviation

The modulation signal amplitude is set for each carrier frequency to be tested by making a Bessel Zero calibration for the reference deviation being used. Using a frequency counter, set the frequency of the modulation source (-hp- 650A Test Oscillator). Tune the receiver to the carrier with no modulation, and then increase the amplitude of the modulation signal until the desired order null occurs. The null can be detected on the meter or with earphones using the receiver BFO. This amplitude of modulation signal will be used for the test of the measurements and the results will be based on a known reference deviation. The reference deviation can be set to 50 kc by setting the modulation signal amplitude

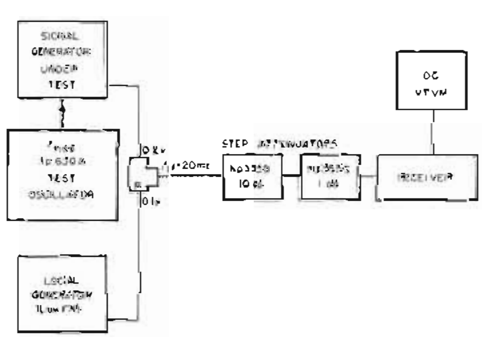


Figure 3. Fidelity Measurement — Step Attenuator Method

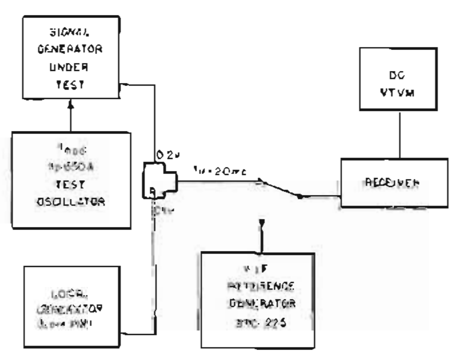


Figure 4. Fidelity Measurement — Reference Generator Method

for the second-order Bessel Zero at a modulation frequency of 9.058 kc.

$$f_{mod} = \frac{\Delta f}{m} = \frac{50 \text{ kc}}{5.520} = 9.058 \text{ kc}$$

After a Bessel Zero calibration is made for 50 kc deviation, it is now known that at a modulation frequency of approximately 10 kc, "X" volts of modulation signal gives 50 kc deviation. Now that the amplitude of the modulation signal has been set, the fidelity measurements can be made. We want to know how the deviation will differ from 50 kc for "X" volts of modulation signal at other modulation frequencies.

Using Step Attenuators

Referring to Figure 3, the ratio E_1/E_c is determined by the amount of attenuation needed to reduce the carrier amplitude to the amplitude of the first-order sideband. The step-by-step procedure for making this determination is as follows:

1. Calculate the theoretical value of the modulation index (m) for the deviation and modulation frequency being used.
2. Using Figure 1B, convert m to E_1/E_c and set the step attenuators to a value at least 3 db greater than this value.
3. Tune the receiver to the unmodulated carrier.
4. Apply the modulation signal and adjust its level to the same value obtained for the reference deviation Bessel Zero.
5. Adjust the receiver RF gain to give a convenient upscale reading on the VTVM.
6. Tune the receiver to a first-order sideband ($F_{if} \pm f_{mod}$).
7. Reduce the step attenuator settings to give the *same* VTVM indication as for the modulated carrier. (The -hp- 355C step attenuator has 1 db steps, however, the VTVM reading can be interpolated to at least one-quarter db).
8. The amount of attenuation removed to adjust the VTVM sideband indication to the same indication as the carrier, is the actual ratio, E_1/E_c, in decibels.
9. This value, minus the theoretical value, is the departure of the signal source from a perfectly flat fidelity characteristic.
10. Both upper and lower sidebands should be checked to make sure the IF signal is properly limited and not distorted by AM. If the amplitudes of the upper and lower sidebands are within 1 db, the average can be used to determine the ratio E_1/E_c.

Checking the fidelity at $f_{mod} = 500$ kc, and $\Delta f = 50$ kc, might produce the results shown in the following example. Under these conditions the theoretical modulation index is

$$m = \frac{\Delta f}{f_{mod}} = \frac{50 \text{ kc}}{500 \text{ kc}} = 0.10$$

From Figure 1 or Table 2, for m = 0.10, E_1/E_c theoretically equals 0.05, or —26 db.

The BRC Type 202J FM fidelity specification is ± 1 db from 5 cps to 500 kc, therefore at $f_{mod} = 500$ kc and $\Delta f = 50$ kc, a E_1/E_c ratio of —25 db to —27 db would be within limits.

Suppose the ratio E_1/E_c was —26.5 db. Since Figure 1 shows a nearly linear relationship between m and E_1/E_c at low values, if E_1/E_c is one-half db (6%) low, m would also be 6% low. Then, the actual modulation index would be 0.094 compared to the theoretical value of 0.1. This means that "X" volts of modulation, which gave 50 kc deviation at approximately 10 kc, would not give 50 kc deviation at $f_{mod} = 500$ kc, but would actually give 6% less deviation or

$\Delta f = f_{mod} = .094 \times 500 \text{ kc} = 47$ kc.

Using a Reference Signal

The reference generator method uses the same IF and procedure for determining modulation signal amplitude. In this set-up, Figure 4, the calibrated output of the reference generator is matched to the carrier and sideband amplitudes. The step-by-step procedure follows.

1. Calculate the theoretical value of m for the deviation and modulation frequency being used.
2. Convert m to E_1/E_c.
3. Tune the receiver to the unmodulated IF carrier of the beating RF signals.
4. Apply the modulation frequency signal and adjust its amplitude to the level obtained for the reference deviation by the Bessel Zero calibration.
5. Adjust the receiver RF gain to give a convenient upscale reading on the VTVM.
6. Switch the receiver to the Reference Generator (tuned to the carrier frequency) and adjust the attenuator for the same vtvm reading as in step 5. Note the attenuator setting.
7. Switch back to the Beating Generators and tune the receiver to the first upper (or lower) sideband.
8. Adjust the receiver RF gain and/or the VTVM range to get an upscale indication.
9. Switch to the Reference Generator and tune it to the Sideband Frequency ($F_{if} \pm f_{mod}$). Adjust the Reference Generator attenuator for the same VTVM reading as for the sideband. (Note Attenuator Setting.)
10. The ratio of the Reference Generator attenuator settings (sideband amplitude divided by carrier amplitude) equals E_1/E_c. The difference in attenuator settings (on the decibel scale) also equals E_1/E_c in decibels.
11. Both upper and lower sidebands should be checked to make sure the IF signal is properly limited and not distorted by AM. If the amplitudes of the upper and lower sidebands are within 1 db, the average can be used to determine the ratio E_1/E_c.

As an example, for $f_{mod} = 200$ kc and $\Delta f = 50$ kc; m = 0.25. Referring to the graphs in Figure 1 or to Table 2, m = 0.25 results in a sideband-to-carrier ratio (E_1/E_c) of 0.125 or —18 db. Therefore, if the modulated carrier amplitude equals 2 K μv (Step 6), the first-order sideband should equal 2K μv × 0.125 = 250 μv (Step 9).

A fidelity specification of ± 1 db at $f_{mod} = 200$ kc means the sideband will be within ± 12% of its ideal values, or 220 to 280 μv, with m = .25 and 2 K μv carrier. Again, using the linear approximation of Figure 1, suppose the ratio of E_1/E_c measured 0.120; 4% below the desired value of 0.125 at $f_{mod} = 200$ kc. In this case, m would also be 4% low, but for "X" volts of modulation signal at $f_{mod} = 200$ kc, the deviation would be 48 kc, or 4% less than the 50 kc deviation observed for "X" volts of modulation signal at $f_{mod} = 10$ kc

CONCLUSION

The numerical examples used in this article are based on measurements made on the BRC 202J Telemetering Signal Generator. However, the concept of the sideband amplitude method of measuring FM fidelity is applicable to any frequency-modulated signal or source.

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EDITOR'S NOTE

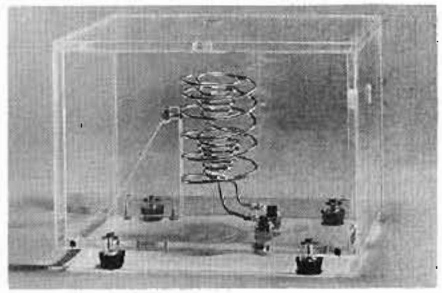
Q Contest Winner

The Q of the coil displayed in the BRC booth at the 1963 IEEE show was 313, as measured on the BRC Type 260-A Q Meter. Two estimates of 313 were actually submitted: one by Mr. Seymour Krevsky of RCA Surfcom Laboratory, and the other by Mr. E. A. Zizzo of the Polytechnic Institute of Brooklyn. In accordance with our contest rules, a drawing was made and we are pleased to announce that the winner is Seymour Krevsky. It is also interesting to note that Mr. Krevsky was a near winner in 1957 and again in 1959, when his estimates were just a shade off the actual value.

Nearly 1000 estimates were submitted, ranging from zero to infinity. In addition to the 313 estimates, there were ten estimates within 1% of the measured Q. A list of the persons who submitted these estimates is given below.

Estimate	*Submitted By*
310	T. D. MacCoun, Budelman Electronics Corp.
311.5	J. H. Humphries, Western Electric Co.
312	"Nick", Tarry Electronics
312	D. Bickor, Sperry Gyroscope Co.
312	R. Lafferty, Boonton Electronics Corp.
314	R. Dormagen, E. Stanwyck Coil Co.
314.16	F. Kilkenny, RCA Institute
314.2	H. P. Hall, General Radio Co.
315	F. J. Logan, NASA, Goddard Space Flight Center
315	R. Haindel, New York University

A photograph of the display coil is shown here for those Notebook readers who did not see it at the show. The unusual configuration, consisting of two conical-wound coils wound inside a helical-wound coil, was devised by "Chuck" Quinn, BRC Sales Engineer.



Q Contest Coil

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NUMBER 35

The NOTEBOOK

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Precision Peak Power Measurements With The Peak Power Calibrator

G. RAYMOND POLEN, *Development Engineer*

INTRODUCTION

The need for accurate measurements of peak RF power of pulsed sources, while having existed for over a score of years, has prompted surprisingly little in the way of simple reliable commercial equipment for performing the task. Today with the increasing number of electronic systems such as radar, air navigation, telemetry, communications, command and control, television, radiosonde, and many others depending on pulsed RF signals, the need is greater than ever. Yet, in many instances, the systems engineer must devise his own method of peak power measurement. While some of these systems are fairly accurate, they are generally time consuming and expensive and often completely unsuitable for high volume or production line measurements. The time factor is an important one, not only from the viewpoint of time efficiency, but from the viewpoint of accuracy, for it is axiomatic in this type of measurement that time and error are quite directly related. Other criticisms of current methods have been that they exhibit a high degree of temperature sensitivity and a rather unwieldy procedure for recalibration.

YOU WILL FIND . . .

Figure 1. Type 8900A Peak Power Calibrator

The techniques of CW power measurement have been quite steadily advanced over the years and to take advantage of this, peak power measurements are often a correlation process in which the performance of a device under the application of pulsed RF power is calibrated to a characteristic it exhibits upon application of a known CW power level.

DEFINITION OF PEAK POWER

Occasionally there is some confusion in formulation of a concept of what peak power actually is. A relationship accepted by groups working in the field is:

$$P_{ave} = P_{peak} \times \text{Duty Cycle}.$$

Duty cycle is the fractional time a pulsed source is turned on. If the source were turned on 100% of the time, the duty cycle would be 1 and peak power and average power would be equal. Peak power, then, could be explained as the average power that would exist if the pulsed source were left on all the time. It is not the instantaneous peak power or envelope peak power that exists at the peak of the RF voltage waveform. Assuming a sinusoidal CW source with an average power of 1 watt, the peak power rating of the source is 1 watt also. If the source is turned off 50% of the time, the average power will be ½ watt, whereas the peak power rating remains 1 watt.

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METHODS FOR MEASURING PEAK POWER

Rearranging factors in the original equation:

$$P_{peak} = \frac{P_{ave}}{\text{Duty Cycle}}.$$

From this equation, it is apparent that one method of pursuing the problem of peak power measurement is to measure average power and duty cycle and correlate the two. While average power can be measured with a fair degree of confidence, duty cycle can become an elusive parameter. Measurement of duty cycle requires that a decision be made as to when a source is "on" and when it is "off". In some systems employing complex waveshapes, this can be an arbitrary decision if left up to electronic circuitry. In our air navigation distance measuring systems (DME), for example, employing a gaussian shaped pulse, when should it be decided that the pulse if "off"? In a television system, where intelligence is being transmitted by both time and amplitude modulation of the pulsed source, there is a similar or even more demanding problem. A versatile peak power measuring system must not leave the determination of duty cycle up to the unimaginative mind of electronic circuitry, because the possible errors due to variations in pulse width, rate, or shape are too great.

Consider, then, another means of determining the peak power of a pulsed RF source. For a sinusoidal CW source it is generally agreed that:

$$P_{ave} = \frac{E_{rms}^2}{R_o} = \frac{(.707E_{peak})^2}{R_o} = \frac{(E_{peak})^2}{2\,R_o}.$$

As mentioned before, the average power and peak power of a source are equal if the duty cycle is 1. The voltage waveforms, then, must be identical for the duration of time the source is turned on, assuming a system of constant characteristic impedance or R_o. The peak voltage of the CW mode will be the same as the peak voltage when the source is pulsed, even though this is not the parameter to be measured. It is, however, a means of correlation between the two. Employing a device known as the peak detector, we have an element which will respond identically for a duration of time to both a CW and a pulsed source of the same power rating. While elements such as a bolometer may respond quite differently to the two, depending on the heating effect, the peak detector does not. This is the principle of the Boonton Radio Company Type 8900A Peak Power Calibrator.

