APPLICATION NOTE

Wide-band linear power amplifiers (470 – 860 MHz) with the transistors BLW32 and BLW33

ECO7806

Philips Semiconductors

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Application Note ECO7806

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10.1 BLW32 AND BLW33 AMPLIFIERS

1 ABSTRACT

For application in T.V. transposers for band IV/V (470 – 860 MHz) some wideband linear power amplifiers have been made with the BLW32 and BLW33. A single stage amplifier with BLW32 gave a minimum output power of 380 mW peak sync. at a 2-tone i.m.d. of –47 dB. The power gain was 11.4 ±0.55 dB. A similar amplifier with the BLW33 produced a minimum output of 790 mW peak sync. at the same i.m.d. and a power gain of 10.4 ±0.85 dB.
A 2-stage amplifier with the BLW32 as driver and BLW33 in the output showed a minimum output power of 750 mW peak sync. at –47 dB i.m.d. with an overall power gain of 22.5 ±2 dB.

2 INTRODUCTION

This report describes the theoretical aspects and practical realisation of some wide-band UHF power amplifiers for TV transposer service in bands IV and V (470 – 860 MHz).

The amplifiers are designed with the BLW32 and BLW33 transistors. These devices are intended for resp. 0.5 and 1 W peak sync output and are developed for ultra linear applications. For this purpose the transistors have to operate in class A.

The transistors are encapsulated in a ¼ inch capstan envelope with ceramic cap.

Because of their high gain they are excellently suited for the realisation of wide-band type amplifiers.

3 THEORETICAL CONSIDERATIONS

3.1 The equivalent circuit of the transistor input and output of the BLW32 and BLW33

For class A operation the BLW32 and BLW33 are specified as follows:

BLW32: V_{CE} = 25 V and I_C = 150 mA

BLW33: V_{CE} = 25 V and I_C = 300 mA.

Although the power gain, the input impedance and the optimum narrow band load impedance versus frequency are given in the Data sheets for the above operating points, the actual class A operation points chosen differ somewhat for the circuits described in this report.

Namely, from earlier investigations it is known that the wide-band properties are more favourable for lower load impedances. So, we have chosen for the following class A operation points:

BLW32: V_{CE} = 22.5 V and I_C = 165 mA

BLW33: V_{CE} = 22.5 V and I_C = 330 mA.

The corresponding typical gain, input and load impedances have been calculated. The values thus obtained for three frequencies are given in Tables 1 and 2.

Table 1 BLW32

f (MHz)	gain (dB)	R_i (SERIES) (Ω)	X_i (SERIES) (Ω)	R_L (SERIES) (Ω)	X_L (SERIES) (Ω)
470	16.5	4.59	1.67	41.5	39.4
636	14.0	4.42	3.61	28.9	38.0
860	11.4	4.12	5.86	17.7	33.0

Table 2 BLW33

f (MHz)	gain (dB)	R_i (SERIES) (Ω)	X_i (SERIES) (Ω)	R_L (SERIES) (Ω)	X_L (SERIES) (Ω)
470	15.0	3.60	1.90	27.1	17.4
636	12.5	3.45	3.26	21.0	18.4
860	10.1	3.19	4.90	14.5	17.3

To facilitate calculations approximate equivalent circuits for the transistor input and output impedances can be given. They are shown in Figs 1 and 2.

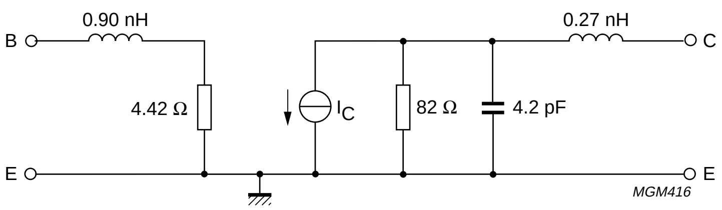

Fig.1 BLW32.

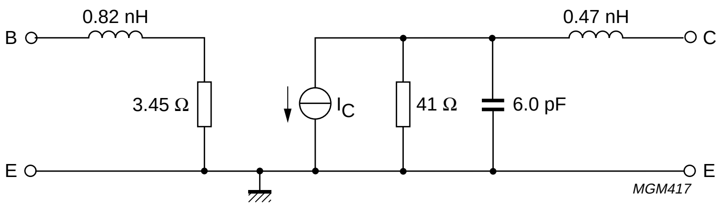

Fig.2 BLW33.

3.2 The output networks

The circuits will be designed on printed circuit boards with P.T.F.E.-fibre-glass as a dielectric with an ε_r = 2.74 and a thickness of 1/16 inch.

The input and output network start with a piece of stripline having a width of 6 mm, being the width of the base and collector leads. For a dielectric of 1/16 inch the characteristic impedance Z_C is 37.6 Ω. The length for the collector leads amounts to 3 mm. The base leads are different in length.

As the output impedance of both transistors is rather high a Chebyshev bandpass-filter configuration will be chosen to match it to the 50 Ω load. As will be shown later 6 elements are sufficient to obtain an acceptable VSWR through the band.

The bandpass filter is derived from a low-pass prototype having a cut-off frequency equal to the bandwidth of the final filter, i.e. 860 – 470 = 390 MHz and a characteristic resistance of 50 Ω (see Ref.1). The transformation procedure is depicted in Figs 3 to 7. These figures are for the BLW32.

To determine $C_1 = C_3$ in Fig.3 we have to keep in mind that the input (and output) RC-product remains constant with impedance transformation, so: $C_1 = C_3 = 82 \times 4.2/50 = 6.9$ pF.

Next, we determine the quantity γ being equal to: $\gamma = \dfrac{1}{\omega_c \times C_1 \times R}$

In which ω_c is the cut-off frequency of the low-pass prototype and $C_1 \times R$ is the already mentioned time constant, so:

$$\gamma = \frac{1}{2 \times \pi \times 390 \times 10^6 \times 6{,}9 \times 10^{-12} \times 50} = 1{,}183$$

Then L_2 can be calculated according to: $L_2 = \left\{\dfrac{R}{\omega_c}\right\}\left\{\dfrac{2\gamma}{\gamma^2 + 0{,}75}\right\}$

in which R is the characteristic resistance of the filter being 50 Ω, so:

$$L_2 = \left\{\frac{50}{2 \times \pi \times 390 \times 10^6}\right\}\left\{\frac{2 \times 1{,}183}{1{,}399 + 0{,}75}\right\} = 22{,}46 \text{ nH}$$

The transformation from low-pass to bandpass is made by:

1. Shunting each capacitor by an inductor and
2. Putting the capacitor in series with each inductor, such that resonance is obtained at the geometric mean frequency of the band: $f_O = \sqrt{860.470} = 635.8$ MHz

