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Technical Information Note

NANOSECOND PULSE HANDLING TECHNIQUES IN I/C INTERCONNECTIONS

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The rapid advancement in the field of high speed digital Integrated Circuits has brought into focus many problem areas in the methods of pulse measurement techniques and new concepts dealing with these problems. This paper is intended to discuss the more common, yet perhaps not well known, pitfalls of measurement systems, a method of detecting them and possible solutions.

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INTRODUCTION

Evidence of the development of "millimicrosecond" instrumentation techniques can be dated back 15 years. Then, millimicrosecond, or as it is now called, nanosecond, instrumentation developments were chiefly in support of Atomic Energy Commission-sponsored nuclear research programs. More recently the gigacycle, now gigaHertz, computer research has intensified the need for better instruments and techniques to detect and to measure subnanosecond events.

Most electrical engineering students, by their senior year, have been exposed to the operation of various oscilloscopes of 10 to 50 megaHertz bandwidth in the laboratory. Many think that about 50 megaHertz is the upper frequency limit for these instruments. It is true that practical bandwidth limitations do exist for conventional "real time" oscilloscopes although the limitations have been overcome in various ways. The two most prevalent methods are the sampling concept and the tunnel diode switching comparator concept.

Aside from the fact that instrumentation is presently available for monitoring nanosecond response times, which even today is pushing the state-of-the-art, there exists problems in the laboratory of handling the signals. One of the major problems in nanosecond pulse measurement is the transmission of pulses from one point to another with a required fidelity. Another problem is the introduction of signals into circuits, and the removal of signals from circuits. This problem is present largely because of the effect of stray parameters associated with the components and wiring used to inject or remove signals.

SYSTEM PRECAUTIONS

An immediate word of caution here concerning the response time being measured by a system with limited inherent response time, since the observed reading may be in error. A good rule of thumb to follow is to have a system response time an order of magnitude better than the response of the device under test ($t_{r\,DUT}$). Table 1 lists some observed rise times ($t_{r\,obs}$) with their corresponding percentage of errors when viewed with a limited system response time. (System response time = t_{rs})

t_{rs} ns	$t_{r\,DUT}$ ns	$t_{r\,obs}$* ns	Error %
1	1	1.41	41
1	2	2.24	12
1	3	3.16	5
1	4	4.12	3
1	5	5.10	2
1	10	10.05	0.5

$^{*}t_{r\,obs} = \sqrt{t_{rs}^2 + t_{r\,DUT}^2}$

TABLE I

To yield a realizable concept in dealing with pulses in the nanosecond region, the practical tool is to relate the 10 to 90% rise time and pulse width to an equivalent frequency bandwidth (BW). In this corrolation, we assume an ideal rectangular pulse passing through a system, and characterize the system's response by analyzing the output pulse width, rise time, and general characteristics of the top of the pulse.

The rise time of the leading edge of the output pulse cannot be much less than one-half of the period of the upper frequency limit of the amplitude response of the system. Further, the total duration of the output pulse cannot be much longer than the half-period of the low frequency limit (and, of course, cannot be shorter than about twice the rise time). The importance of the concept of phase bandwidth must be emphasized in pulse circuits. It is preferable, from the point of view of obtaining minimum distortion, that frequency components should not be reproduced at all rather than arrive at the output in the wrong phase. In the latter case, the high frequency components not only do not add correctly to give a rapid pulse rise but also produce an undesirable series of "wriggles" which appear superimposed on the main pulse. The effect is easily recognizable, but difficult to eliminate, since the phase and amplitude bandwidths are often intimately related and it is not always easy to arrange for the amplitude response curve to "cut off" before the phase characteristic departs from linearity.

The rise time is then approximated by the relation

$$t_r = 0.45/f_c \qquad (1)$$

where f_c is the upper frequency limit defined by the 3 db point.

When a number of networks are connected in cascade, we are concerned with the cumulative deterioration in pulse rise time which occurs on passing through the whole system. If the quantities t_r are the rise times of the individual sections, when a rectangular pulse is applied to each separately, then the overall rise time T of the whole is

$$T = \sqrt{\Sigma t_r^2} \qquad (2)$$

This relationship was used to derive Table 1. The relation only applies strictly if the pulses are Gaussian in shape and provided there is no overshoot, but the equation gives an invaluable guide for general use in so-called minimum phase networks, composed of lumped components, and where the amplitude cut-off frequency alone is specified. Systems possessing wave propagating properties, however, are not minimum phase networks, and phase distortion may well determine the useful bandwidth if care is not taken. In normal practice, minimum phase networks are used with the exception of coaxial lines.