TYPE 8900A PEAK POWER CALIBRATOR

Figure 2 is a block diagram illustrating the basic operation of the 8900A. It can be seen that a signal applied to the front panel input connector is sent through two paths by virtue of the power divider. In one path, the signal passes through a 10 db attenuator and is absorbed in a 50-ohm termination. In the other path, the signal arrives at a diode peak detector which develops a dc level equal to the peak voltage of the RF waveform applied to it. The diode is forward biased to bring it to an operating point of maximum stability and to bring it away from the square-law region to produce a somewhat more linear change in output voltage for a change in the appied RF level. As the diagram indicates, a variable dc supply is included also. The output of the supply is connected to a dc meter, which monitors its voltage, and to one leg of a mechanical chopper. If the chopper is set in operation and its selecting arm is connected to an oscilloscope, one can look, first at the dc level produced by the peak detector in response to an RF voltage, and then at the dc level from the variable supply. In operation, the supply is adjusted until the two voltages are exactly equal. The dc meter monitoring the output of the variable supply has been calibrated in terms of RF level required to produce a given dc from the peak detector and hence peak RF power can be read from it directly. CW power, of course, is correctly indicated also, since the calibration is in terms of the peak voltage waveform.

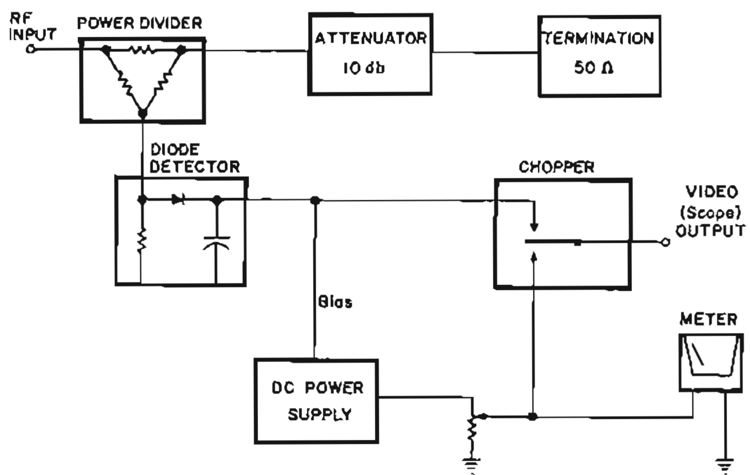

Figure 2. Block Diagram - Type 8900A

The peak detector has a very important job, and if it does not do this job well, errors will be introduced. The output capacitor must be charged to the true peak of the waveform within the duration of the pulse, though not necessarily on the first cycle of the RF carrier. It may charge up in staircase fashion over a period of several cycles, but must reach the peak before a measurement is made. The 8900A specification states that 0.25 μs should be allowed for this, although 0.10 μs is typical with normal cable lengths and oscilloscopes connected to the video output. Since no isolating amplifier is employed, extremely high external capacitance could increase the time required for the video output to rise to the true peak of the voltage waveform. The impedance of the output circuit is approximately 150 ohms. The peak detector also has a responsibility to remain faithful at low RF carrier frequencies. It must not start to discharge while it is waiting for the crest of the next cycle of the RF voltage waveform to appear. If it falls by even a few per cent, the dc output level, within the duration of the pulse, would be lower than the true RF voltage peak and an error would be introduced. The 8900A was designed to meet its accuracy specification at carrier frequencies down to 50 mc and has been found capable of doing this. The preliminary catalog lower limit of specification was placed at 150 mc as a gesture of conservarism. In a like manner, the preliminary specification of upper frequency limit is 1500 mc, although all units tested have been found to be within the accuracy limit up to 2.0 Gc.

Figure 1 is a photograph of the front panel of the 8900A. The 5½ inch meter actually occupies almost two-thirds of the front panel and was included to enable the user to take full advantage of the accuracy and stability of the instrument with readout easily to 0.1 db. Meter tracking accuracy and repeatibility are of necessity a tightly controlled characteristic of the unit. A front panel "NULL" control is included to permit the user to "erase" the static dc bias on the diode from the video presentation. While this control need not be reset for a repetitive series of measurements, it gives the operator a range of adjustment to compensate for any possible long term aging effects on the diode. This adjustment is made with the function switch in the "CAL" position. In this position, also, a voltage divider from the reference power suppy applies a preset voltage to the dc meter to deflect the needle to a calibration mark. This was included to give the operator confidence that the dc meter and power supply are operating properly should he question it at any time during a measurement. It should be noted, however, that unlike some measuring systems, the reference supply is not really a critical parameter in the measurement because it is being used only for comparison rather than as an absolute reference for the measurement. The meter calibration is the absolute reference.

Now consider the signal path which attenuates the incident power and dissipates the remainder in a 50-ohm load. This is provided as a convenient means of calibrating or standardizing the instrument. If the 50-ohm load is replaced by an accurate CW power measuring device such as a bolometer or calorimeter, and a CW source is connected to the input connector, the effect of the applied power level can be monitored on the average reading CW standard and the peak reading diode detector simultaneously. Therefore, one need only to know accurately the attenuation between the front panel input connector and the CW standard to determine what effect a known power level at the input has upon the peak RF detector. The 10 db pad was introduced merely to reduce the CW level to one within the range of several commercially avaiable standards. The -hp- 431 Power Meter, with the 478A Bolometer or the -hp- 434 Calorimeter, are quite satisfactory for this application. The CW source requirements also are met by readily available units.

A basic objective of the 8900A is to provide a peak power measurement instrument of sufficient accuracy to serve as a working standard, without the usual rigorous limitations of standards laboratory equipment. The high quality forward-biased diode is an order of magnitude more stable environmentally than the uncompensated bolometer or calorimeter. Operator skill level, also, has been reduced to absolute minimum. The objective then is to "capture" the established accuracy of known standards and to faithfully repeat this knowledge under a much more demanding set of conditions.

While the specified accuracy of the BRC 8900A Peak Power Calibrator is ±0.6 db, when frequency correction is applied, it should be explained that utilization to a higher degree of accuracy, by virtue of its inherent stability, is both practical and recommended. The ±0.6 db figure is based on absolute worst case error without benefit of some error theories which propose a probable error as the RMS value of the worst case. The worst case error is also based on a minimum of standards equipment to perform a calibration. With high quality standards laboratory type equipment for calibration, operation to about a ±0.3 db worst case error is considered practical. Major potential sources of error are in the measurement of the attenuation path from the input connector to the CW standard output connector and error of the CW standard itself. Other worst case errors included in the 8900A analysis are:

1. Input VSWR reflection error.
2. Meter tracking and repeatibility error.

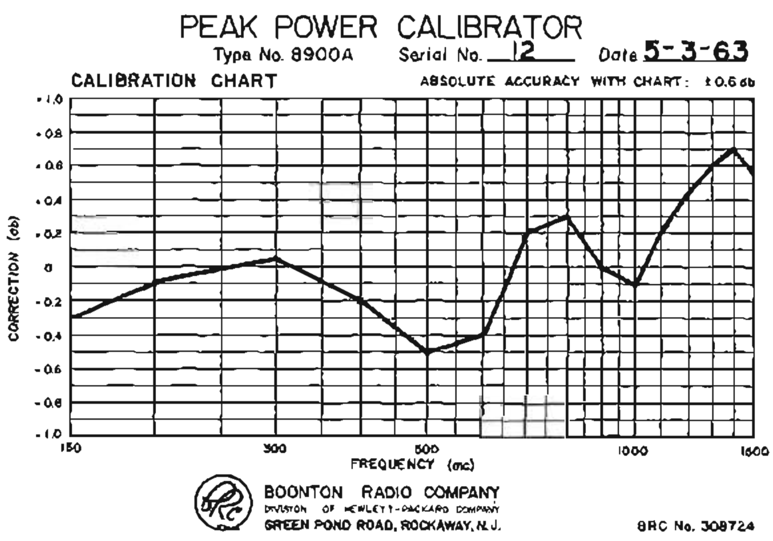

Figure 3. Typical Frequency Correction Curve - Type 8900A

3. Operator readout error.
4. Possible errors due to aging of the detector diode.

The aging error is included as a safety factor. Calibration checks of two prototype units, which had been in constant use for 6 months, exhibited less than .05 db discrepancy when rechecked at the end of that period. This amount of discrepancy could be attributed to any of the other errors first mentioned, but in a conservative evaluation, possible effects of aging must be considered. Provisions for recalibration permit the user to connect the standardizing equipment quite easily to external connectors on the instrument. Should field replacement of the detector diode become necessary, internal adjustments have been incorporated to permit adjustment of the front panel meter to accommodate its particular idiosyncrasies without necessity for an additional correction curve. A frequency correction curve is necessary when working to the highest degree of accuracy to remove errors due to the frequency sensitivity of the power divider diode detector, 10 db attenuator, and type N connectors. Without frequency correction, overall accuracy is conservatively rated at ±1.5 db over the specified frequency range. Figure 3 illustrates a typical individual frequency correction curve which can be supplied with the 8900A.

It is possible, then, to standardize the 8900A with a CW source, a CW power standard, and an oscilloscope. The procedure permits calibration at any frequency desired, and the user need not, as in some instruments, standardize at dc and wonder what the instrument is actually doing at a proposed measurement frequency.

A basic factor in the philosophy of the 8900A is that of actual observation of the pulse waveform during the measurement. While this requires the use of a suitable auxiliary oscilloscope, it was considered important in the reduction of subtle errors; some of which are variations in pulse width, rate, or shape, as previously mentioned. It has an inherent advantage, however, in permitting measurement of intermediate levels of power in a complex waveshape. The operator may ignore characteristics, such as overshoot, if they contribute nothing to the effectiveness of his system or he may measure them as he chooses. In some applications, the user may be monitoring the effectiveness of a system at the time the measurement is made in an effort to correlate system performance to peak RF power. It then becomes important that he verify the output power has not changed by even a few tenths of a db at the time of the reading. Some frequently useful methods of peak power measurement have the disadvantage that the operator stops looking at the waveform at the precise moment of measurement, which is the most important time of all. A typical oscilloscope display from the 8900A during a measurement is shown in Figure 4.

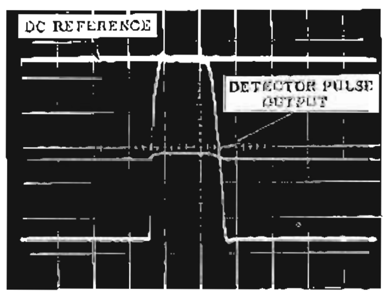

Figure 4. Typical Oscilloscope Display - Type 8900A

While the 8900A has a basic range of 200 mw full scale, it was designed with the intention that higher power measurements would be desired. Use of a high quality attenuator or directional coupler at the input will provide higher scale ranges. The directional coupler can provide an "in line" measurement with the source delivering useful power to an external load, thus providing a continuous monitoring capability. The accuracy of the external attenuating device is necessarily a factor in the overall accuracy of the measurement. If, for example, the 8900A were calibrated to an accuracy of ±0.4 db, and a range extending 40 db attenuator known to an accuracy of ±0.2 db were added to its input, accuracy of the overall measurement would be ±0.6 db. The accuracy of the measurement, reduced to its basic meaning, is really the degree of confidence that the operator has in the measurement and this confidence can come only from a thorough understanding of the theory and practical limitations of the measuring system.

CONCLUSION

In conclusion, the 8900A is BRC's approach to the general problem of accurate peak power measurement. Its basic features are high stability, ease of standardization, and elimination of some of the subtle errors of many present-day systems. Its characteristics have been conservatively rated and it is recommended that the user fully understand the theory of operation and knowledgeably apply it to measurement applications, especially where his requirements demand greater than specified accuracy. The complete specifications for the 8900A are given below.

SPECIFICATIONS

Radio Frequency Measurement Characteristics

RF RANGE: 150 to 1500 MC

RF POWER RANGE:
200 mw* peak full scale
*may be readily increased through use of external attenuators or directional couplers

RF POWER ACCURACY: ± 1.5 db*
*± 0.6 db with custom calibration curve

RF POWER PRECISION: ± 0.1 db

RF PULSE WIDTH: > 0.25 μsec

RF REPETITION RATE: 1.5 MC maximum

RF IMPEDANCE: 50 ohms

RF VSWR: < 1.25

Physical Characteristics

MOUNTING:
Cabinet for bench use; readily adaptable for 19" rack mounting

FINISH:
Gray engraved panel; green cabinet (Other finishes available on special order)

DIMENSIONS:
Height 6-1/8" Width 7-3/4"
Depth 11"

WEIGHT: Net: 10 lbs.

Power Requirements

8900A: 105-125/210-250 volts, 50-60 cps

Low Level Measurements Using the Signal Generator Power Amplifier

CHARLES W. QUINN, *Sales Engineer*

The normal applications of the 230A Power Amplifier (Fig. 1) have been discussed in Notebook Number 32.

Because of the choice of amplifier tubes, the noise figure of this new versatile amplifier is in the order of 6 to 8 db. Further, for most of the range, the noise figure is closer to 6 db. This feature opens another field of application—Low Level Measurements. It is the purpose of this article to discuss the many ways this power amplifier can be utilized in low level work.

TUNED MICROVOLTMETER

One of the most useful applications is the wedding of the 230A with an HP 411-A RF Millivoltmeter. The 411-A is connected as shown in Fig. 2, without a termination, but driving into the high impedance probe. Stub tuning at the output may improve the gain and VSWR at some frequencies. Under these conditions, the 230A will provide approximately 40 db of gain. The result is that full scale maximum sensitivity (which is normally 10 mv) is now approximately 100 μv, and 10 μv can be observed with ease. This configuration can be used to detect leakage and for harmonic analysis, using substitution to determine the gain at the frequency of operation. In this application it is possible to measure approximatey 80 db of insertion loss with 1 volt as a source voltage.

PREAMPLIFIER FOR FREQUENCY COUNTERS

When used in combination with the HP 524 and the 5243L and 5245L series counters with appropriate converters, the 230A Power Amplifier can make direct counter measurements possible with signal levels approximately 30 to 40 db below normal counter requirements.