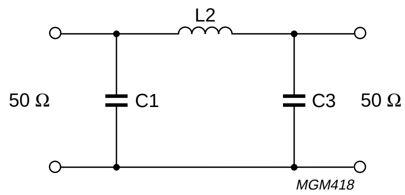

Fig.3

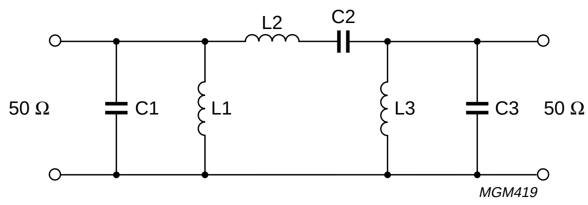

Fig.4

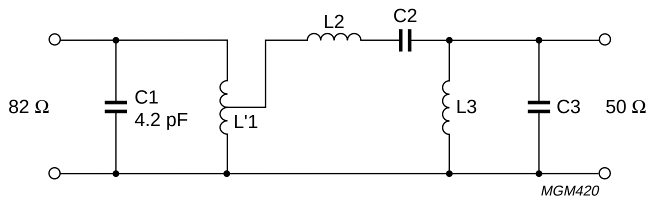

Fig.5

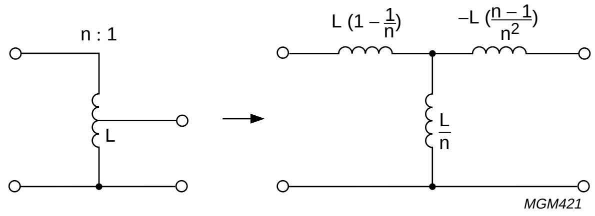

Fig.6

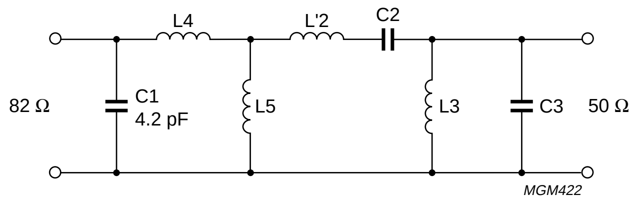

Fig.7

This means in Fig.4 that:

$L_1 = L_3 = 9.081$ nH

$C_2 = 2.79$ pF

Now we transform the input part of the filter to the actual output impedance of the transistor (see Fig.5). This is done with an ideal transformer having a primary inductance, $L'_1 = 82/50.9.081 = 14.89$ nH and a transformation ratio n : 1 in which

which $n = \sqrt{82/50} = 1.281$

This ideal transformer is subjected to a Norton transformation (see Fig.6) in which the inductances become 3.27 nH, 11.62 nH and –2.55 nH resp. The final situation is then depicted in Fig.7 with $L_4 = 3.27$ nH, $L_5 = 11.62$ nH and $L'_2 = L_2 - 2.55$ nH $= 22.46 - 2.55 = 19.91$ nH.

The resulting maximum VSWR is equal to:

$S = \{(x^3 + 1)/(x^3 - 1)\}^2$ in which:

$$X = \gamma + \sqrt{\gamma^2 + 1}$$

For our filter this becomes: S = 1.22, which is a very acceptable value justifying the conclusion that the number of filter elements is sufficient.
The next step is the transformation of this filter to a stripline circuit. If a transmission line is shorter than 1/8 of a wavelength the following equivalence is reasonably accurate (see Fig.8):

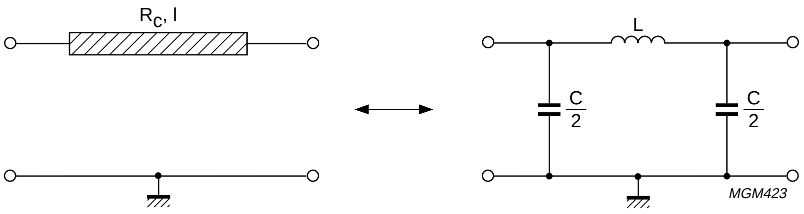

Fig.8

$$L = \frac{R_C \times l}{v} \text{ and } C = \frac{l}{R_C \times v}$$

in which v is the propagation speed being 3.10^8 m/s for air-line. Starting from Fig.7 we must keep in mind that C_1 and part of L_4 (viz. 0.27 nH) are inside the transistor.
The inductors are then replaced by striplines. The consequence of this transformation is that at some points in the circuit parasitic capacitances are introduced.
This can be corrected by adjusting the values of L_5 and C_3. A second reason for correcting the latter is its parasitic series inductance. The remaining part of L_4 is composed of 2 pieces of stripline with different characteristics impedances. The part with the lower R_C serves to accomodate the collector lead of the transistor. C_2 has a parasitic series inductance of appr. 1 nH which must be subtracted from L_2'.
The remaining part is split up in 2 equal pieces on both sides of C_2. The result of the transformation described above is shown in Fig.9 and Table 3.

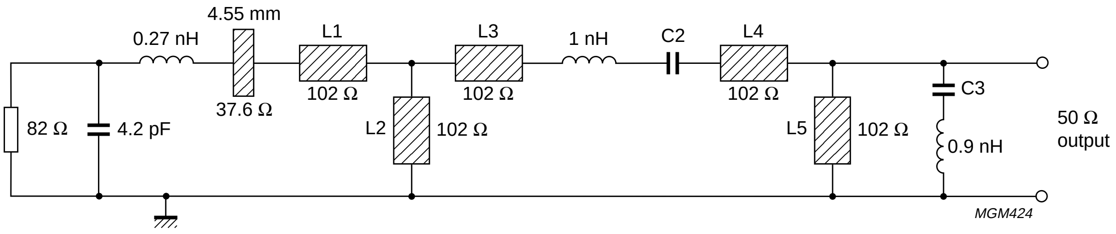

Fig.9 See also Table 3.

An exact analysis of this circuit results in a maximum output VSWR of 1.66 which is rather high. Therefore the complete circuit is subjected to a computer optimization procedure with the object to reduce the maximum output VSWR to the lowest possible value. The results are given in Table 3.

Table 3

ELEMENT	BEFORE OPTIM.	AFTER OPTIM.	UNIT
l_1	7.15	6.84	mm
l_2	28.4	27.7	mm
l_3	27.8	29.8	mm
C_2	2.79	2.76	pF
l_4	27.8	29.8	mm
l_5	26.7	33.4	mm
C_3	5.14	4.01	pF
S_{max}	1.66	1.35	–

As a final step the practical dimensions of the striplines must be determined. the printed-circuit board material is P.T.F.E. fibre-glass with a thickness of 1/16 inch and a dielectric constant ε_r of 2.74. The lines with R_C = 102 Ω must have a width of 1 mm. The length reduction factor is 1.43. The collector pad (R_C = 37.6 Ω) must have a width of 6 mm and its length reduction factor is 1.52. The results of the latter transformation can be found in the final circuit diagram and parts list.