Using equation (1) and solving a typical application for a one nanosecond rise time pulse, we see that the frequency bandwidth is 450 megaHertz. In normal sinusoidal applications in the IF-RF region, a person is well aware of the precautions necessary for signal handling techniques, yet when the same persons deal with pulses, they tend to be negligent.

Our first major problem of concern is that of transporting our pulses from one point to another with the best fidelity. When transporting a pulse, one usually desires a minimum of change in the distribution of the pulse energy in time-space. This means a minimum of attenuation and dispersion in the transport process. For this reason, most nanosecond pulse propagation is done with the transverse electro-magnetic mode (TEM) in coaxial, stripline, or wire-over-ground-plane type transmission lines.

Stripline has its advantages in some applications which will be discussed briefly later. Wire-over-ground-plane comes into play in most test module construction and will likewise be mentioned later. Since coaxial transmission lines are practically incorporated in most every case, they will be treated immediately.

COAXIAL TRANSMISSION LINES

The discussion will be confined to the TEM mode propagation. Since nanosecond pulse work requires bandwidths of several handred megaHertz, dielectric losses are usually not a severe problem, until at the upper frequency limit when the dielectric losses become comparable to the skin effect loss. This occurs at about one gigaHertz for solid polyethelene dielectric cables. The major loss in good quality transmission lines, properly used in the nanosecond region is skin effect. When the dimensions in the direction of propagation begin to exceed the order of a wave length at the upper end of the required bandpass, the stripline and wire-over-ground type lines tend to have excessive radiation losses and the coaxial line, or other shielded construction, is superior.

Skin effect increases with the frequency according to

$$R = 4.2 \times 10^{-8} \sqrt{f} \left(\frac{1}{a} + \frac{1}{b}\right) \Omega/m \qquad (3)$$

where copper conductors with radii a and b are used in the coaxial cable. The attenuation is seen to rise with frequency resulting in a loss of the higher signal component frequencies. The distortion only appears as a slowing in the rise time, however, since there is no accompanying phase distortion (to the first order).

It may be noted that skin resistance may be reduced by plating the conductors with silver to a thickness equal to several times the skin depth. It may also be noted that the resistance of an outer conductor (shield) composed of thin wires braided together may be several times that of the corresponding metal tube and the variation with frequency seems to be greater, at high frequencies, than predicted by simple skin effect theory. The step function response of a skin effect limited cable has been shown to be non-Gaussian. It rises slowly for a short time to a small amplitude, and finally very slowly to a final value equal to the input pulse amplitude. (This assumes that the dc resistance is not significant compared to the characteristic impedance "Z_0" of the cable.) It is found that a unit step function input pulse, after traversing a length ℓ of cable, has a shape given by:

$$1 - \text{erf}\left(\frac{\ell \alpha_o}{2\sqrt{\pi f_o t}}\right) \qquad (4)$$

where α_o is the attenuation (nepers per meter) at some particular frequency f_o (Hertz), and t is the time (sec) excluding the normal delay down the cable. The function (4) is plotted in Figure 1. It is interesting to compare the rise time taken from Figure 1 with that predicted by the relation (1). If α_o is the attenuation constant at the "cut-off" frequency, i.e., at the 3 db point, then

$$e^{-\alpha_c \ell} = 1/\sqrt{2}$$

Now $$\alpha_c = \alpha_o \sqrt{f_c / f_o}$$

thus we find $$f_c = \frac{f_o}{\alpha_o^2 \ell^2} (\ln 0.707)^2$$

We then apply $t_r = 0.45/f_c$ with the result tabulated below along with the tangent-intercept and 10 to 90% rise time derived from Figure 1.

Measurement	Rise Time	
$0.45/f_c$	3.74	
10–90%	9.99	$\times \alpha_o^2 \ell^2 / f_o$ (all three)
tangent-intercept	1.64	

TABLE II — PULSE DISTORTION DUE TO LOSS IN COAXIAL CABLE

It is seen that the value given by equation (1) falls between the two extremes and is thus adequately justified as a rough practical measure.