Figure 1. Type 230A Signal Generator Power Amplifier

"OFF THE AIR" MEASUREMENTS

This application is especially useful when it is desirable to make transmitter measurements without interrupting transmission, when direct connection to transmission line is not desirable, or when the output of the transmitter is too low for normal measurements. In cases where direct connections are not made, an antenna may be substituted and remote measurements made up to several miles, depending upon the radiated power. (See Fig. 3.)

There are some precautions which should be taken in this application.

Figure 2. Tuned Microvoltmeter Setup with Stub Tuning

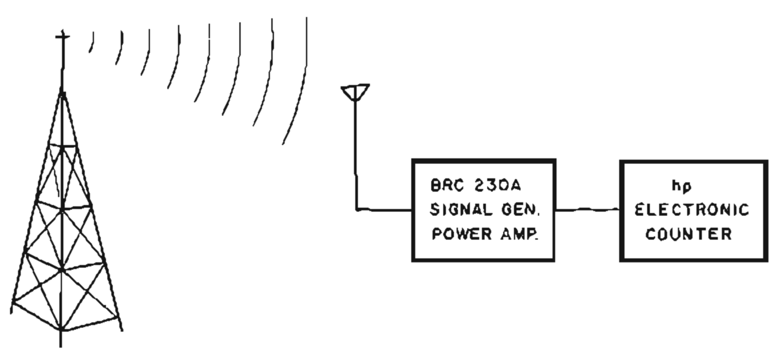

Figure 3. Setup for Remote Frequency Monitoring

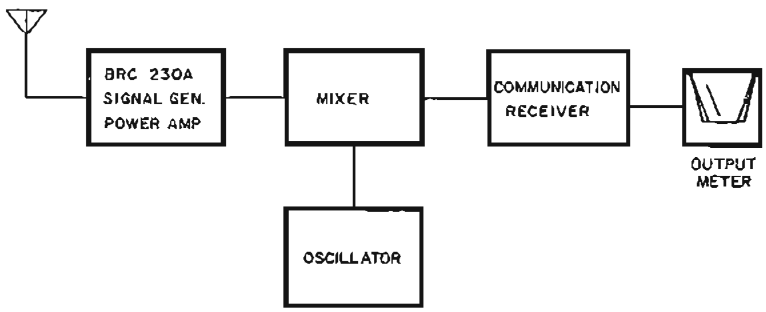

Figure 4. Setup for Leakage Detection

1. The effects of noise and modulation should be considered.
2. For AM signals, the negative modulation peak must not go below the triggering level.
3. For FM, the period must be sufficiently long for good averaging.
4. Even-order distortion, or carrier shift under modulation will be observed.
5. The absolute value of the peak noise voltage must be less than the triggering uncertainty, or hysteresis value.
6. The measurement must be made in the absence of interfering signals to the extent of the above noise limitation.
7. It is most desirable to check frequency at zero modulation.

RF LEAKAGE DETECTION

In the design of RF equipment, it is often necessary to detect very small signals, much less than 1 μv, usually picked up on a standard loop. When the frequencies involved get above 30 mc, it is quite common to use a good communication receiver as an IF amplifier and precede it with a broadband mixer. Normally, the insertion loss of the mixer will degrade the 10 db signal-to-noise ratio to approximately 10 to 20 μv. Adding the 230A Power Amplifier, as shown in Figure 4, will improve this figure to .2 to .5 μv for the same bandwidth of approximately 10 kc.

RECEIVER DESIGN

In the early stages of receiver design, the 230A Power Amplifier has numerous applications:

1. In the development of the IF amplifier stages, the 230A can serve as a temporary front end or RF preselector.
2. It can be used to provide high levels for limiters and detectors at IF frequencies above 10 mc.
3. It can be used to increase the output of a signal generator to determine proper mixing levels, thereby optimizing mixer gain and noise figure.

CONCLUSION

The 230A Power Amplifier is an extremely versatile instrument, capable of amplifying very small signals, as well as providing large signals, greater than 10 volts, for high level measurements with moderate input levels. With these features it becomes a valuable laboratory tool and also has many applications in the production line.

BRC CUSTOMER SERVICE DEPARTMENT

In order to offer improved service on repairs, accessories, and replacement parts, BRC has established a Customer Service Department which will operate as part of our Sales organization. The new department is staffed with personnel highy skilled in the repair and servicing of the complete line of BRC equipment.

Under the direction of Ray Tatman, the department is responsible not only for the scheduling and processing at the factory of repair instruments, but for handling all communications with our engineering representatives and customers regarding field repair and servicing of BRC instruments. Much of this communication will be handled by means of special "Service Notes" which will be issued periodically by Ray to keep BRC Engineering Representatives and customers posted with the latest servicing information.

Ray Tatman

Ray Tatman, head of the new operation, comes to BRC from the Hewlett-Packard Company in Palo Alto, California, where he spent two and one-half years in that Company's Customer Service Department. Ray, who hails from Oroville, California, attended Chico State College from 1949 to 1951 where he studied accounting and business administration. From 1951 to 1955 he served as an electronic technician in the U. S. Navy. In 1959 he received his degree in Electronics Engineering from California State Polytechnic College in San Luis Obispo, California.

The basic objective of the Customer Service Department is to provide the best possible service to our customers. If you have any questions regarding service, calibration, or repair, please do not hesitate to call or write us.

DON'T MODIFY YOUR RX METER YET

In a Service Note on Page 7 of Notebook Number 31, a new method was described for modifying the Type 250-A RX Meter which would provide for both a reduced signal level and increased sensitivity. It has since come to our attention that a number of instruments, modified in accordance with the instructions given, would no longer meet factory specifications. The calibrated oscillator frequency in these instruments changed beyond the specified ±1% limits and, in some cases, additional RF leakage occured which prevented bridge balance above 150 kc.

In view of these problems, we are recommending the discontinuance of this modification until such time as the problem is resolved. Work is being continued toward a solution to the problem and additional information will be published as soon as it becomes available.

ELECTRONICS, ELECTRONICS, EVERYWHERE

Over the years, BRC has received inquiries at the IRE show from a number of unusual organizations; sources which one might not readily associate with the electronic instrumentation field. This year, we did it again. A request for information about BRC equipment was received from the Dept. of Pediatrics, University of Washington, Seattle, Washington.

202H and 202J Instruction Manuals Now Available

The final instruction manuals for the Types 202H and 202J Signal Generators are now available. Included in the manuals are complete calibration and maintenance data, and parts lists. Copies of the manuals are being distributed through the BRC Engineering Representatives' offices. If you own a 202H or 202J and have not received your copy of the new instruction manual, contact our representative nearest you for your copy. Addresses and telephone numbers of our Representatives are given on Page 8 of this issue. Requests should include name of department and person to whom manual should be mailed.

MEET OUR REPRESENTATIVES

RMC Sales Division

One of the newest members of the BRC sales family is the RMC Sales Division of the Hewlett-Packard Company (formerly RMC Associates). RMC has two offices and handles BRC equipment sales in the Metropolitan New York City and Northern New Jersey areas. The Division's main office is located at 236 East 75th Street. A branch office is located at 391 Grand Avenue in Englewood, New Jersey.

RMC was founded in 1953 by Robert Asen, Milton Lichtenstein, and Charles Sargeant, all of whom were previously with Burlingame Associates as Field Engineers. Charles Sargeant has since retired. Robert Asen is Manager of the new Division and Milt Lichtenstein is Sales Manager. Rod Foley is Manager of the Englewood Branch.

Robert Asen, RMC Manager

The organization has an efficient and skilled technical staff consisting of eight Field Engineers, who are equipped to provide complete engineering service on our products; three Staff Engineers who are responsible for inside sales functions and provide "on-the-spot" technical assistance to customers; and a Customer Service Department which provides complete repair and recalibration services on all Hewlett-Packard instruments. An extensive stock of spare and replacement parts is also available to provide local service to customers.

We are proud to welcome RMC into the BRC family and invite all of our customers in the New York/Northern New Jersey area to contact them for complete information on BRC products.

Milton Lichtenstein, RMC Sales Manager

New York City Headquarters - RMC

WESCON 63!

FRONTIERS IN ELECTRONICS

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EDITOR'S NOTE
Q Meter Award

Seymour Krevsky, winner of the increasingly popular Q estimating contest, held each year at the IEEE show, was presented with a Type 160-A Q Meter by Bill Myers, General Manager of BRC. With near winning estimates in both 1957 and 1959, Mr. Krevsky is proving to be a perennial threat in the competition.

Mr. Krevsky received his B.S.E.E. from Newark College of Engineering in 1942 and his M.S.E.E. from the same college in 1950. He is currently studying for his doctorate at Polytechnic Institute of Brooklyn.

In 1944 Mr. Krevsky entered the service as a member of the technical staff of the Aircraft Radio Laboratory at Wright Field, Dayton, Ohio, where he performed studies on measurement techniques of parasitic FM in AM signal generators and in pulse modulated systems. Since that time he has done development and research work at the Coles Signal Laboratory and USARDL, and is presently engaged in advanced communications systems analysis and synthesis at the RCA Surface Communications Systems Laboratory.

Seymour Krevsky

Mr. Krevsky is a senior member of the IEEE, PGCS, PGAP, and PGMTT. He has had numerous articles published in IRE and SCEL pubications, and was written up in "Who's Who in Engineering" in 1958 and 1962.

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JULY, 1964

NUMBER 36

The NOTEBOOK

BOONTON RADIO COMPANY · ROCKAWAY, NEW JERSEY

A Division of Hewlett-Packard Company

A New System for Calibrating DME and ATC

JAMES E. WACHTER, *Senior Production Engineer*

INTRODUCTION

With the number of jet aircraft flights continually increasing, since their inauguration in 1960, it has become necessary to augment the existing VOR (Visual Omni-Range) and ILS (Instrument Landing System) navigational aids with two new sophisticated systems; DME (Distance Measuring Equipment) and ATC (Air Traffic Control). Boonton Radio, with an established line of specialized instrumentation for the design, test, and calibration of airborne VOR and ILS equipment, and recognizing the need for the same type of instrumentation in conjunction with the DME and ATC airborne equipment, has designed the 8925A DME/ATC Test Set (Figure 1) for this purpose. A block diagram of the Test Set is shown in Figure 2.

PURPOSE OF DME AND ATC

In brief, DME provides the pilot of an aircraft with a read-out of his distance in miles from a given ground station. In addition, DME equipment, used in pairs or in conjunction with VOR equipment, can give the pilot his exact location on a continuous basis, avoiding separate measurements for triangulation and/or calculation. ATC provides ground control personnel with positive individual identification and location of aircraft within their area.

YOU WILL FIND . . .

Figure 1. Type 8925A DME/ATC Test Set

Both the DME and ATC systems function through the exchange of pulse coded information between the airborne and ground stations. The pulse coding, plus the time delays associated with transmission and reception, constitutes the information. In the case of DME, the airborne equipment interrogates the ground station which, in turn, replies. In the ATC system, the ground station interrogates the airborne equipment.

TYPE 8925A DME/ATC TEST SET

The basic concept of the 8925A Test Set was that it should be, in essence, a calibrated precisely controllable, low power, ground station. The minimum characteristics of the station were to be those as specified by the cognizant authorities for a DME and ATC ground station and for checking out airborne DME/ATC equipment.[1]

Analysis of the requirements led to the conclusion that existing "tried and proven" test instruments were available which, when assembled in building block fashion, could provide the basis of the calibrated pulsed RF source and a means of measuring the airborne transmitter peak power. With modification of some of these units, the exact requirement could be met. Having established this course, there remained to devise means of measuring the ATC airborne transmitter frequency, interconnecting the Test Set components, connecting to the equipment under test, and monitoring the various signals involved. This was accomplished by the development of two new specialized test instruments: the BRC 8905A Wavemeter to measure the transmitter frequency and the BRC 13505A Isolator-Monitor to provide the required interconnection, isolation, and monitoring facilities.

Signal Generator

The basic CW RF signal is generated by a Hewlett-Packard H01-8614A Signal Generator. This instrument is a slightly modified version of the standard production unit. The frequency range is restricted to 950 to 1250 mc in order to optimize the characteristics over this range, and the attenuator calibration is offset to compensate for system losses. The modifications are minor and the instrument can be readily returned to its original state. The generator incorporates automatic leveling of the RF signal which permits tuning over the entire frequency range with no adjustment of level required. The attenuator dial is calibrated to read di-

1 - See References at end of article.

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ulation is essentially independent of frequency and, in actual use, when set at one frequency, will hold across the entire frequency range of 950 to 1250 mc.

The video signals required to actuate the HO3-8714A Modulator are derived from equipment external to the 8925A Test Set, such as the Collins 578D-1 Test Set (DME) or 578X-1 Test Set (ATC). These units

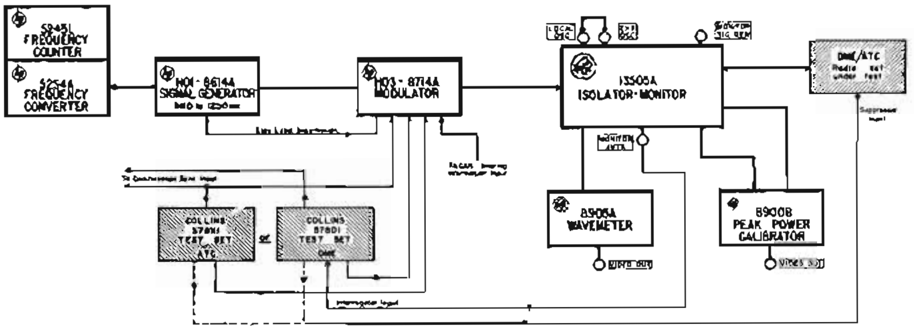


Figure 2. Block Diagram — Type 8925A DME/ATC Test Set

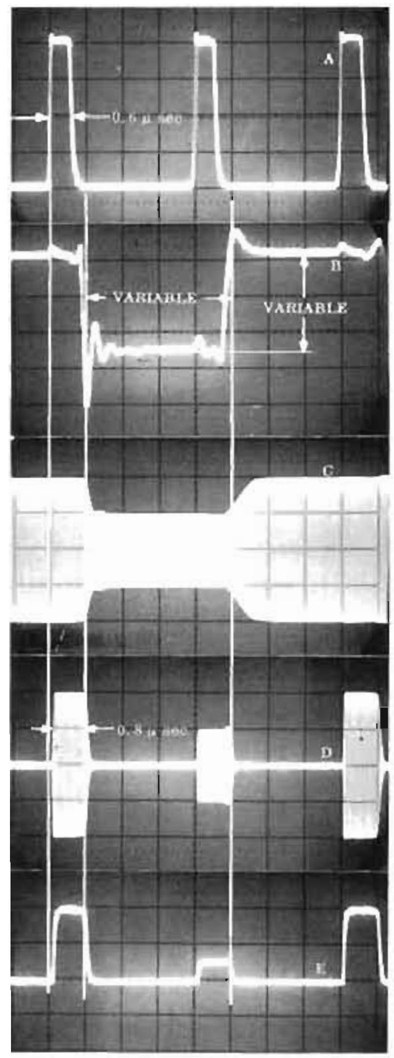
rectly the system output into 50 ohms over the range of −10 to −120 dbm. The generator contains an internal modulator which employs PIN diodes as essentially resistive modulator elements. The frequency controls permit adjustment to within 50 kc as the frequency is simultaneously monitored and displayed by an -hp 5245L Frequency Counter in combination with a 5254A (3.0 Gc) Frequency Converter.