3.3 The input networks

The approach followed for these networks is rather similar to those described in Refs 2 and 3. However, there is one exception:

Because of the higher input impedance of the BLW32, as compared to the BLW98, a smaller amount of transformation is required. As a consequence of this a single section matching network is sufficient. Generally speaking it can be said that this section must have a loaded Q-factor of appr. 4 to produce the required gain compensation.

This means that the total inductive reactance (of transistor and matching network) must be 4 times the input resistance of the transistor (see Fig.10).

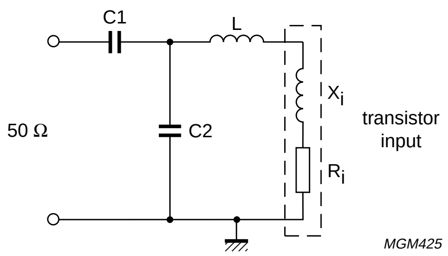

Fig.10

The tuning of this circuit is then at appr. 860 MHz.

The single section transforms the transistor impedance to a value above 50 Ω, so that some ‘back-transformation’ is required. This is done by reducing the value of the blocking capacitor C_1 (8.2 pF instead of 100 pF).

In case of the BLW33 the input impedance is somewhat lower, so a 2 section network is advisable. Here, the blocking capacitor is 100 pF.

The final steps are the same as in the previous section, viz. transformation to stripline and computer optimization. The results can again be found in the circuit diagrams.

4 THE SINGLE AMPLIFIERS WITH BLW32 AND BLW33

4.1 Practical considerations

On previous pages the theoretical approach has been discussed. In practice it was the intention to realize small compact amplifiers on a printed-circuit board with the input and output terminals (R_C = 50 Ω) in-line for easy cascading of several amplifiers.

The printed-circuit board needs to be double copper clad and has a P.T.F.E. fibre-glass dielectric for low losses at UHF. With a typical ε_r of 2.74 we had the choice between a thickness of 1/16 and 1/32 inch.

In principle 1/32 inch is possible but especially the 102 Ω lines in this design become so narrow, that the risk of under etching is too high.
So, we have decided to use a dielectric of 1/16 inch, what means that these 102 Ω lines are 1 mm wide.

Another problem being met with the application of 1/32 inch is the very small surface for soldering of the multi-layer chips and also an unusual ratio between the width of chip and track.

Figure 11 shows the circuit diagram of the BLW32 class A amplifier with biasing, whilst in Fig.12 the circuit is drawn for the BLW33. Both circuits contain a bias network with a PNP transistor type BD136. The lists of components are given in Tables 7 and 8.

The printed-circuit boards are in Figs 13 and 14.
For a correct earthing the upper earth sheet parts are directly connected to the lower sheet by soldering copper straps at the edges of the printed-circuit board.

The emitters were grounded as short as possible by applying copper straps under the emitter leads. For that reason the holes in the board were square instead of round.

All components are situated on one side of the board viz. the side of the tracks.

The transistors were screwed to an extruded aluminium heatsink, resulting in an average stud temperature of appr. 50 °C.
In both circuits tuning capacitors are applied. They are of the film dielectric type with three tags. Both earth terminals are fed through holes and soldered to the upper and lower earth plane.

In the experimental phase, the coupling capacitors in the collector bandpass sections are of the ‘Micro thin-trim’ type 1.0 – 4.0 pF. Later, they can probably be replaced by fixed multilayer chip capacitors.

The coaxial connectors are of the SMA 50 Ω type. They are soldered to the upper and lower earth plane.

In general, earth connections have to be made as short as possible.

4.2 Practical optimization method

Both prototypes were built according to the result of the theoretical part. It proved that such a theoretical approach was very valuable. It led, via some relatively small modifications to the ultimate design.

Optimization on a small signal basis has been done with the circuits inserted in a network analyzer chain, having swept S (scattering)-parameter facilities. From this set-up one can examine under dynamical conditions the effect on the power gain (represented by S_{21}), input- and output impedances (resp. S_{11} and S_{22}) and feedback (S_{12}) by tuning and modifications on the circuit of the amplifier.

The correction of power gain over the whole 470 – 860 MHz range is arranged by applying appropriate mismatch. The resulting high VSWR, especially at low frequency, is not reduced in an equalizer.

So, tuning on an S-parameter set-up means that the S_{21} has to be optimized for flat gain, and at the same time the input reflection damping S_{11} tuned for minimum at the highest frequency of 860 MHz.

4.3 Measured results

Figures 16 to 21 show the practical results for resp. the BLW32 and BLW33 amplifiers.
The results for two units are both given. They are numbered (1) and (2).

For a clear interpretation of the S_{11} and S_{22} readings the expression for the reflection damping and some figures are given in Table 4:

Table 4 Refl. damping = 20 log (S + 1)/(S – 1) dB in which S is the voltage standing wave ratio (VSWR)

REFL. DAMPING (dB)	REFL. COEFF.	S
0	1	∞
4	0.631	4.42
8	0.398	2.32
12	0.251	1.67
16	0.158	1.38
20	0.1	1.22
24	0.063	1.13
28	0.04	1.08

Examining Figs 16 and 19 one will see that in both cases, the reflection damping is about 20 dB so the input VSWR is about 1.2 near 860 MHz, whilst the output VSWR, with respect to 50 Ω, varies between 3 and 1.06 over the band.

After optimization at small signals, the i.m.d. has been measured in a 2-tone set-up. Although it is advised to apply the 3-tone test method for determining the i.m.d. in case of TV systems, the 2-tone test shows a good correlation with the former (see Ref.3).

In practice this means that for a 3-tone test with the tones at –7, –8 and –16 dB below the 0 dB peak sync. level, the in-band i.m. product has to be at least –60 dB down, whilst the same amplifier has to show a distance of two i.m. products d_3 of at least –47 dB with respect to one of the equal tones in case of the 2-tone test with equal peak output power.

The results are given in the curves of Figs 22 to 25.

For these amplifiers, of which two each have been constructed and measured, the results are according to the expected ones for wide-band operation. At three channels the peak sync power for a 3-tone i.m.d. of –60, –56 and –52 dB has been measured too. Table 5 gives an impression of the average –60 dB results of the BLW32 and BLW33.

Table 5

CHANNEL NO.	BLW32	BLW33
21 (471.25 MHz)	445 mW	1286 mW
39 (615.25 MHz)	671 mW	1308 mW
70 (863.25 MHz)	500 mW	941 mW
P_O sync. for –60 dB i.m.d. (average of two amplifiers)		

Tables 10 and 11 show the more complete test results.