It is pointed out by Kirsten, that if we normalize the abscissa in time and define T_o as the time for the wavefront to attain 50% amplitude, then

$$T_o = 4.56 \times 10^{-16} A^2 \ell^2 \text{ seconds} \qquad (5)$$

where A = attenuation of cable at 1000 MHz - db/100 feet

ℓ = length of cable in feet

In cases where the attenuation is known only at a frequency other than 1000 MHz, or the frequency dependence of attenuation departs somewhat from the one-half power law (say, where $\alpha = Kf^n$ in the region $0.4 < n < 0.7$), then

$$T_o = \frac{4.56 \times 10^{-7} \alpha_f^2 \ell^2}{f} \qquad (6)$$

where α_f = attenuation of cable at frequency f – db/100 feet

f = frequency – Hertz

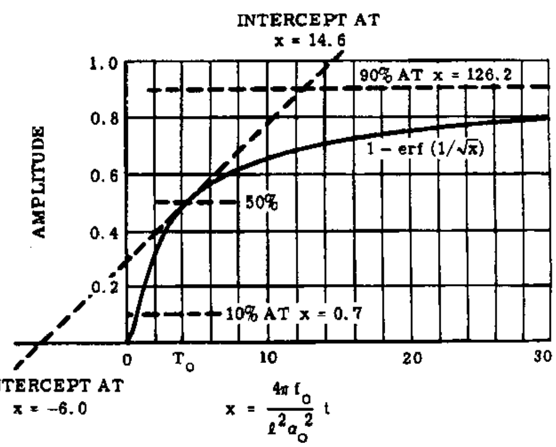

FIGURE 1 — CALCULATED DISTORTION SUFFERED BY A UNIT STEP FUNCTION ON TRAVERSING A LENGTH OF COAXIAL CABLE

Values of T_o for some commonly used cables are given in Table 3.

Cable Type	Z_o (ohms)	A (db/100' @ 1000 MHz	T_o (sec × 10^{-11}) ℓ = 10 feet
RG 8A/U	52	8.8	0.35
RG 9A/U	52	9.0	0.37
RG 58/U	53.5	20	1.8
RG 58A/U	50	23.5	2.5
RG 174/U	50	30	4.1
RG 196	50	46	9.6
RG 213/U	50	8.8	0.35
GR 874-A2	50	10.5	0.50
GR 874-A3	50	22	2.2

TABLE III

The times to reach other percentages of the input step amplitude are given in Table 4.

% Amplitude	Rise Time (% T_o)
10	0.17
20	0.28
50	1.0
70	3.1
90	29
95	110

TABLE IV

Of note here is the fact that the 10 to 90% response time of such a skin effect limited cable is

$$t_r = (29 - 0.17)\,T_o = 28.83\,T_o$$

which in rounding off is roughly 30 times as long as the 0 to 50% response time. Taking this into consideration, one can see where both the type and length of coax used are especially important in the subnanosecond region.

As mentioned previously and evidenced by equations (5) and (6), since the rise time T_o is proportional to ℓ^2, if two equal lengths of a given type of cable are cascaded, the rise time of the combination is four times the rise time of either length alone. This is in contrast to the well-known "Gaussian" response in which the rise time varies as the square root of the number of identical sections. For this reason, the rule of thumb for the overall response time as given in equation (2) is not applicable. Instead, the response of a system which includes cables with other elements may be obtained by other means. A practical method is given later using the Time Domain Reflectometer (TDR) method.

Another look at Table 3 will prove most significant later in the discussion of TDR and various anomalies or discontinuities in systems. Of the popular cables listed, the first three are not 50-ohm cables.

To summarize what has been presented so far:

1. Nanosecond and subnanosecond pulses contain frequency components in the UHF region and must be treated accordingly for minimum distortion.

2. Observed rise times may be considerably in error – caution.

3. The relationship $T = \sqrt{\Sigma t_r^2}$ only applies strictly if all components follow a Gaussian frequency distribution.

4. Skin effect is the major contributor to rise time degradation in coaxial cables in the nanosecond region.

5. The form of a step function, after traversing a skin effect limited coaxial cable of reasonable length, is a complementary error function.

6. The relationship $f_c = 0.45/t_r$ is a valid approximation.

7. The pulse analysis of systems, which include skin effect limited coax, does not follow the familiar $T = \sqrt{\Sigma t_r^2}$ Gaussian rule of thumb.