Modulator

The RF output of the Signal Generator is fed to an -hp- HO3-8714A solid state Modulator. Like the modulator in the Signal Generator, this instrument employs electrically controlled PIN diodes mounted in a strip transmission line. Modulation is accomplished by varying the attenuation of the strip line. The HO3-8714A is an extensively modified version of the -hp- 8714A. Special front panel controls and connectors are provided and special pulse shaping circuitry is included. The pulse shaping circuitry is necessary to compensate for the inherently fast PIN diode switching time (of the order of 10 ns), permitting the RF envelope to closely reproduce the video modulating signals. The linearity of the modulator is approximately ±5% over the upper 30 db of the dynamic range. This method of post generation modulation

provide pulse coded video signals which simulate DME/ATC ground emission information as regards to pulse spacing, repetition rate, number of pulses, and, to some extent, pulse shape.

In the case of ATC, it is required that the second pulse of a train of two or three pulses be varied in amplitude from +1 to −10 db relative to the amplitude of the first pulse. The special circuitry within the HO3-8714A Modulator provides this capability by generating a video gate (Figure 3B), which is related in time to the second pulse of the video pulse train and adjustable in amplitude by means of a front panel control. This video gate pulse is returned to the modulator in the HO1-8614A Signal Generator where it effectively modulates the RF signal upward or downward the desired amount (Figure 3C). The generator output, then, is modulated by the HO3-8714A which turns completely on for each pulse (Figure 3D). The resulting pulse amplitudes are determined by the previously modulated generator output. The linearity of the HO3-8714A internal circuitry is such that the front panel control is calibrated in 0.5 db increments over the range of +1 to −10 db and is accurate to ±0.5 db. It is apparent that in order to permit the upward modulation, it is necessary to bias the modulator within the generator to some level below the full on condition. The bias which is derived from the HO3-8714A Modulator, can be made to vary the RF output over a range of several db by means of a control available at the front panel. In practice, the level which is set corresponds to the Test Set calibrated CW output.



(A) VIDEO SIGNAL APPLIED TO HO3-8714A MODULATOR

(B) VIDEO GATE GENERATED BY HO3-8714A MODULATOR AND APPLIED TO HO1-8614A SIGNAL GENERATOR

(C) HO1-8614A SIGNAL GENERATOR RF OUTPUT (VIEWED USING 13505A ISOLATOR-MONITOR HETERODYNE MONITOR)

(D) TEST SET RF OUTPUT (VIEWED USING 13505A HETERODYNE MONITOR)

(E) TEST SET RF OUTPUT (VIEWED USING 13505A DIODE MONITOR)

Figure 3. Generation of ATC Ground Station Signal (Time Axis 1 Microsecond per Centimeter)

While the DME system does not require any differences of relative pulse amplitudes, a closely allied system, TACAN (Tactical Air Navigation) does. The 8925A Test Set has provided means to simulate both. See Figs. 4 and 5. TACAN utilizes the basic DME pulse coded signal but, in addition, requires up to 55% amplitude modulation of the pulse train with a composite signal comprised of 15 cps and 135 cps sine waves. This AM provides bearing information.

While the 8925A Test Set is capable of providing full TACAN RF signals, there is no currently available single source of synchronized pulse and sine wave signals. However, the video TACAN signal may be simulated by using an audio oscillator (such as the -hp- 200CD) in conjunction with the Collins 578D-1. The upward modulation requires that the HO3-8714A Modulator be biased to a level, in the absence of AM, some 4 db below the full on condition. The bias is derived internal to the Modulator and, as in the ATC case, a control available at the front panel permits variation of the RF output over a range of several db. In practice, this is set to a level corresponding to the Test Set calibrated CW output.

Another control, available at the front panel of the HO3-8714A Modulator, permits variation of the Test Set CW output signal over a range of several db. This allows ease of recalibration after extended periods of use.

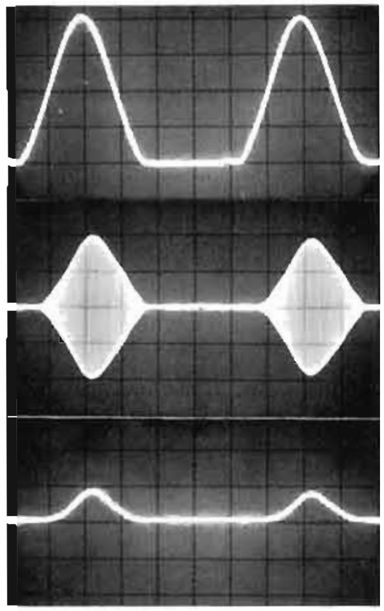
Isolator-Monitor

The simulated ground station RF signal is fully generated at this point. The BRC 13505A Isolator-Monitor connects the RF signal to the airborne instrument under test and monitors the signal for measurement purposes. In the normal mode of operation, the incoming RF signal is passed through a low attenuation path (approximately 1 db) of a properly terminated four-port circulator, through a coaxial switch electrically actuated from the front panel, to the output connector. It is the signal level at this connector for which the HO1-8614A Signal Generator attenuator is calibrated. A cable from this point to the antenna jack of the receiver/transmitter under test completes the path. The transmitter signal, gener-

(A)

(B)

(C)



(A) VIDEO SIGNAL APPLIED TO HO3-8714A MODULATOR

(B) TEST SET RF OUTPUT (VIEWED USING 13505A HETERODYNE MONITOR)

(C) TEST SET RF OUTPUT (VIEWED USING 13505A DIODE MONITOR)

Figure 4. Generation of DME Ground Station Signal (Time Axis 2 Microseconds per Centimeter)

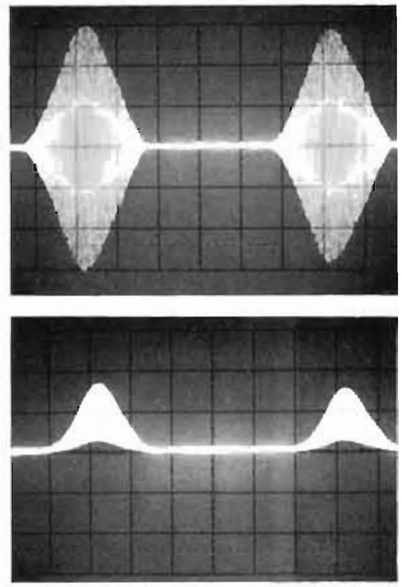
ally of high peak power (up to 2000 watts), returns by way of the same route, but is isolated from the equipment preceding the 13505A by the 30 db minimum insertion loss of the circulator in this direction. Some of the transmitter signal is available at the remaining two ports of the circulator, and, after suitable attenuation, is brought out at the 13505A front panel for frequency and power measurements.

When the 13505A is used as a monitor, the test signal is diverted from the output connector by means of the coaxial switch and routed to the monitoring circuits. Because the load is removed from the antenna jack of the transmitter under this condition of operation, a pair of interlock terminals are provided on the 13505A rear panel which may be used to automatically de-energize the transmitter, if so desired. Two monitoring modes are available by operation of the front panel switch: a linear heterodyne mixer and a diode detector.

When the 13505A is operated as a heterodyne monitor, the signal from an internal oscillator operating at 1025 mc is mixed with the test signal

(A)

(B)



(A) VIEWED USING 13505A HETERODYNE MONITOR

(B) VIEWED USING 13505A DIODE MONITOR

Figure 5. Simulated TACAN Ground Station Signal (Approx. 45% AM — Time Axis 2 Microseconds per Centimeter)

and the difference frequency (IF), either plus or minus, is delivered to an amplifier. The combined circuitry has a bandwidth, at the 3 db points, of approximately 10 mc. The 1025 mc frequency was chosen because it lies between the frequencies of the middle DME ground station channels (channels 61, 62, and 63) and the ATC beacon frequency of 1030 mc. Thus, the difference frequency in either case will fall within the pass band of the amplifier. Best results are obtained, naturally, when the IF frequency is centered in the amplifier pass band. Actually, a region of about 4 mc in range gives equally good results due to the presence in the mixer output of vestigal sidebands which extend far beyond the amplifier frequency limits. This being the case, if the IF frequency is not centered within the amplifier pass band, the sidebands which are not passed on one side of the IF frequency are, in part, made up for by the additional sidebands passed on the other side of the IF frequency. This, in effect, increases the apparent bandwidth of the mixer-amplifier combination. The output of the heterodyne monitor is available at the 13505A front panel for viewing on a suitable oscilloscope (such as the -hp- 175A). A minimum of 1v peak to peak is obtained when the Test Set signal is

−10 dbm, and the presentation is linear within ±0.5 db over the range of test signal levels of +4 to −10 dbm. This provides an accurate means of relative amplitude measurements such as are required for setting the ATC side lobe suppression pulse and the percent amplitude modulation of the TACAN signal.

The bandwidth of the heterodyne monitor is more than adequate for making pulse shape measurements of the DME (or TACAN) pulses, which have rise and fall times of 2.5 μs nominally. However, in the case of ATC pulses, which may have rise and fall times of 50 ns, an error of up to +40% could result. For accurate measurement of the ATC pulse shape, it is recommended that a sampling type oscilloscope (such as the -hp- 185B) be used to directly observe the 8925A DME/ATC Test Set output signal.

It should be mentioned, that, due to the frequency independence of the PIN modulator units, pulse settings made using the heterodyne monitor (that is with the HO1-8614A Signal Generator tuned to the restricted band of frequencies which provide a heterodyne signal), will remain constant at any other frequency to which the generator is set. Nevertheless, should it be desired to employ the heterodyne principle at any other frequency, provision is made on the back panel of the 13505A Isolator-Monitor for the substitution of an external oscillator (capable of 60 mw output power) for the internal oscillator. Also, the internal oscillator is capable of being tuned over a small range of approximately ±5mc by a control available at the 13505A back panel.

When operating the 13505A as a diode monitor, the internal oscillator is turned off and the mixer then becomes a diode detector with an output amplified by the same amplifier discussed previously. A minimum of +0.1v peak is obtained at the 13505A front panel for a Test Set signal of −10 dbm at any frequency within the range of the Test Set. As with most diode detectors, the output is not a linear presentation and care must be exercised when using it for making measurements. (See Fig. 6). The relationship between the diode monitor output and the Test Set RF signal may be readily examined by varying the Signal Generator attenuator in fixed amounts and noting the corresponding changes in the diode monitor output amplitude. While it is not as accurate as the heterodyne monitor, the diode monitor does provide a rapid means of making relative measurements at any frequency within the range of the Test Set. This signal is also used when calibrating the overall DME system (including the 578D-1 video modulator) delay.

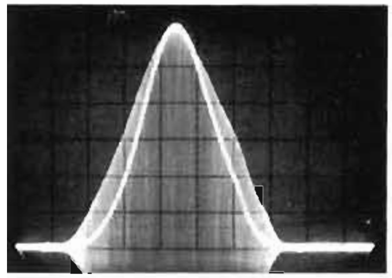


Figure 6. Comparison of Diode Monitor and Heterodyne Monitor Displays of the Same RF Signal (Time Axis 1 Microsecond per Centimeter)

With the 13505A operating in the normal mode, portions of the transmitter power are available at the front panel Wavemeter and Pwr Meter connectors. The amounts of power available are controlled by the inclusion of fixed precision attenuators within the 13505A.

Peak Power Calibrator

A BRC 8900B Peak Power Calibrator is connected to the Pwr Meter output of the 13505A. This is a production unit specially calibrated to account for variations in attenuation in the system (including cables) preceding it. The unit will read peak power of 100 to 2000 watts over the frequency range of 960 to 1215 mc to an accuracy of ±1.2 db. The accuracy of measurement may be improved to ±0.6 db with special calibration. Should measurement of other power ranges be desired, it may be accomplished by changing the fixed attenuation within the 13505A. This would be handled on a special order basis. For peak powers in excess of 2000 watts, the user may, with some degradation of accuracy, insert additional attenuation between the 13505A and 8900B, external to the system.

In making a power measurement, the 8900B basically compares the demodulated pulse envelope of the signal to be measured and the output from an internal dc reference supply. A mechanical chopper permits both signals to be viewed simultaneously using a suitable oscilloscope (-hp-175A). The dc reference is adjusted, by means of a front panel control, to be equal in amplitude to the demodulated pulse. The resulting dc reference is indicated on the front panel meter, which is calibrated to read peak RF power. Provision is made for readily recalibrating the instrument against an external bolometer or calorimeter.

The 8900B also provides a means for monitoring the transmitter output pulses. The demodulated pulse is brought to a back panel connector through a two-stage emitter follower. By means of intercabling, this signal is available at the Xmtr. Monitor connector on the front panel of the 13505A for monitoring purposes, and, in the case of DME, as the demodulated transmitter interrogation signal required by the 578D-1 for distance measurement.