Comparing the afore mentioned 3-tone (–60 dB) results with the 2-tone (–47 dB) figures it appears that: As for the single amplifiers the typical low frequency decoupling elements are divided over the bias circuit and the amplifier circuit. This was done because in case of concentrating all low-frequency decoupling elements on the bias board, parasitic oscillations occurred.

To get some idea of the reproducibility of practical amplifiers, two equivalent units have been built. They are screwed to heatsinks having a thermal resistance of appr. 2 °C/W. So for an average ambient temperature of 25 °C the stud temperature rises to about 48 °C for a power dissipation of about 11.5 W.

For the single amplifiers described, the straight forward method of small signal tuning at the S-parameter equipment and afterwards the 2-tones test was followed. Via this method reliable results could be obtained.
However in case of the double amplifier it appeared to be rather difficult to find the optimum between an acceptable flat gain curve and sufficient output power with low i.m.d.

Because there is a correlation between the single tone 1 dB compression point and the i.m.d. of a linear amplifier, the following dynamic way of tuning was applied: For the BLW33 compression starts around 2 W output. The gain of the complete amplifier is appr. 20 dB so appr. 20 mW (+13 dBm) drive power should be sufficient. This amount of swept power is available from the sweep oscillator HP8620C in combination with RF plug-in unit HP86222A. Because of internal ALC the swept output power level is very stable and almost unaffectable by the strong input VSWR variation of the BLW32 amplifier.

- For the BLW32, the channel 21 and 39 results are somewhat better (3 – 18%) when measured under 3-tone conditions, whilst the channel 70 figures are about 6% lower.
- In case of the BLW33, the 3-tone results are somewhat different for both tested units. As an average they are in between those of the 2-tone test.
- The gain figures are about the same. For a description of the 2-tone and 3-tone test chain and the way of testing one is referred to Ref.3.

5 COMBINED AMPLIFIER WITH BLW32 AND BLW33

5.1 Practical considerations

As an experiment both amplifiers, being tuned for optimum performance, were cascaded. In that case the output of the BLW32 has to accept the mismatch, specially at lower frequencies, caused by the BLW33 amplifier.

The gain (S_{21}) becomes appr. 20 dB, whilst the input and output reflection damping of the combination show somewhat different values. However, the trend is the same.

From this experiment the idea came to integrate both amplifiers on one printed-circuit board. The easiest solution was the direct connection of the two 50 Ω striplines, maintaining the blocking capacitor C_1 (see Fig.12) and concentrating the elements C_{11} (see Fig.11) and C_2 (see Fig.12) in a new single tuning element.

Figure 26 shows the total circuit diagram, Fig.15 the printed-circuit board of the amplifier part and Table 9 the list of components.

To create an amplitude variation the sweep generator is amplitude modulated (AM input) with a square wave pulse generated with a PM5715 pulse generator.

The output of the amplifier under test is now again attenuated (about 20 dB), detected and supplied to an oscilloscope being horizontally driven by means of the available sweep output of the HP8620C.

Figure 27 shows the block diagram of the measuring set-up, and Figs 28 to 30 screen pictures of the swept output power for different input levels (+7, +10 and +13 dBm) in which compression is clearly shown. From above mentioned experiments it appeared that the blocking capacitor, C_{12} in Fig.26, better could be decreased to 8.2 pF.

5.2 Measured results

Being tuned for a good large signal behaviour the i.m.d. results of both amplifiers constructed are given in Figs 31 and 32. Figures 33 and 34 resp. show the gain curves for an input power of 5 mW (+7 dBm) and Figs 35 to 37 the S-parameter results.

The peak sync. power for a 3-tone i.m.d. of –60, –56 and –52 dB has been measured at three channels. Table 6 shows the –60 dB results for the combination no. (2).

Table 6

CHANNEL NO.	BLW32 – BLW33
21 (471.25 MHz)	1045 mW
39 (615.25 MHz)	895 mW
70 (863.25 MHz)	587 mW
P_O sync for –60 dB i.m.d. (amplifier no. 2)	

Table 12 shows the more complete test results.

Comparing the 2-tone (–47 dB) and 3-tone (–60 dB) results it appears that:

- The combination BLW32/33 shows appr. equal results for channels 21 and 39. However the channel 70 results are somewhat lower.
- The gain figures are about the same.

6 CONCLUSIONS

On preceding pages the theoretical and practical designs have been described of some wide-band (470 – 860 MHz) high quality linear amplifiers, utilizing the BLW32 and BLW33 transistors operating in class A.

The class A operation points, in these wide-band applications, differ somewhat from those being published in the data sheets for narrow band applications. This led to a better wide-band loading, what again resulted in a better i.m.d. performance.

The flatness of the power gain versus frequency is obtained by applying an increasing amount of mismatch when the frequency becomes lower.

This can be seen as a disadvantage of the system, because the preceding driver stage has to accept that mismatch. In the present combination, however, the driver is so large that no difficulties are experienced.

7 RECOMMENDATIONS

A flat power gain characteristic combined with a low input VSWR may be obtained by:

1. Applying an equalizing network at the input.
2. A coaxial isolator or circulator at the input.
   However, these devices have restricted bandwidths.
3. Applying the method in which two equal amplifiers have been connected in parallel with the aid of two wide-band 3 dB –90 °C coaxial hybrids on a 50 Ω basis. In that configuration the output power will be nearly doubled, whilst the input VSWR of the system becomes max. 1, 2 a value mainly given by the properties of these devices. The reflected power will be absorbed in the resistor matching the isolated port.

   A suitable type of hybrid is the ultra-miniature 3 dB –90 °C coupler, model 10264-3 (range 0.5 – 1.0 GHz) or 1H 0264-3 (range 0.44 – 0.88 GHz) from Anaren Microwave Inc.

   Compared with the rest of the components such an arrangement is rather expensive.

   If one wishes to construct the couplers oneself it can be done with ‘wire-line’ type BH 10 of Sage. The assembling, however, asks for special tools.

In this report a number of measuring methods are given. It could be considered to apply the swept compression test method with an external wide-band amplifier (no need for linearity) and the external ALC possibility for testing amplifiers asking for more drive power.