PARALLEL-PLATE TRANSMISSION LINES

We will now give attention to the planar, parallel plane or stripline, which may be considered as an evolution of the coaxial line. Thus, as illustrated in Figure 2, a circular-section coaxial line may be deformed in such a manner that both the center and outer conductors are first square and then rectangular. The narrow side walls are then taken to infinity to give a flat strip transmission system. This is a symmetrical line, but by removing one of the outer plates, an unsymmetrical system is obtained. The symmetrical line is normally referred to as "stripline", while the unsymmetrical is called "microstrip". Most practical striplines are unbalanced electrically. Parallel-plate lines may be dielectric supported or air-spaced. The conductors are usually of copper and the dielectric is a low-loss material such as teflon or silicon-impregnated fiberglass.

In striplines, a TEM mode solution is applicable as an approximation if certain boundary conditions are met. In such systems, the usual circuit elements can be constructed and counterparts of most coaxial line components may be realized. They are capable of broadband performance and complicated circuits can be readily constructed. Striplines have additional advantages of relatively small bulk and weight while they are capable of economic manufacture in quantity by printing techniques. They are finding increasing application in the frequency range up to 12 gigaHertz.

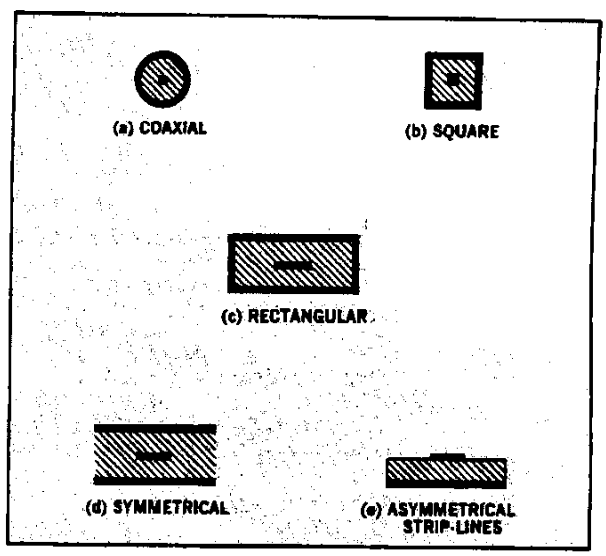

FIGURE 2 — EVOLUTION OF FLAT STRIP TRANSMISSION LINE

Figure 3 represents a typical microstrip transmission line. If the width of the ground plane is large compared with that of the strip, the electric field does not extend appreciably outside the dielectric and dominant propagation is in the TEM mode. If it were not for fringing and leakage flux, the theoretical characteristic impedance would be

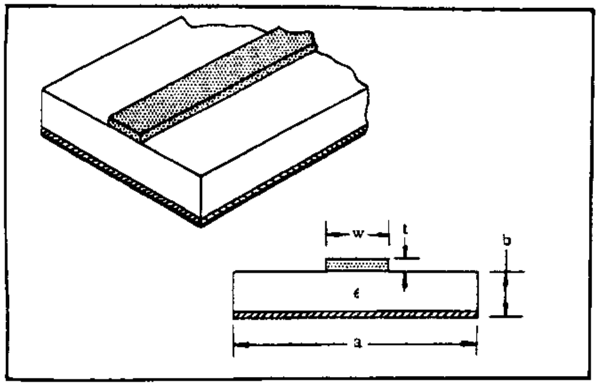

FIGURE 3 — A SOLID DIELECTRIC SUPPORTED "MICROSTRIP" TRANSMISSION LINE

$$Z_o = \frac{1}{C\,v_p} \qquad (7)$$

where C is the capacitance per unit length and v_p is the velocity of propagation within the dielectric.

To further insure a dominant TEM mode, the thickness b of the dielectric must be less than one-quarter wavelength in the dielectric.

Much analysis of parallel-plate system has been made. An analysis made by Bowness, involving the use of a measured constant for the fringing capacitance of the line, gave the characteristic impedance as

$$Z_o = \frac{10^4}{3\sqrt{\epsilon}\,\{7 + 8.83\,(w/b)\}} \qquad (8)$$

This result assumes that the strip conductor is then compared with its width and that w is greater than b.