Wavemeter

The BRC 8905A Wavemeter is connected to the Wavemeter connector on the front panel of the 13505A and is used to measure the frequency of the pulsed output signal of the ATC transmitter. The instrument is composed of a transmission type tuneable wavemeter and associated metering circuitry. The front panel meter indicates a peak reading when the cavity is tuned to the incoming frequency. This frequency can be read to within ±0.5 mc directly from a dial calibrated in 0.5 mc increments from 1070 to 1110 mc. The sensitivity of the unit is adjusted for individual Test Set losses, so that meter indications are obtained for any ATC transmitter signal within the specified frequency range and within the peak power limits of 250 to 1000 watts. A video output is provided whereby the detected cavity output may be monitored with a suitable oscilloscope for frequency measurements of signals having peak power as low as 10 watts.

Interconnecting Cables

Because the calibration of the 8925A DME/ATC Test Set takes into account all the known system losses and the many interfaces which exist, the over-all system accuracy is dependent upon all critical interconnecting cables remaining in their proper locations. To insure this, all cables and their associated connectors are color coded.

CONCLUSION

The 8925A is the most complete and universal Test Set available for the checking of DME and ATC airborne equipment and, because of its building block construction, is readily adaptable to many special applications. Its high degree of stability and continuous frequency tuning are highly desirable features when regarded in the light of possible future expansion of the DME system; i.e., channel splitting.

The Test Set is the result of a Hewlett-Packard corporate effort and the responsible groups are deserving of recognition. Individual instruments which form part of the Test Set and the divisions responsible for their development are listed below.

Instrument		*-hp- Division*
5245L	Electronic Counter	Frequency and Time Division
5254A	Frequency Converter	Frequency and Time Division
HO1-8614A	Signal Generator	Microwave Division
HO3-8714A	Modulator	Microwave Division
13505A	Isolator-Monitor	Boonton Radio Division
8900B	Peak Power Calibrator	Boonton Radio Division
8905A	Wavemeter	Boonton Radio Division

REFERENCES

1 - *Air Traffic Control Transponder, ARINC Characteristic No. 532C, Mar. 1, 1961.*

2 - *Airborne Distance Measuring Equipment (DME) ARINC Characteristic No. 521D, Nov. 1, 1963.*

3 - *Minimum Performance Standards Airborne Distance Measuring Equipment (DMET) Operating within the Radio-Frequency Range of 960-1215 Megacycles, RTCA Paper 167-59/DO-99, Sept. 8, 1959.*

4 - *Minimum Performance Standards Airborne ATC Transponder Equipment, RTCA Paper 181-61/DO-112, Dec. 14, 1961.*

X-Y Plotting with the Types 202H and J FM-AM Signal Generators

CHARLES W. QUINN, *Applications Engineer*

INTRODUCTION

A rear panel jack on the Types 202H and 202J FM-AM Signal Generators provides a means for introducing an external dc voltage to control the output frequency over a limited range. The jack permits direct coupling, at low rates, to the reactance tube modulator in the generator. This feature, together with the improved linearity, automatic leveling, and low FM noise characteristics of these generators, opens the door to applications new to the FM signal generator art.

Two major areas of application which make use of the direct-couple feature of the 202H and J signal generators will be discussed in this article. One area deals with tests and measurements that can be made with the signal generator connected with an X-Y Plotter. The other area of application involves tests and measurements that can be made with the signal generator connected in an automatic frequency control setup or system.

X-Y PLOTTER APPLICATION

The aforementioned features of the 202H and J signal generators, together with the advantages of the X-Y Plotter; i.e., large display, extremely good linearity, permancy, and reproducibility, make these instruments an excellent combination. In this application, sweep widths from a few kc to 1 mc are possible. This pair of instruments, plus a 207H Univerter and a 230A Power Amplifier used as a doubler, yields a potential frequency range of 100 kc to 500 mc for the applications to be discussed in this article.

The auxiliary equipment required depends on the specific test to be made. This equipment will be listed as each application is discussed.

The major areas of X-Y Plotter application are receiver testing and narrow-band filter testing.

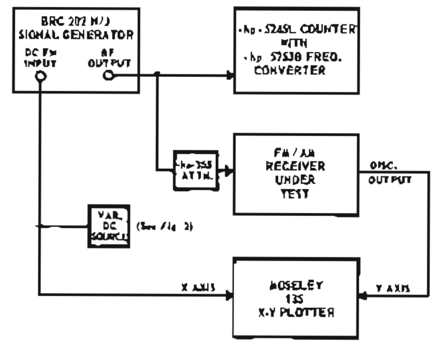


Figure 1. Setup for Checking Discriminator Output vs. Frequency

Receiver Testing

Receiver tests that can be made to advantage with the 202H/J and X-Y Plotter are as follows:

1. Selectivity or bandwidth versus frequency and input.
2. Discriminator output versus frequency.
3. AGC or detector voltage versus input level or overload characteristics.
4. Audio/video output versus input level.
5. Audio/video fidelity or response versus level.

The first tests to be discussed will be those tests which use a minimum of equipment; i.e., selectivity or bandwidth versus frequency and RF input level, and discriminator output versus frequency. If it is assumed that the receiver to be tested is in the range of the 202H and J, the connections and equipment are shown in Fig. 1. The Plotter may be a Moseley Type 2D, 135, etc. The Y axis input to the Plotter is connected to the AM detector, limiter grid, or discriminator output. The X axis input is connected to the variable dc source. The simplest form of variable dc is a multi-turn variable potentiometer of approximately 10,000 ohms and two 6-volt batteries connected as shown in Fig. 2. This supply will produce approximately 600 kc sweep on the

202H low band, and about 1200 kc on the 202H high band and the 202J. Reference to the dc FM input curve (Fig. 3) will indicate how to optimize this sweep if necessary. (A function generator, such as the -hp- 202A, could be used in place of the dc source described above.) With the equipment connected as shown in Fig. 1, typical curves, such as those shown in Figs 4 and 5 can be plotted.

Calibration of Scales

Before the data in the plotted curves can be of value, the X and Y axes must be calibrated. The vertical scale can usually be read directly, using the indicated sensitivity on the Y amplifier controls. If more precise calibration is necessary, an -hp- 412A DC Voltmeter may be used in shunt with the Y terminals.

The horizontal X axis may be calibrated in a number of ways, depending upon the accuracy required and the equipment available. The curve in Fig. 3 could be plotted as a function of the X input sensitivity, as indicated by the calibration on the X amplifier controls. A much more precise method, however, utilizes a crystal calibrator or an -hp- 5243L or 5245L counter with an -hp- 5253B Converter plug-in. Fig. 1 shows the setup using the counter. Note that additional attenuators are necessary if continuous frequency monitoring is desirable. This enables the counter to operate at a reasonable level with the 202H and J. The Type 202JA has 50 mv available for continuous monitoring.

It is good practice to calibrate the frequency or X axis beginning at the most important point, usually the center, peak, or zero crossing, depending upon the curve being plotted. There are a number of choices in the method of marking. Two possibilities are given below.

1. Retrace the curve with "pen up" and mark a vertical line at the desired frequency increments, using the Y axis "Zero" control and the "pen up/down" control. This procedure produces a trace marked as shown in Fig. 4.

2. After checking that the reference point is at the desired location, reduce the Y signal to zero. Sweep the signal generator through the same limits and mark the desired increments with a vertical line, using the Y axis "Zero" control. This procedure produces a trace calibration as shown in Fig. 4.

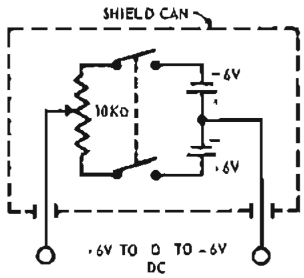


Figure 2. Variable DC Source Using Multi-turn Variable Potentiometer

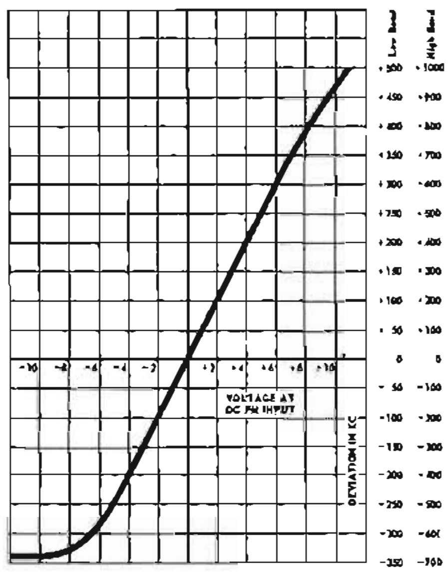


Figure 3. DC FM Input Curve

AUTOMATIC FREQUENCY CONTROL APPLICATION

Since it is possible to feed a dc signal into the reactance tube modulator directly through the DC FM INPUT jack, this dc signal might easily be an error signal which is a function of some reference frequency. This arrangement, shown in block diagram form in Fig. 6, is helpful in reducing the frequency drift per hour of the 202H and J signal generators by a significant factor. Frequency stability of 0.001% per hour is easily obtainable with this setup. This improved stability is valuable in the testing of narrow band (about 10 kc) systems where drift is a problem. System frequency can also be measured more precisely on a continuous basis with this method, utilizing a frequency counter connected as shown in Fig. 6. At this point, it is well to point out that the AFC correction is not instantaneous. In fact, there is a time constant of approximately one-half second, so that short-term frequency changes, such as FM modulation rates, are not cancelled out by the AFC loop. This permits all of the standard tests to be performed while the long-term drift is corrected.

Phase Locking

A slight modification in the area of the DC FM INPUT circuit will per-

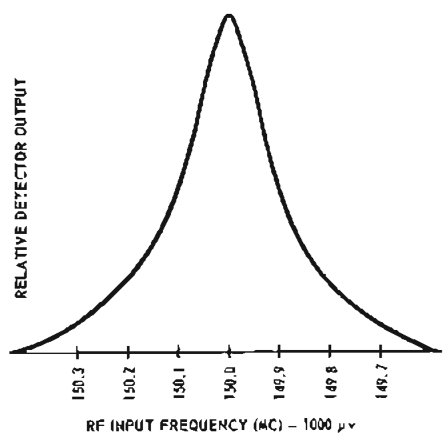


Figure 5. Overall Receiver Selectivity Curve

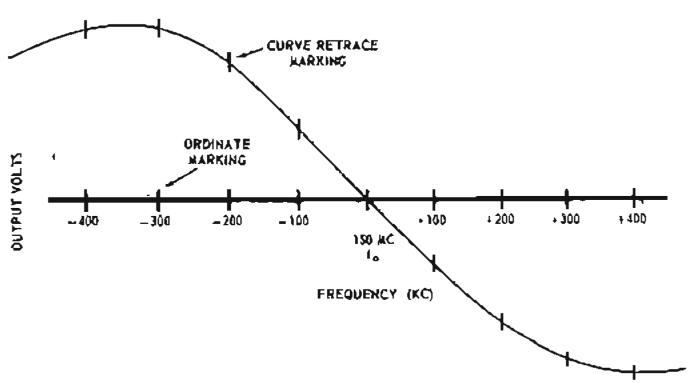


Figure 4. Discriminator Output vs. Frequency Curve

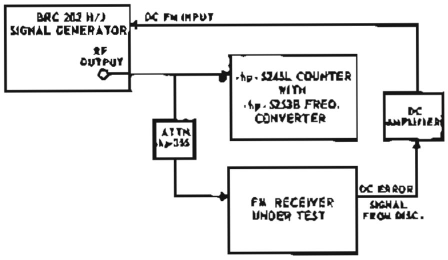


Figure 6.
Setup for Automatic Frequency Control

mit phase locking, to a suitable reference, of the 202H and J signal generators. This could considerably reduce phase noise in critical applications, but the FM function would be cancelled out by the phase lock loop. Frequency stability in this case would be that of the reference.

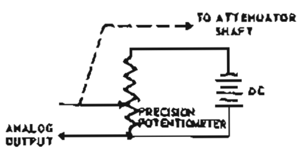


Figure 7.
Connections for Analog Output vs. Attenuator

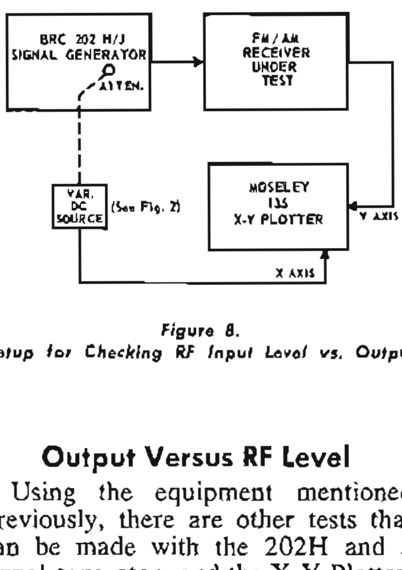


Figure 8.
Setup for Checking RF Input Level vs. Output

Output Versus RF Level

Using the equipment mentioned previously, there are other tests that can be made with the 202H and J signal generators and the X-Y Plotter. It is a simple matter to connect a double-ended precision potentiometer directly to the attenuator shaft by removing the knob. Using the circuit shown in Fig. 7, a voltage analogous to RF input or attenuation can be obtained. The equipment is connected as shown in Fig. 8. Typical curves are shown in Fig. 9. When the Y axis is a function of audio/video voltage, the "AC Voltage" position in the Type

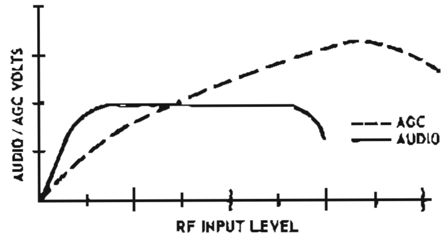


Figure 9.
Typical Output vs. RF Input Level Curves

2D Plotter may be used; otherwise, an ac to dc converter or detector must be used to obtain a dc signal proportional to the audio/video output.