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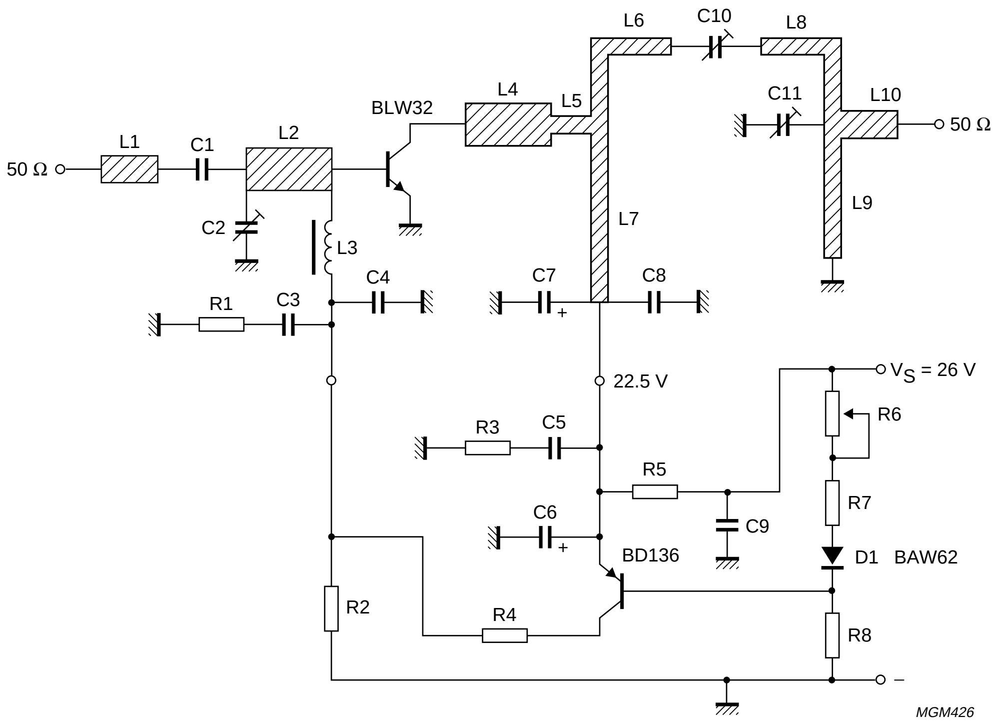

Fig.11 BLW32 circuit.

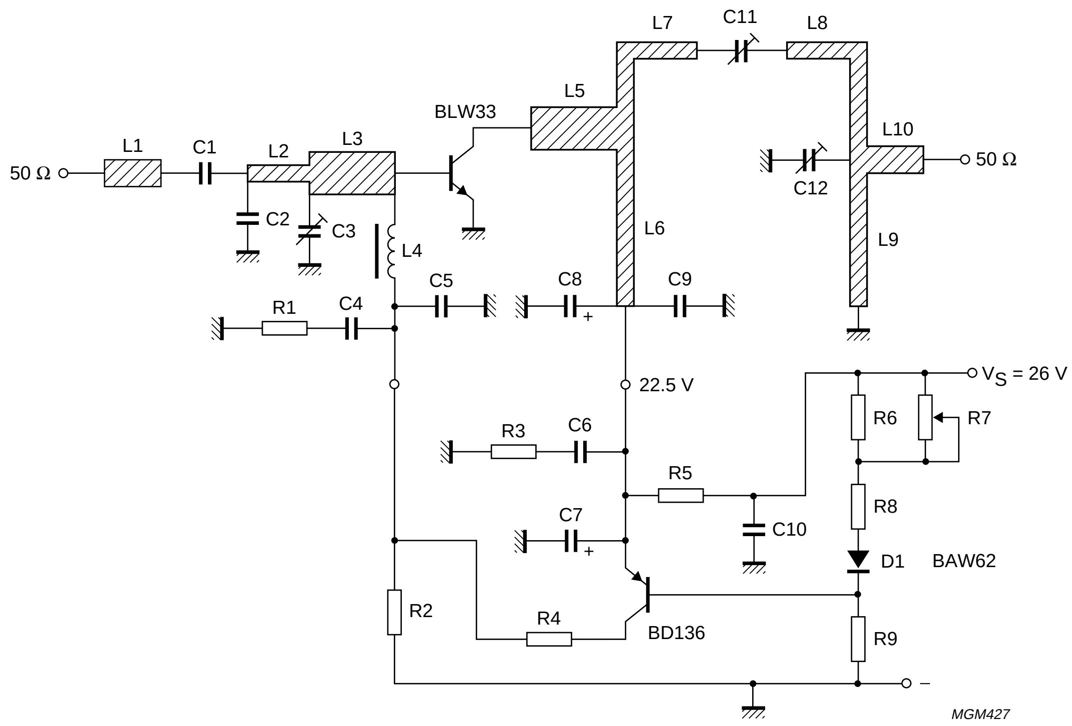

Fig.12 BLW33 circuit.

Table 7 List of components of the BLW32 amplifier (Figs 11 and 13)

C_1	8.2 pF	multilayer ceramic chip capacitor, ATC (Amercian Technical Ceramics) type 100A-8R2-J-Px-50
$C_2 = C_{11}$	1 to 3.5 pF	film dielectric trimmer (cat. no. 2222 809 05001)
$C_3 = C_9$	100 nF	polyester capacitor
$C_4 = C_8$	100 pF	multilayer ceramic chip capacitor (cat. no. 2222 852 13101)
C_5	470 nF	polyester capacitor
$C_6 = C_7$	6.8 µF; 63 V	electrolytic capacitor
C_{10}	1 to 4 pF	micro thin-trim, Tekelec Airtronic part no. AT9401-4-SL1
$L_1 = L_{10}$	stripline (Z_C = 50 Ω)	width 4.0 mm
L_2	stripline (Z_C = 37.6 Ω)	11.5 × 6.0 mm^2
L_3	470 nH	microchoke
L_4	stripline (Z_C = 37.6 Ω)	3.0 × 6.0 mm^2
L_5	stripline (Z_C = 102 Ω)	4.8 × 1.0 mm^2
$L_6 = L_8$	stripline (Z_C = 102 Ω)	20.8 × 1.0 mm^2
L_7	stripline (Z_C = 102 Ω)	19.3 × 1.0 mm^2
L_9	stripline (Z_C = 102 Ω)	23.4 × 1.0 mm^2
$R_1 = R_3$	10 Ω (±5%)	carbon resistor CR25 type
R_2	33 Ω (±5%)	carbon resistor CR25 type
R_4	220 Ω	power metal film resistor PR37 type
R_5	18 Ω (±5%)	power metal film resistor PR52 type
R_6	220 Ω	cermet preset potentiometer
R_7	150 Ω (±5%)	carbon resistor CR25 type
R_8	1.8 kΩ (±5%)	carbon resistor CR25 type
D_1		BAW62

Table 8 List of components of the BLW33 amplifier (Figs 12 and 14)