At this point you might be asking what importance does microstrip play in present systems. Take a look at a copper clad circuit board mounted against a conducting surface or for that matter most printed circuit boards. You are looking at microstrip transmission lines which have characteristic impedances of a magnitude that can be detrimental to your circuit or system.

Figure 4 illustrates a general curve for microstrip transmission lines. Note that as t/w and w/b increase, Z_o decreases.

Figures 5 and 6, taken from the forementioned article are practical cases of the asymmetrical and symmetrical strip transmission lines. Figure 6 is experienced in multilayer printed circuit board techniques. Note that by adding the other ground plane, the characteristic impedances for a given line width are halved.

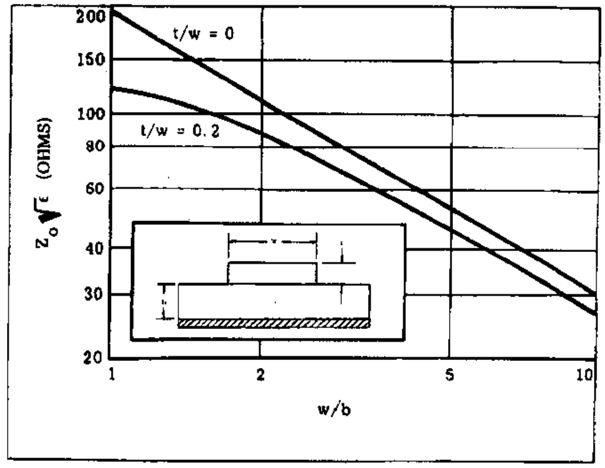

FIGURE 4 — CHARACTERISTIC IMPEDANCE OF ASYMMETRICAL STRIP-ABOVE-GROUND

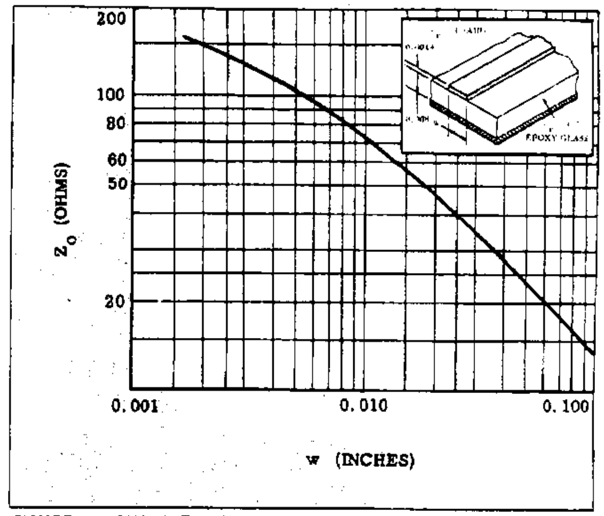

FIGURE 5 — CHARACTERISTIC IMPEDANCE OF SURFACE CONDUCTORS

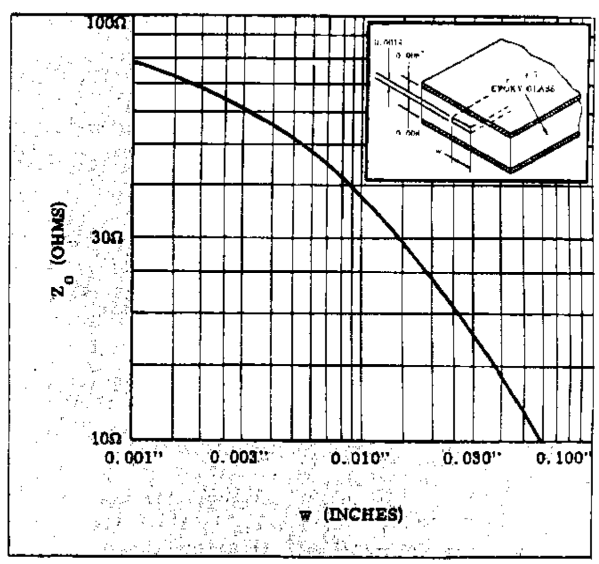

FIGURE 6 — CHARACTERISTIC IMPEDANCE OF BURIED CONDUCTORS

WIRE-OVER-GROUND TRANSMISSION LINES

Common to all systems is the familiar wire-over-ground construction. Using the relationship

$$Z_0 = \left(\frac{138}{\sqrt{\epsilon}}\right) \log_{10}\left(\frac{4D}{d}\right) \qquad (9)$$

where ϵ = dielectric constant between the wire and ground plane, D is the height of the wire of diameter d above the ground plane; one might see at first glance that Z_0 can be indefinitely large simply by making D/d indefinitely large. This fails rapidly in practice. If one used the diameter of an electron for d, and 50 light years for D, one would have an impedance of only about 3000 ohms. This apparently low value of 3000 ohms occurs because an order of magnitude increase in D/d results, at best, in only adding 138 more ohms to Z_0.