Audio/Video Fidelity or Response

If an audio/video oscillator with a dc analog output is available, such as the DYMEC Model DY-207A, response curves may be plotted directly.

EASTERN SERVICE CENTER TO BE ESTABLISHED AT BRC

In order to provide improved parts and repair services for customers in the Eastern U.S.,. the Hewlett-Packard Company is establishing a new Eastern Regional Service Center at Boonton Radio Company in Rockaway, N. J. The new facility will provide complete parts support and an extensive factory-level instrument repair service. Parts and factory repair service will be available for all Hewlett-Packard Company equipment and for equipment manufactured by -hp- Divisions; including BRC, Dymec, Moseley, and Sanborn.

The new operation will be set up in the present BRC plant and will occupy about 25,000 square feet of space: 10,000 square feet for parts warehousing and repair areas and 1,500 square feet for administrative functions. Approximately 90 employees will be hired to handle the various administrative and technical tasks.

Manager of the new service center, under the direction of Bill Myers, BRC General Manager, will be Al Thoburn, formerly manager of the materials handling group in the Western Service Center. Service Manager will be Bob Wolfe, who served as Service Manager for RMC Sales Division in New York City. Dick Love,

Al Thoburn

Bob Wolfe

Dick Love

formerly parts manager at the Western Service Center, will be Parts Manager.

It is expected that the new service center, scheduled to begin operations on August 17, will provide even better service to all our customers in the Eastern region of the country.

ERWIN CONRAD TO BE SERVICE ENGINEER

Erwin Conrad joined BRC in January of this year as Customer Service Supervisor, replacing Ray Tatman who has been on special assignment pending his return to Corporation headquarters in Palo Alto. Erwin came to BRC from the RMC Sales Division in New York City, where he was Assistant Service Manager and had gained seven years of valuable experience in the service of Hewlett-Packard products. In his present assignment as Customer Service Supervisor, he has been responsible for all BRC factory repairs, replacement parts, and general service support.

With the establishment, in August, of the new Eastern Regional Service Center, announced in this issue, Erwin's present operation will be integrated into the service center and he will assume new duties as Service Engineer for all Boonton Radio products. In this capacity, he will be responsible for aiding in the preparation of field maintenance procedures and providing the necessary customer support and training for these instruments. His duties will also include the preparation of Service Notes and other service publications, to better enable -hp- field repair stations and customers to properly maintain BRC products.

Q CONTEST WINNER

The IEEE Show at the Coliseum in New York in March found BRC's booth bustling with contest hopefuls all determined to "out-estimate" each other in the annual BRC "Guess the Q" contest. This year's display coil was formed with heavy copper wire into a configuration not too unlike the BRC logograph. It proved challenging to the parade of booth visitors.

Bill Myers, BRC General Manager, Presents Q Meter to Contest Winner George Engert

The Q of the coil, it turns out, is 242.6. This is the "indicated Q" and represents the average of ten measurements made at 10 mc on the Type 260A Q Meter in the BRC Standards Laboratory.

Winner of the contest, with an estimate of 242.5, is George W. Engert. Project Engineer with MEPCO, Inc. in Livingston, N. J. Runners-up are Alan Budner of U. S. A. Electronic R & D Lab. (240); John Mulqueen of MEPCO, Inc. (246.5); and T. A. Metz of Polyphase Inst. Comp. (248).

Mr. Engert was awarded a factory reconditioned Type 170A Q Meter at BRC on June 17.

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A Compact, Versatile 10 to 500 MC Oscillator

CHANNING S. WILLIAMS, *Electrical Engineer*

The 3200A VHF Oscillator (Figure 1) is a compact, versatile general purpose instrument intended for use in receiver and amplifier testing, driving bridges, slotted lines, antennas and filter networks, and as a local oscillator for heterodyne detector systems in the frequency range from 10 to 500 mc. Completely self-contained, the instrument is packaged in the new Hewlett-Packard modular cabinet, permitting convenient bench use as well as rack mounting for system applications. The oscillator is housed in a rugged aluminum casting for maximum stability and extremely low leakage. Six frequency ranges are provided for adequate band spread on a slide rule dial.

Figure 1. 3200 VHF Oscillator

CIRCUIT DESIGN

The oscillator circuit, shown in Figure 2, employs push-pull 6DZ4 tubes with capacitive tuning and a turret system which permits switching of the tank circuit inductance on the various frequency ranges. Feedback is accomplished with a capacitive divider from one plate to the opposite grid, using the grid-to-cathode capacitance of the tube, together with a fixed mounted capacitor from the other plate. This two-tube oscillator is particularly well suited to this design because it provides more power than a single 6DZ4 tube and feedback is obtained by fixed capacitance on the top four bands. On the two lower bands, drive is reduced by switching in additional capacitance from the grids to ground. The two-tube oscillator also works well with a split-stator capacitor which requires no wiping contacts, eliminating a potential source of noise and instability.

In this oscillator, the center of the tank is at ground potential and therefore the rotor of the capacitor is also at ground potential for RF frequencies. Since the center of the oscillator coil is also roughly at the neutral or ground plane, plate power can be injected from this point from a common supply ring on the turret. This ring is a slip ring rather than a switchable contact. Actually, the oscillator turret is so constructed that the center of each coil is permanendy tied back to this common slip ring to individual resistors. These resistors serve to break up undesirable RF paths, without introducing appreciable plate voltage or RF loss.

FREQUENCY STABILITY

The oscillator is specified for a ±0.002% stability after a 5-minute warmup. However, *typical* data shown in Figure 3 indicates that, under controlled conditions, 5-minute stabilities of 0.0001%, or 1 part per million, have been measured at some frequencies.

Frequency stability has been achieved through careful design of the circuit components and the use of a substantial aluminum casting which provides a large thermal mass. Particular care has also been given to the mechanical design of the turret assembly. The detent mechanism is positive, assuring accurate and stable positioning of the active oscillator inductor. The turret itself is precision molded from orlon filled diallyl phthalate. Turret contacts are constructed of coin silver and tuning ca-

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pacitor contacts are of coin silver and beryllium copper laminate. Both the detent spring and tuning capacitor contacts have been subjected to long life tests without failure. The circuit is mounted on a silverplated brass chassis in such a way that lead lengths have been minimized and fundamental circuit symmetry has been maintained.

The position of the tuning capacitor is indicated on a large slide rule dial which is simultaneously rotated by the oscillator range switch mechanism to display only the active frequency range. The frequency drive mechanism is backlash-free and employs a cable drive. The cable consists of a glass core, for dimensional stability, enclosed in braided Nylon for long wear and traction. Nonconductive cable is used so that entrance into the RF enclosure can be made through waveguide-below-cutoff tubes, permitting low RF leakage without wiping grounds. The turret drive shaft also utilizes nonconductive Fiberglass Epoxy which passes through a waveguide-below-cutoff tube into the RF enclosure.

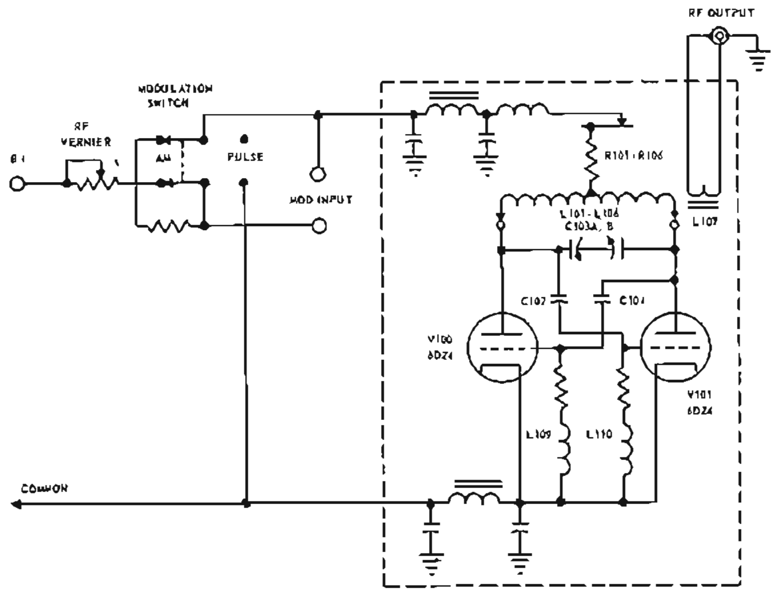


Figure 2. Oscillator Circuit

MODULATION

The simplified modulation circuit is shown in Figure 4. AM plate modulation is injected through front panel terminals from an external source of audio power. The modulation signal is impressed across a resistor in series with the plate supply to the oscillator. The oscillator is specified at less than 1% AM distortion at a level of 30% AM. However, modulation up to 50% AM can be obtained with a resulting increase in distortion. Approximately 30 volts RMS into 600 ohms is required from an external source to achieve 30% AM. 60-cycle modulation can be conveniently obtained at almost any amplitude desired. If modulation percentages in excess of 40% are required, they may alternately be obtained by switching to pulse operation and applying a transformer-coupled audio signal in series with an adjustable dc (Figure 5); the resultant RF level, however, may be less than the specified CW signal. For 100% modulation, a dc offset voltage equal to the peak voltage of the audio signal is applied at the modulation terminals. The maximum voltage (dc offset voltage plus RMS audio voltage) should not exceed 135 volts. On the range from 260 mc to 500 mc, it will not be possible to modulate linearly 100% since the oscillator will not start before the plate voltage reaches some appreciable value.

Because of the direct plate modulation, some small amount of frequency modulation will occur. Typical FM deviation for a fixed audio input level is shown graphically in Figure 6.

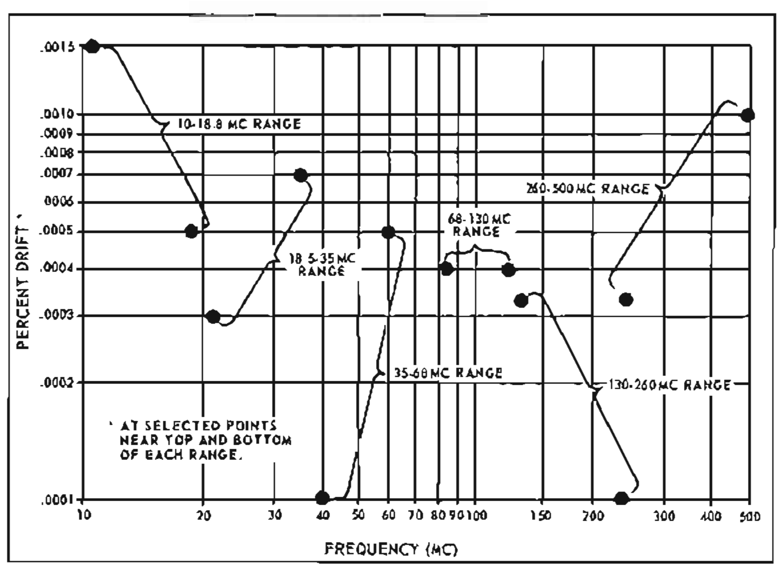


Figure 3. Typical Frequency Drift — For 5 Minutes After 4-hour Wormup

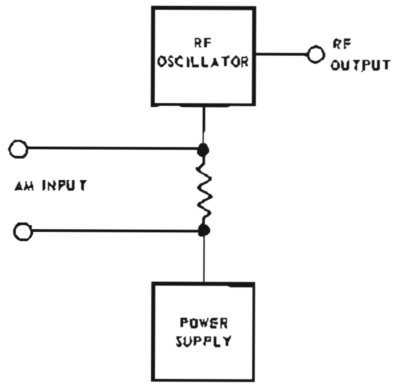


Figure 4. Amplitude Modulation

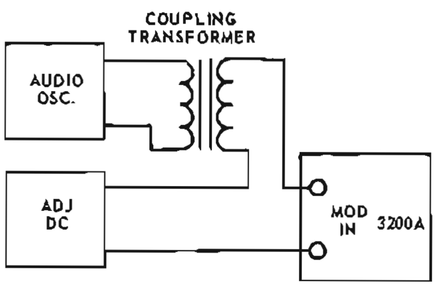


Figure 5. Setup for Obtaining Modulation in Excess of 40%

Provision is also made for external squarewave and pulse modulation through the front panel terminals (Figure 7). With the front panel modulation switch in the "pulse" position, the internal plate voltage supply to the oscillator is disabled and the oscillator plates are fed from an external source. Any varying source of signal, within the ratings of the 6DZ4 oscillator tubes, whose frequency is not limited by the input RF filters, may be applied. The signal source may be squarewave or pulse or remotely programmed dc. Again, some FM will be experienced due to direct plate modulation.

A power capability of 140 volts into 2000 ohms will drive the oscillator to maximum specified output on all ranges. Typically, however, approximately 10 volts peak (except 50 volts on the 260-500 mc range) will produce 1 milliwatt peak power output. For maximum tube life the peak voltage in pulse position should be limited to:

$$V_{peak\ pulse} = 150 - 1.8 \times V_{mod}$$

where V_{mod} = the voltage drop across the modulation terminals in the CW position. Since the modulation input circuit is ungrounded, either positive or negative pulses may be used, provided the more positive terminal from the generator is connected to the left-hand modulation terminals on the 3200A front panel.