$C_1 = C_5 = C_9$	100 pF	multilayer ceramic chip capacitor (cat. no. 2222 852 13101)
C_2	2.7 pF	multilayer ceramic chip capacitor, ATC type 100A-2R7-B-Px-50
$C_3 = C_{12}$	1 to 3.5 pF	film dielectric trimmer (cat. no. 2222 809 05001)
$C_4 = C_{10}$	100 nF	polyester capacitor
C_6	470 nF	polyester capacitor
$C_7 = C_8$	6.8 µF; 63 V	electrolytic capacitor
C_{11}	1 to 4 pF	micro thin-trim, Tekelec Airtronic part no. AT 9401-4-SL1
$L_1 = L_{10}$	stripline (Z_C = 50 Ω)	width 4.0 mm
L_2	stripline (Z_C = 102 Ω)	9.6 × 1.0 m^2
L_3	stripline (Z_C = 37.6 Ω)	5.3 × 6.0 m^2
L_4	470 nH	microchoke
L_5	stripline (Z_C = 37.6 Ω)	3.0 × 6.0 m^2
L_6	stripline (Z_C = 102 Ω)	16.2 × 1.0 m^2
$L_7 = L_8$	stripline (Z_C = 102 Ω)	21.7 × 1.0 m^2
L_9	stripline (Z_C = 102 Ω)	20.4 × 1.0 m^2

$R_1 = R_3$	10 Ω (±5%)	carbon resistor CR25 type
R_2	33 Ω (±5%)	carbon resistor CR25 type
R_4	220 Ω (±5%)	power metal film resistor PR37 type
R_5	10 Ω (±5%)	enamelled wire-wound resistor WR0617E style
R_6	1 kΩ (±5%)	carbon resistor CR25 style
R_7	220 Ω	cermet preset potentiometer
R_8	150 Ω (±5%)	carbon resistor CR25 style
R_9	1.8 kΩ (±5%)	carbon resistor CR25 style
D_1		BAW62

Printed-circuit boards 1/16 inch PFTE double CU clad

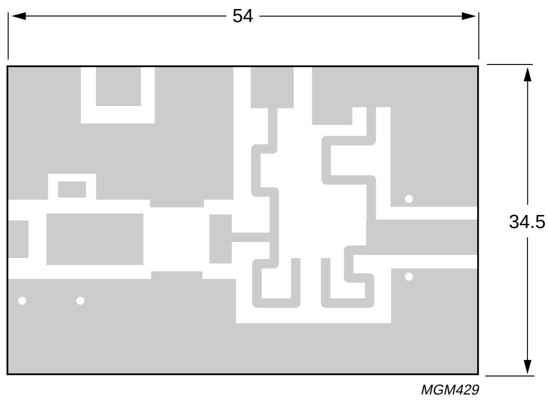

Fig.13 Printed-circuit board BLW32.

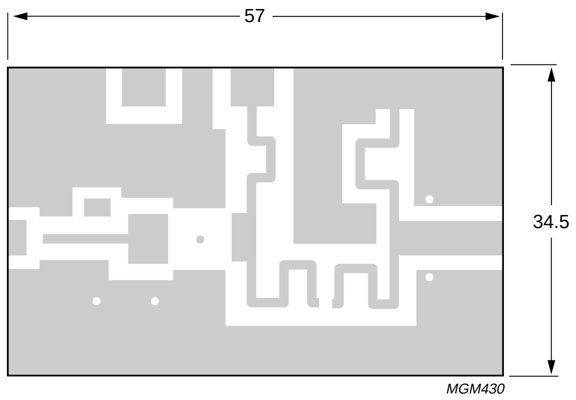

Fig.14 Printed-circuit board BLW33.

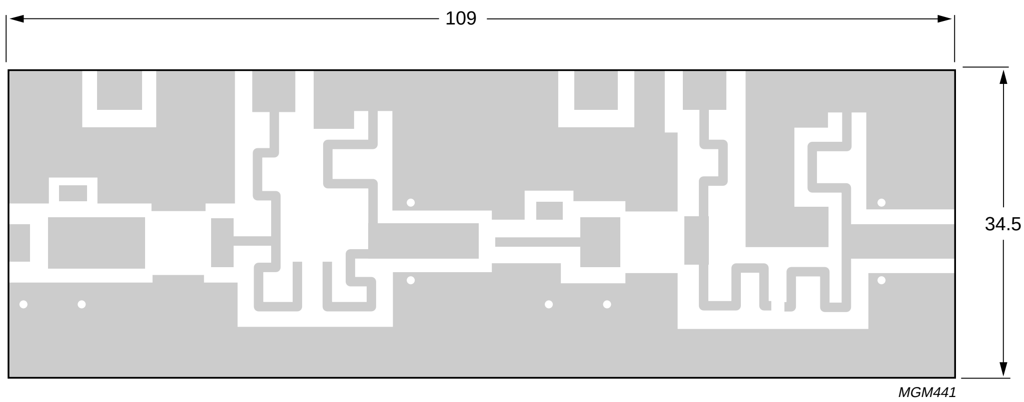

Fig.15 Printed-circuit board BLW32-33.

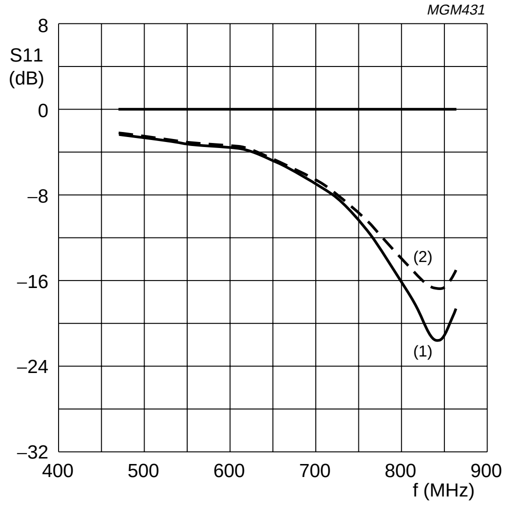

Fig.16 BLW32.

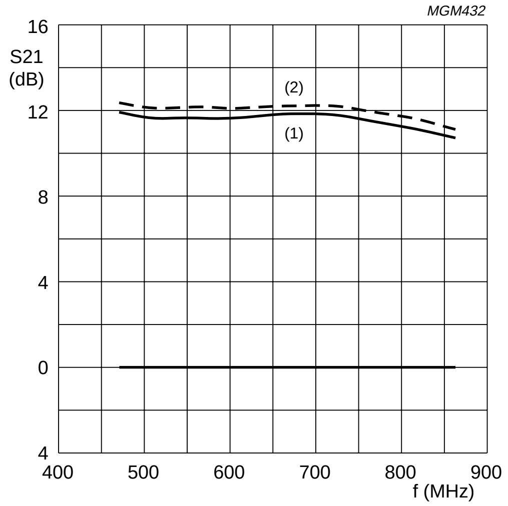

Fig.17 BLW32.