The significance of this becomes apparent when one investigates the inductance L and capacitance C of a section of this transmission line. The applicable relationships for an air dielectric are

$$L = 11.7 \log_{10}\left(\frac{4D}{d}\right) \text{ nH/in.} \qquad (10)$$

$$C = \frac{0.612}{\log_{10}(4D/d)} \text{ pF/in.} \qquad (11)$$

Due to the low rate of change of Z_0 with conductor spacing, clearly the more sensitive way to influence L or C is to change the length, ℓ. Since it is difficult to go much under 50 ohms or over about 200 ohms with wire-over-ground construction, 100 ohms is a useful number to keep in mind. With a 100-ohm "air" dielectric wire-over-ground type line, one has about 9 nH/in. of inductance and 0.8 pF/in. of capacitance.

Let us assume that we have a configuration where we have a one-inch length of wire as an intermediate wiring of a 50-ohm system. A length of 50-ohm line which would have the same 0.8 pF capacity would have about 4 nH of inductance. Thus there is an excess of inductance of 5 nH. The L/R time constant would then be 50 ps. (There is a 100-ohm loop impedance.) For a 1 ns rise time pulse, we would have about 20 time constants, or more than enough time to reach well over 90% of steady state conditions.

The approximation works well because the one-inch leads specified are short compared to the distance occupied by a 1 ns rise time pulse leading edge (nearly 1 foot in an "air" line). The significance of this is that in nanoseconds (or slower) wiring, the leads may be viewed as short lengths of transmission line whose impedance is fairly easily estimated; this allows the quick calculation of the excessive inductance or capacity over that required to match the operating impedance of the system, since leads are of necessity kept short compared to the distance in space occupied by the transient portion of the signal of interest.

One can now chuckle in retrospect at the superstitions and hysterical concern about wiring techniques in the microsecond and slower region. The impedance levels of thousands of ohms usually encountered implies that the leads will almost always have a Z_0 lower than the circuit impedance level, and therefore will usually have an excess of capacity. Note, however, that once a wire is 2.5 diameters above the chassis, moving it 25 diameters makes little difference, compared to typical tube and transistor capacities if the wire is only an inch or so long. By far, the more important effect is length. If the wires were two inches long, there would be 1.6 pF and 0.8 pF, respectively, for the 2.5 and 25 diameter-height lines; but going to 10 inches results in 16 pF and 8 pF. Thus raising a lead farther above the chassis probably promotes undesirable cross-coupling with other leads far more rapidly than it cuts down capacity. Decreasing the length of the lead is very desirable since this directly decreases capacity and cross-coupling. However, the inductance argument is usually ridiculous. Even if one had 10-inch leads of 500 ohms (about 2500 diameters above the chassis), in a 1000-ohm circuit, this would result in a time constant of less than 0.2 ns.

In nanosecond work, however, the impedance levels are frequently low enough that one is limited by stray wiring inductance, rather than capacity. The big advantage in nanosecond work is that circuit impedances are usually within the range of realizable transmission line impedance, thus one can arrange for an impedance match resulting in neither an excess of inductance or capacitance.

COMPONENT CONSIDERATIONS

Components also have their problems. High values of resistors suffer from capacitive troubles, since the RC time constant will generally be greater than L/R; and conversely low valued resistors have inductive problems. The inductance problem frequently shows up in terminations at the bottom of dividers used to either make a voltage source or signal attenuator, and in current metering devices such as a shunt; in other words, when one is interested in the voltage across the resistor because something is being placed in shunt with the resistor. The capacitive problem frequently shows up in terminations, in large dropping resistors used to either make a current source or voltage divider, or in general when one is interested in current through the resistor because something is going to be placed in series with it.

Inductance is usually not a problem, because it is sometimes convenient to arrange several resistors in parallel to further reduce the inductance; this process carried to the limit results in the disk-type resistor. A consequence of this parallel construction is increased capacity.