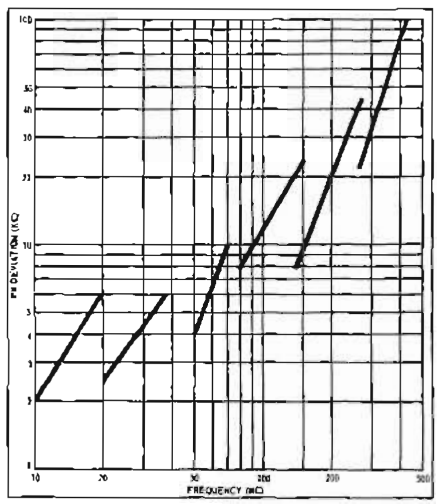


Figure 6. Typical FM Deviation for Fixed Audio Input Level

RF OUTPUT

The RF output of the oscillator is available through a unique, simplified waveguide-below-cutoff piston attenuator, shown in Figure 13. This attenuator provides a minimum of 120 db attenuation from maximum output and is adjusted by positioning the piston which can be readily locked in place by a rotary clamp. The attenuator piston is marked at intervals of 10 db attenuation. These graduations permit setting the attenuator to precise ratios. The attenuator has a bore diameter of 0.757 inches, providing an attenuation of 42 db per inch for all frequencies in the range of the instrument. The gradua-

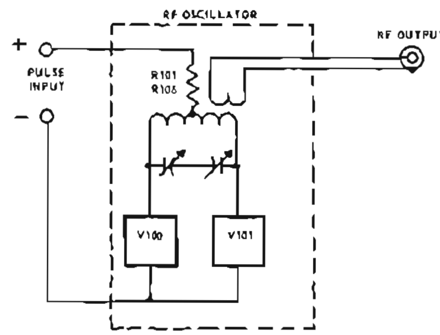


Figure 7. Pulse Modulation

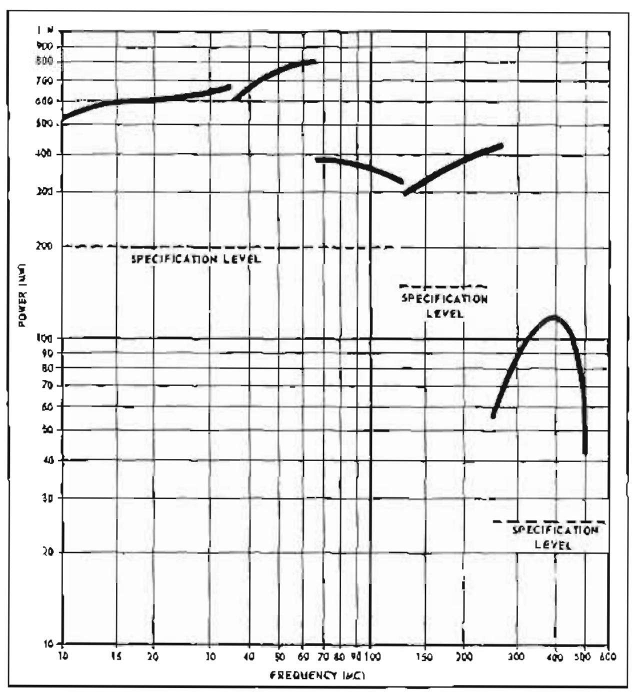


Figure 8. Typical Maximum Power Output

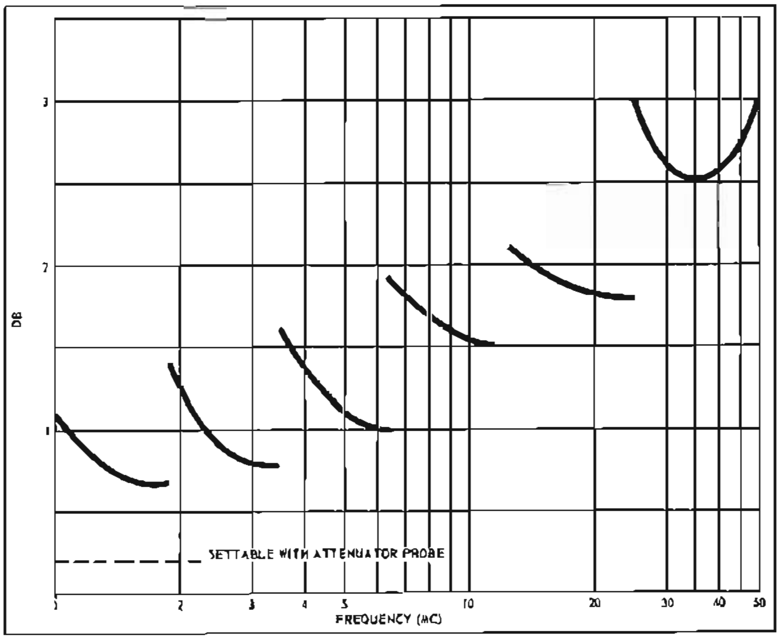


Figure 9. Typical Settability Using Electrical Vernier

tions also permit resetting to a particular power level. The first graduation will provide roughly 1 volt into 50 ohms and the last graduation will provide roughly 1 microvolt into 50 ohms.

The attenuator piston is completely removeable, and the pickup loop, employing a ferrite core for maximum low frequency coupling, is completely encapsulated in a low dielectric constant resin for maximum stability and protection. A simplified schematic of the output system is shown in Figure 2. In order to provide maximum available power output, no internal dissipative elements are employed. The output circuit is designed to feed an external 50-ohm load. If critical match or low VSWR is required, a suitable pad may be readily connected in series with the attenuator output.

Maximum RF output power is specified at greater than 200 milliwatts, 10-130 mc; greater than 150 milliwatts, 130-260 mc, and greater than 25 milliwatts, on the highest range, 260-500 mc. Curves of *typical* maximum output power, as a function of frequency, are shown in Figure 8. RF shielding, consisting of aluminum castings and compressed braid, will permit measurements at levels down to 1 microvolt.

In addition to the control of RF output level by positioning the attenuator piston, a front panel electrical vernier control is provided. This control varies the plate voltage on the oscillator and provides precise settability over a typical range as shown in Figure 9.

With the 3200A operating into a 50-ohm load, frequency shift from no load to specified output typically is less than 2%. If the 3200A is operated at maximum specified output into 50 ohms and then mismatched so as to increase VSWR up to 40 db, frequency shift will be less than 2% for typical data. See Figure 10.

POWER SUPPLY

All necessary operating voltages are provided by an internal solid-state power supply, shown in the simplified schematic, Figure 11. This supply provides regulated dc for the oscillator filaments and plates for minimum hum modulation and maximum tube life. In the circuit, the B supply reference is returned to regulated B— instead of unregulated B—. This reduces the current change through the reference tube

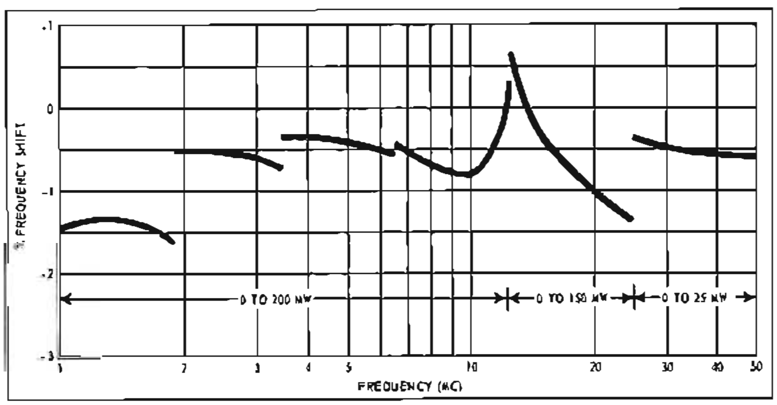


Figure 10. Typical Frequency Shift — No Load to Specified Output and Worst Mismatch

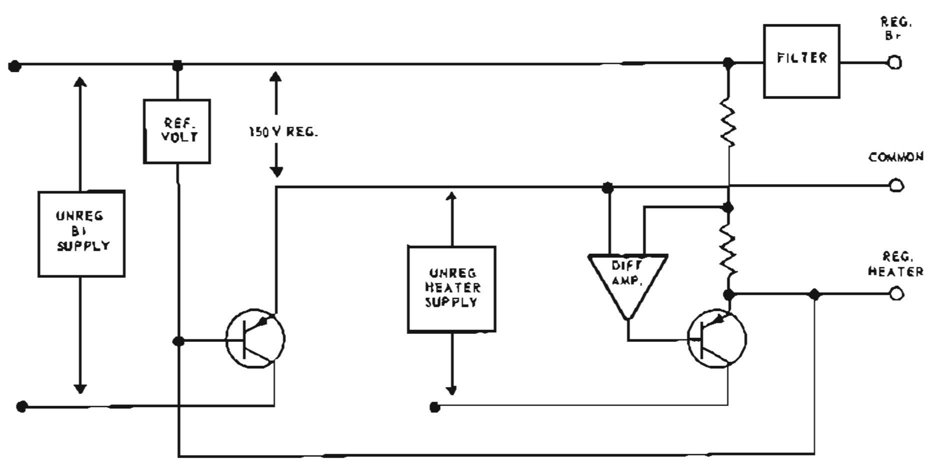


Figure 11. Power Supply

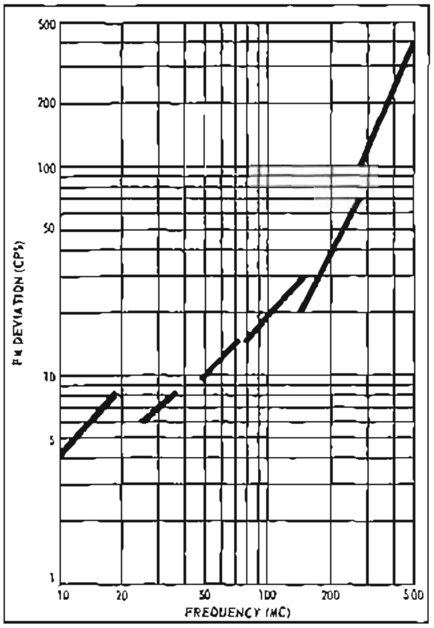


Figure 12. Residual FM Due to Hum and Noise

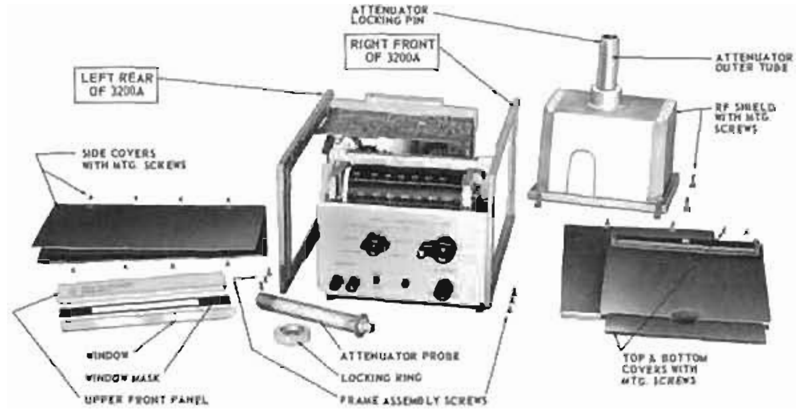


Figure 13. 3200A — Disassembled

caused by changes in unregulated B— or line voltage. The heater reference is tapped down from the regulated 150 volt supply, eliminating the need for a second reference. A filter in the B+ supply reduces hum, providing almost constant power operation of the oscillator tubes. Further reduction in the B supply hum is obtained by minimizing the common resistance in the B— and HTR+ return leads, and establishing the common tie point at the RF filter on the instrument baseplate casting. Residual FM, due to hum and noise, for a typical instrument is shown in Figure 12.

Frequency stability of the oscillator, as a function of input line voltage changes, is specified at 0.001% for a 5-volt change. The 3200A oscillator is designed for operation with an input line voltage of 105 to 125 volts and 210 to 225 volts, 50 or 60 cps. The unit has been operated from a 400 and 1000 cps power source, however, and has typically met its specifications at these higher line frequencies. Power consumption is 30 watts.

MAINTENANCE

As shown in Figure 13, considerable attention has been given to ease of maintenance. All of the power supply circuits are mounted on a stable circuit board and the power supply may be readily disconnected from the oscillator assembly through a multi-conductor plug. The entire RF casting can be readily removed by loosening just four screws.

New Transistor Test Jig Provides Y Parameters—500 KC to 250 MC

CHARLES W. QUINN, *Applications Engineer*

Figure 1. 13510A Transistor Test Jig

INTRODUCTION

Transistor measurement techniques utilizing the 250A RX Meter and special jigs built in the Boonton laboratory were described in articles published in Notebook Numbers 19, 20, and 26 and drawings showing the construction details of these jigs were made available to customers upon request. Over the past several years, however, Boonton has received numerous requests from customers to build and market these jigs and, as a result, has designed the 13510A Transistor Test Jig (Figure 1), an improvement over the original design. This new jig is designed to provide consistant, convenient, and precise readings over the entire the 500 kc to 250 mc frequency range of the RX Meter.