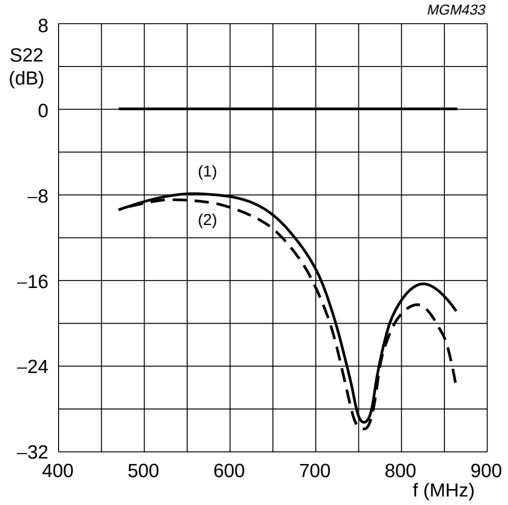

Fig.18 BLW32.

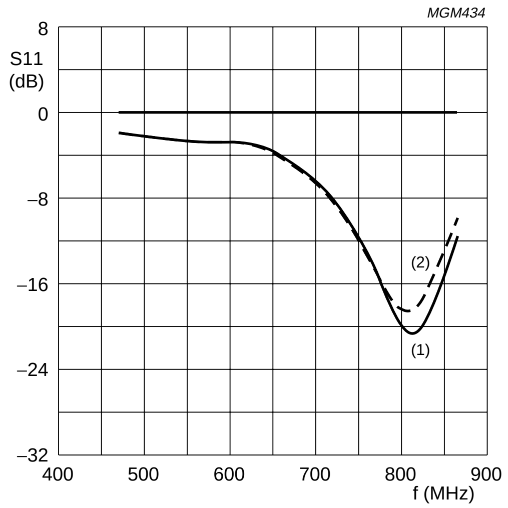

Fig.19 BLW33.

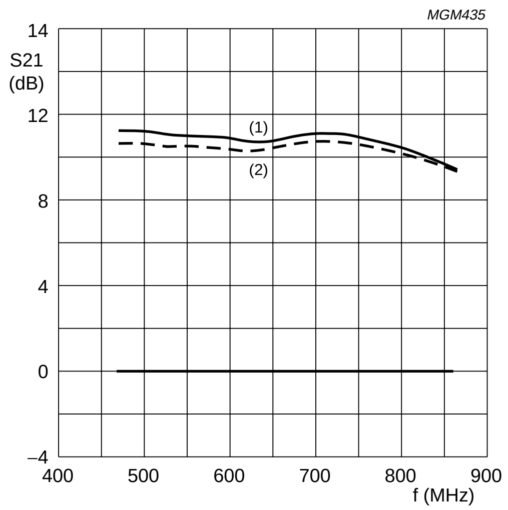

Fig.20 BLW33.

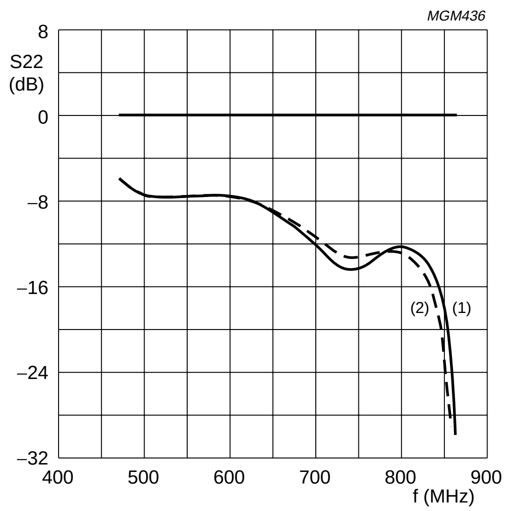

Fig.21 BLW33.

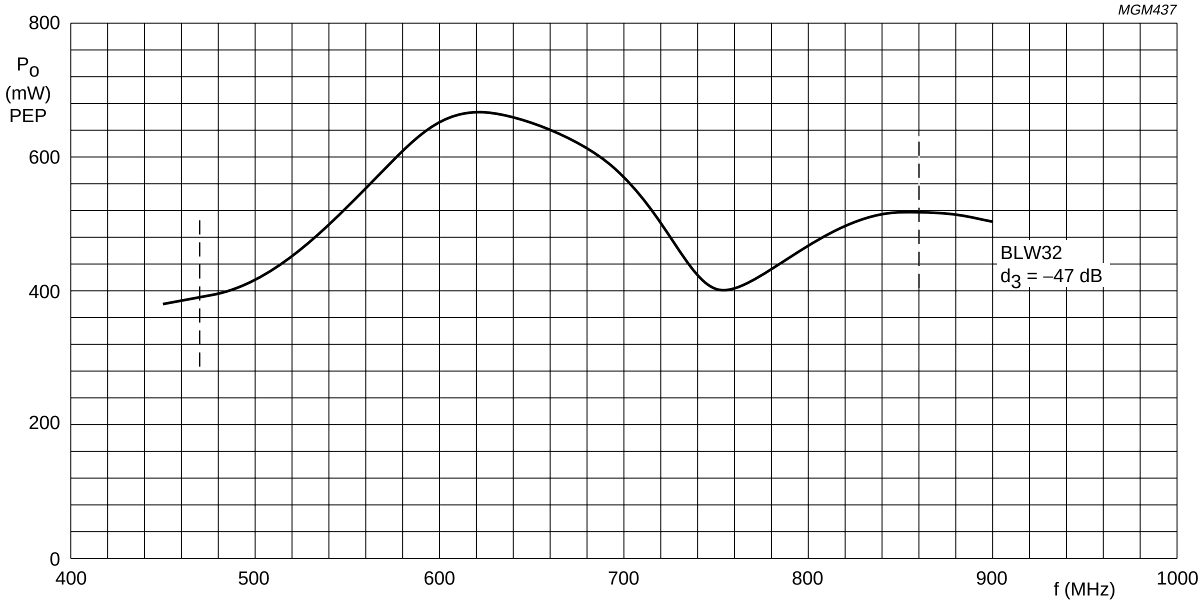

Fig.22 Ampl. no.1.