Resistor capacity is separated into two components—radial and axial. The axial capacity tends to peak the leading edge of a step waveform, while the radial capacity tends to roll off the leading edge.

A relatively large resistor used to change a voltage source into a "current" source is a good example of the problem. A 1000-ohm, 1/2-watt film resistor, 1/8" diameter and 1/2" long will most likely have under 10 nH of inductance; thus the L/R time constant will likely be under 10 ps. The axial capacity is in the order of 1/2 pF, resulting in a 1/2 ns time constant. The total of the distributed radial capacity is about 1/2 to 1/4 pF. With low source and load impedances, the radial capacity is relatively unimportant. Remember the radial capacity will not contribute to a single time constant due to its distributed nature. However, the axial capacity is going to peak up the leading edge of the step voltage. This can be reduced by placing the resistor through a hole in a ground plane, and placing the voltage source on one side and load on the other. Radial capacity difficulties can be compensated by adding axial capacity, preferably by using a wire extending from the high potential end to the load end and bowed away from the center of the resistor.

TIME DOMAIN REFLECTOMETRY

The question that is probably uppermost in your mind in reading this article might be this, "How do I find out what the true rise time of my system is?" or "What input waveform do I actually have at the input of my device?"

Up to this point, we have not touched on the problems in connection with impedance mismatches, reflections, VSWR's, discontinuities, etc. In obtaining a true picture (oscilloscope presentation) of your incident signal at the point you want, you will also be looking at the mentioned anomalies. The conditions which can cause them are too numerous to mention. In taking another look at Table 3, you can see that the first three cables listed are not 50-ohm cables. The worst case of RG58/U can cause a reflection of better than 5%. Now, if your system is not properly backmatched, this signal is re-reflected into your system. Depending on the position of the cable in the system, and the source to test point time relationship, this re-reflection could add to your original signal in most any fashion.

Returning to the original questions at hand, the only way to be certain of the rise time or waveshape of the test "step" at a particular point in a system is to use a time domain reflectometer (TDR) set-up and observe the reflection returning from a short at that point.

Many articles have been written on the subject of TDR. Among the best are Hewlett-Packard's Application Notes 62 and 67, and Tektronix's articles by Gordon D. Long on the subject of Pulse Reflection Measurements.

Of note is the fact that using present equipment in TDR set-ups one can detect minimum discontinuities resulting from $L = 10^{-12}$ Henry and $C = 4 \times 10^{-16}$ Farad. Of advantage is that discontinuities can be resolved as inductive, capacitive, or resistive, along with the physical location in the system. The practical limit of location or position resolution is about one-half the width of the impulse response, i.e., a little more than half the nominal rise time. The HP 140A/1415A can resolve discontinuities down to about 1 cm. Since this is essentially the discontinuity that any signal with frequency components below about 3 GHz will see, the test is valid for this frequency range (i.e., dc to 3 GHz). Herein lies another important advantage of TDR – it is a wide-band system. This in itself has many implications.

For example, consider the wide-band input impedance (r_b') of a transistor. With a TDR system, and the utilization of microstrip techniques, r_b' can be determined over the quoted frequency bandwidth, by reading it on a CRT display as one would read an ohmmeter. This is opposed to the many signal generators, test fixtures; dc, audio, and RF voltmeters; etc., that would be required for a point-by-point analysis.

Many applications of TDR have yet to be explored, but as we can see in this discussion, it can be an invaluable tool in merely finding out if a test system or test fixture does have problems, and what and where those problems are.

CONCLUSIONS

We have, in the course of this note, touched on systems as a whole; the different transmission lines (apparent and non-apparent) and individual components used in the system; and a method whereby deficiencies can be detected, observed, and possibly eliminated. A paper can be written on each of the topics mentioned. However, as as originally stated, the intention of this note is to create a general awareness of the common, yet not well known problems, and generally mention possible solutions.

REFERENCES

I.A. Lewis and F.H. Wells: "Millimicrosecond Pulse Techniques," Pergamon Press, 2nd Ed., 1959.

A. F. Harvey: "Microwave Engineering," Academic Press, 1963.

C. N. Winningstad: "Nanosecond Pulse Measurements." Tektronix, Inc.

F. Kirsten: "Pulse Response of Coaxial Cables," Lawrence Radiation Laboratory, Counting Note No. CC2-1A.