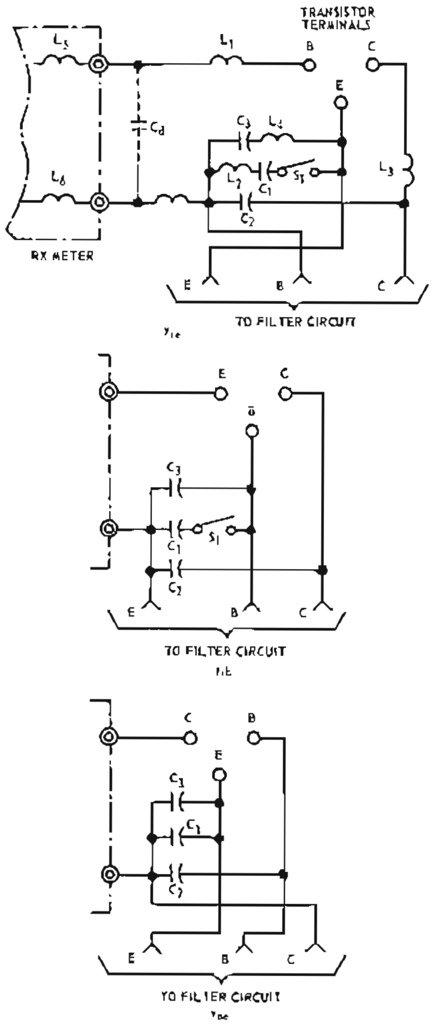
DESCRIPTION

The 13510A Transistor Test Jig consists of four basic components: a mounting adapter and three separate plug-in test circuits for measuring y_{ie}, y_{ib}, and y_{oe}. Included as part of the jig are bias feed and bypassing for an external power supply. The jig may be readily mounted on the RX Meter using the existing rear set of accessory holes on the ground plate surrounding the terminals. The parameters obtainable from the RX Meter readings are listed and defined below:

y_{ie} = Input admittance, common emitter output circuit short circuited

$$= \frac{1}{h_{ie}} = \frac{1}{R_p} + j\omega C_p \cong y_{11e}$$

y_{ib} = Input admittance, common base, output circuit short circuited



$L_1 = 3$ nh
$L_2 = 4$ nh
$L_3 = 8.5$ nh — lead and terminal inductances
$L_4 = 34$ nh
L_5, $L_6 = 3$ nh (RX Meter terminal inductances)
C_d = distributed capacitance of HIGH terminals
$C_1 = 29$ µf GE Tantalum, 30 v, bypass, HF
C_2, $C_3 = 0.1$ µf Scionics Polot, 50 v, bypass, VHF
S_1 = RF switch

Figure 2. Equivalent Test Circuits

$$= \frac{1}{h_{ib}} = \frac{1}{R_p} + j\omega C_p = y_{11b}$$

y_{oe} = Output admittance, common emitter, input circuit short circuited

$$= \frac{1}{R_p} + j\omega C_p = y_{22e}$$

DESIGN CONSIDERATIONS

A jig (or fixture) designed to work over the 250 kc to 500 mc frequency range of the RX Meter must, of course, be a broadband or untuned device. This means that the residual inductance must be held to a minimum and made as consistant as possible from one test circuit to another. These requirements have been met through the careful design of the test circuits. The circuits have been designed in printed circuit form for stability and are precision manufactured to insure maximum repeatability. Design of the 13510A jig is such that the RX Meter may be balanced for one test circuit and then used with other test circuits without rebalancing.

The bypassing system (Figure 2) is switchable so that capacitor C_1 may be removed from the circuit over a portion of the frequency range, thereby simplifying the corrections which would have to be made in the "crossover region" (1 to 10 mc). This switch is normally closed for measurements above 10 mc.

TRANSISTOR BIAS

Bias for the transistor under test is provided from an external supply, through the filter circuit on the 13510A jig, when the plug-in test circuit board is connected to terminals T_1, T_2, and T_3 (Figure 2) and the RX Meter terminals. Any dc supply which has less than 1 millivolt ripple (such as the -hp- 721A) may be used to provide transistor bias. Batteries may also be used if desired. The ideal supply for this application is one which provides a combination constant-current, emitter-base supply, and a constant-voltage (with adjustable current limit) for collector-base biasing. Many configurations are usable. Some examples are shown in Figure 4. It should be noted that the

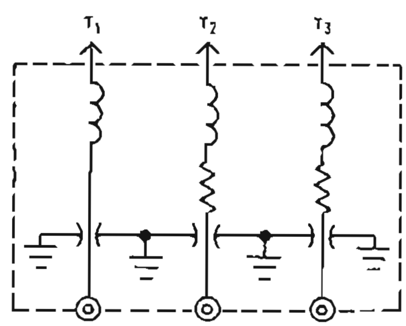


Figure 3. Equivalent Filter Circuit

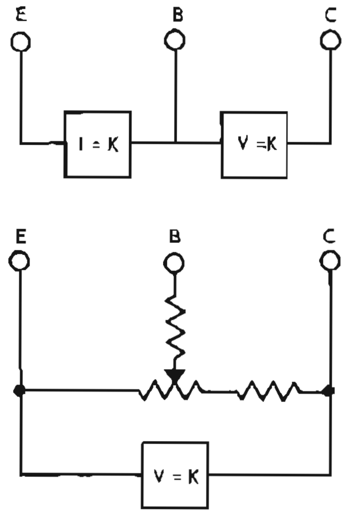


Figure 4. Recommended Bias Supply Configurations

limit for current passing through the RX Meter terminals is 50 milliamperes. Collector-to-base voltages should not exceed 30 volts. Levels at the bridge terminals should usually be kept below 20 millivolts.

CORRECTIONS

For most measurements, direct readings of the C_p and R_p dials on the RX Meter may be used. In cases of extremely low impedance (below 200 ohms) and high frequencies (above approximately 50 mc), however, some correction of these dial readings may be desirable. When measuring h_{ib}, for example, correction of the C_p and R_p dial readings would be indicated.

Corrections for series inductance can be made by adding the series inductance of the jig to that of the RX Meter. The equation to be used for these corrections is (Figure 5):

$$(1) \quad y(\ ) = \frac{y(\ )\, y_s}{y_s - y(\ )}$$

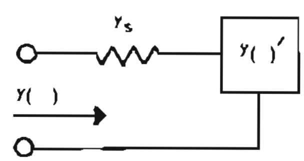


Figure 5. Correction Equation for Series Inductance

For frequencies above 10 mc, the value of y_s is nominally 10 nh. Below 10 mc, y_s is nominally 41 nh; most of which is contributed by the jig bypass electro-

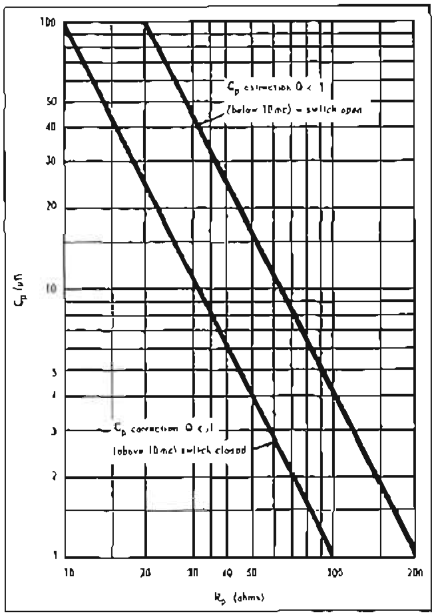


Figure 6. y_s vs. Frequency Correction — Below 50 mc

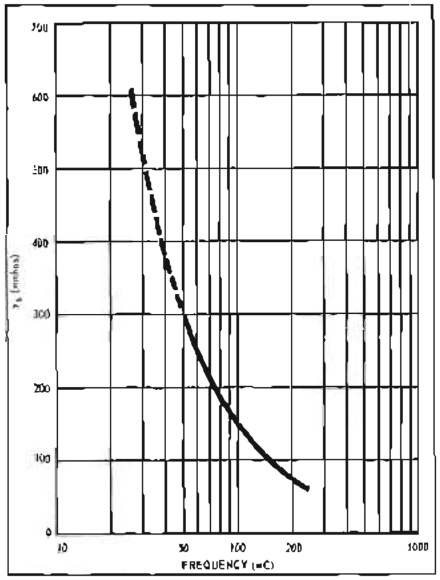


Figure 7. y_x vs. Frequency Correction — Above 50 mc

lytic capacitor, C_3. These corrections may be obtained more quickly through graphical solution as shown in Figures 6 and 7. For frequencies below 50 mc and impedances below 200 ohms, the correction curves in Figure 6 may be used. No correction is necessary for R_p below 50 mc. Above 50 mc, equation 1 should be used with the value of y_s obtained from Figure 7.

With the corrected values for y_{ie}, y_{ib}, and y_{oe}, known, y_{fe} can be calculated to a good approximation by the equation:

$$(2) \quad y_{fe} = \frac{y_{ib} - y_{ie} - y_{oe}}{y_{ie}}.$$

This equation is a more exact expression than the equation:

$$(3) \quad h_{fe} = \frac{h_{ie} - h_{ib}}{h_{ib}},$$

used in Notebooks Numbers 19 and 26.

EXAMPLE

The following is an example of the procedure and calculations used to determine the y_{ie}, y_{ib}, and y_{oe} parameters of an RCA 2N706 transistor utilizing the RX Meter and the 13510A Transistor Test Jig. Measurements were made at 250 mc with a bias supply of 6 volts, 2 milliamperes.

Determination of y_{ie}

RX Meter Readings:

Rp	*Cp*
74 ohms	+ 3.2 pf
74 ohms	— j 200 ohms

$$y_{ie} = G + j\beta$$
$$y_{ie} = 13.5 + j\,5.0 \text{ mmhos}$$
$$= 14.4 \,\angle{+20.6^\circ} \text{ mmhos}$$

Series Inductance RX Meter and Jig:

$$L_s = 10 \text{ nh}, X_s = 15.7 \text{ ohms},$$
$$y_s = -j\,62.8 \text{ mmhos}$$
$$y_s = 62.8 \,\angle{-90^\circ}$$

Corrected y_{ie}:

$$y_{ie}' = \frac{y_{ie}\, y_s}{y_s - y_{ie}}$$

$$y_{ie}' = \frac{14.4 \angle{+20.6^\circ} \times 62.8 \angle{-90^\circ}}{-j\,62 - 13.5 - j\,5.0}$$

$$= \frac{900 \angle{-69.4^\circ}}{-13.5 - j\,67.8}$$

$$= \frac{900 \angle{-69.4^\circ}}{68.8 \angle{-101.3^\circ}}$$

$$y_{ie}' = 13.1 \angle{+31.9^\circ}$$
$$= 11.05 + j\,6.9 \text{ mmhos}$$

Determination of y_{ib}

RX Meter Readings:

Rp	*Cp*
85 ohms	— 2.55 pf
85 ohms	+ 250 ohms

$$y_{ib} = G + j\beta$$
$$= 11.8 - j\,4.0 \text{ mmhos}$$
$$= 12.5 \angle{-18.7^\circ} \text{ mmhos}$$

Series Inductance of RX Meter and Jig:

$$y_s = -62.8 \text{ mmhos}$$
$$= 62.8 \angle{-90^\circ} \text{ mmhos}$$

Corrected y_{ib}:

$$y_{ib}' = \frac{y_{ib}\, y_s}{y_s - y_{ib}}$$

$$y_{ib}' = \frac{12.5 \angle{-18.7^\circ}\, 62.8 \angle{-90^\circ}}{-j\,62.8 - 118 + j\,4.0}$$

$$= \frac{785 \angle{-108.7^\circ}}{-11.8 - j\,58.8}$$

$$= \frac{785 \angle{-108.7^\circ}}{59.8 \angle{-101.4^\circ}}$$

$$= 13.1 \angle{-7.3^\circ} \text{ mmhos}$$

$$y_{ib}' = 13.0 - j\,1.67 \text{ mmhos}$$

Determination of y_{oe}

RX Meter Readings:

Rp	*Cp*
140 ohms	+ 5.3 pf
140 ohms	— j 120 ohms

$$y_{oe} = \frac{G}{7.1} \;\; \frac{+ j\beta}{+ j\,8.35} \text{ mmhos}$$

$$= 10.9 \angle{+49.6^\circ}$$

Series Inductance of RX Meter and Jig:

$$y_s = -j\,62.8$$
$$= 62.8 \angle{-90^\circ}$$

Corrected y_{oe}:

$$y_{oe}' = \frac{y_{oe}\, y_s}{y_s - y_{oe}}$$

$$y_{oe}' = \frac{10.9 \angle{+49.6^\circ} \times 62.8 \angle{-90^\circ}}{-j\,62.8 - 7.1 - j\,8.3}$$

$$= \frac{684 \angle{-40.4^\circ}}{-j\,62.8 - j\,71.1}$$

$$= \frac{684 \angle -40.4^\circ}{71.1 \angle -95.7^\circ}$$

$$= 9.63 \angle +55.3^\circ \text{ mmhos}$$

$$y_{no}' = 5.47 + j\,7.92 \text{ mmhos}$$

CONCLUSION

The 13510A Transistor Test Jig provides a simple and convenient method for measuring the Y parameters of transistors over the range from 500 kc to 250 mc. Through careful design, residuals have been minimized and, by employing printed circuit techniques, excellent unit-to-unit uniformity has been achieved. By applying the corrections described in this article, absolute measurements to accessories in the order of 10% can be obtained. Measurements, based directly upon RX Meter dial readings, provide a good basis for judging relative characteristics.

NEW DIRECT READING VECTOR IMPEDANCE METER TO BE SHOWN AT IEEE SHOW

The IEEE Show, March 22-25 in New York City will mark the introduction of our new and unique Vector Impedance Meter, Model 4800A, which provides automatic direct reading impedance measurements continuously from 5 cps to 500 kc. Impedance magnitude from 1 ohm to 10 megohms and phase angle from 0 to 360 degrees is instantaneously displayed on two front panel meters. Analog outputs, directly proportional to impedance magnitude, phase angle, and frequency are also available so that, by simple connection to an X-Y recorder, direct reading plots of impedance as a function of frequency may be conveniently obtained.

The Vector Impedance Meter will also function as a direct reading L-C meter covering ranges of 1 microhenry to 100,000 henries and 0.1 picofarad to 10,000 microfarads. By employing the "Q by delta f" approach, Q measurements can also be made.

See the new Vector Impedance Meter, the new Frequency Doubler Probe, as well as the VHF Oscillator and Transistor Test Jig in Booths Nos. 3501-3503.

NEW ACCESSORY 500-1000 MC FREQUENCY DOUBLER PROBE

A new Frequency Doubler Probe, Model 13515A, providing additional frequency coverage from 500-1000 mc from the 3200A VHF Oscillator, will be introduced at the IEEE Show, March 22-25. In operation, the Frequency Doubler Probe is merely substituted for the standard 3200A attenuator probe and doubles the output frequency of the oscillator in the 250-500 mc range. The doubler circuit, housed in the probe tip, is all solid-state and requires no tuning. Tentative specifications are given below:

RF RANGE: 500 to 1000 mc*
*With 3200A operating 260-500 mc (Range No. 6), 250-260 mc (Range No. 5)

RF OUTPUT:
Maximum Power: > 4 mw
(Across external 50-ohm load with VSWR < 1.1)

HARMONIC SUPPRESSION:
Fundamental: > 16 db*
Higher Order:
> 16 db* (500-800 mc)
> 14 db* (800-1000 mc)
* below desired signal

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