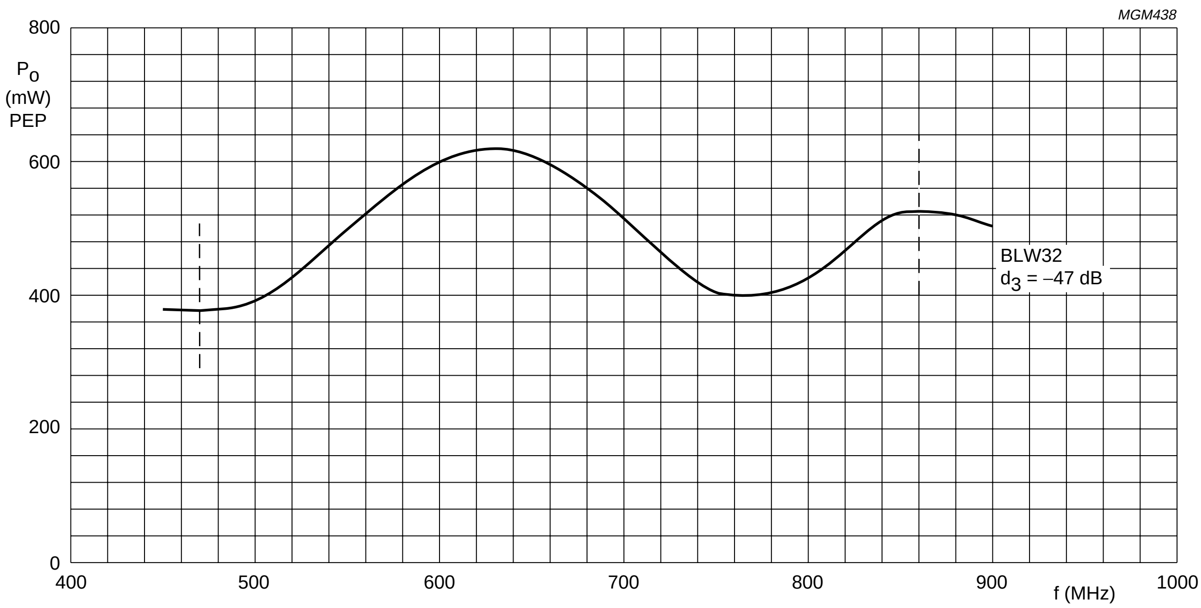

Fig.23 Ampl. no.2.

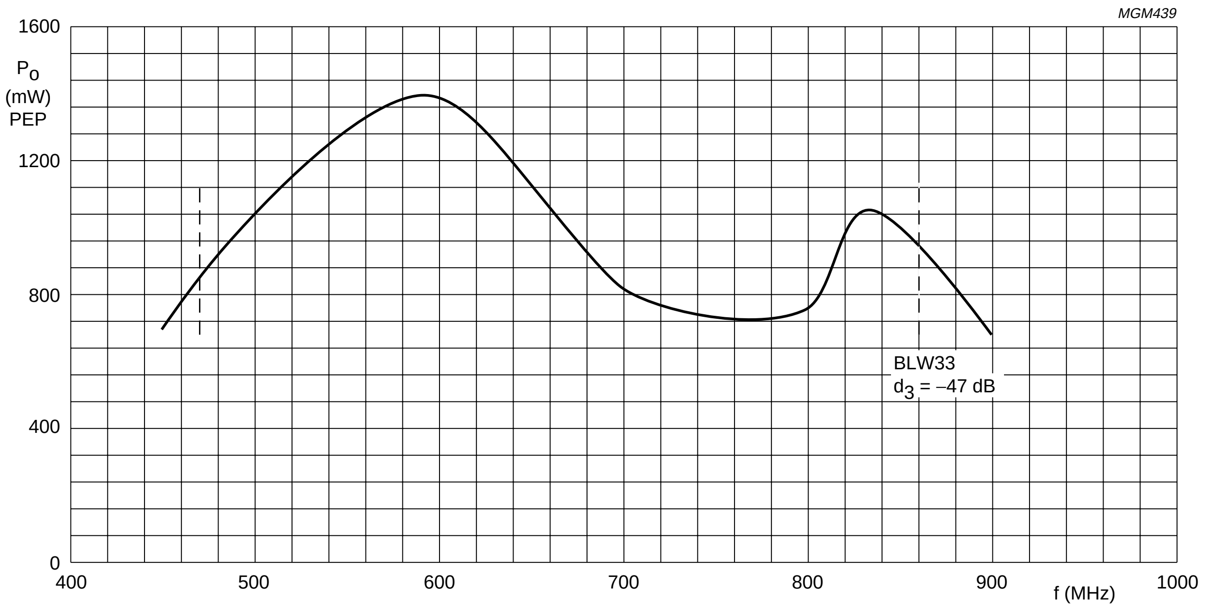

Fig.24 Ampl. no.1.

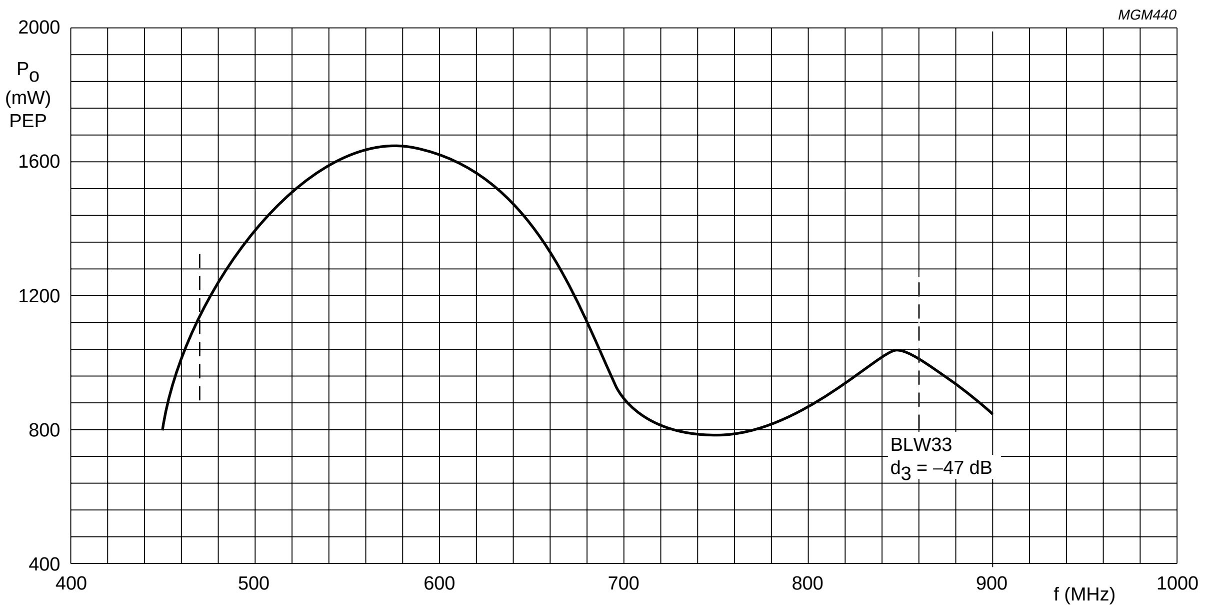

Fig.25 Ampl. no.2.

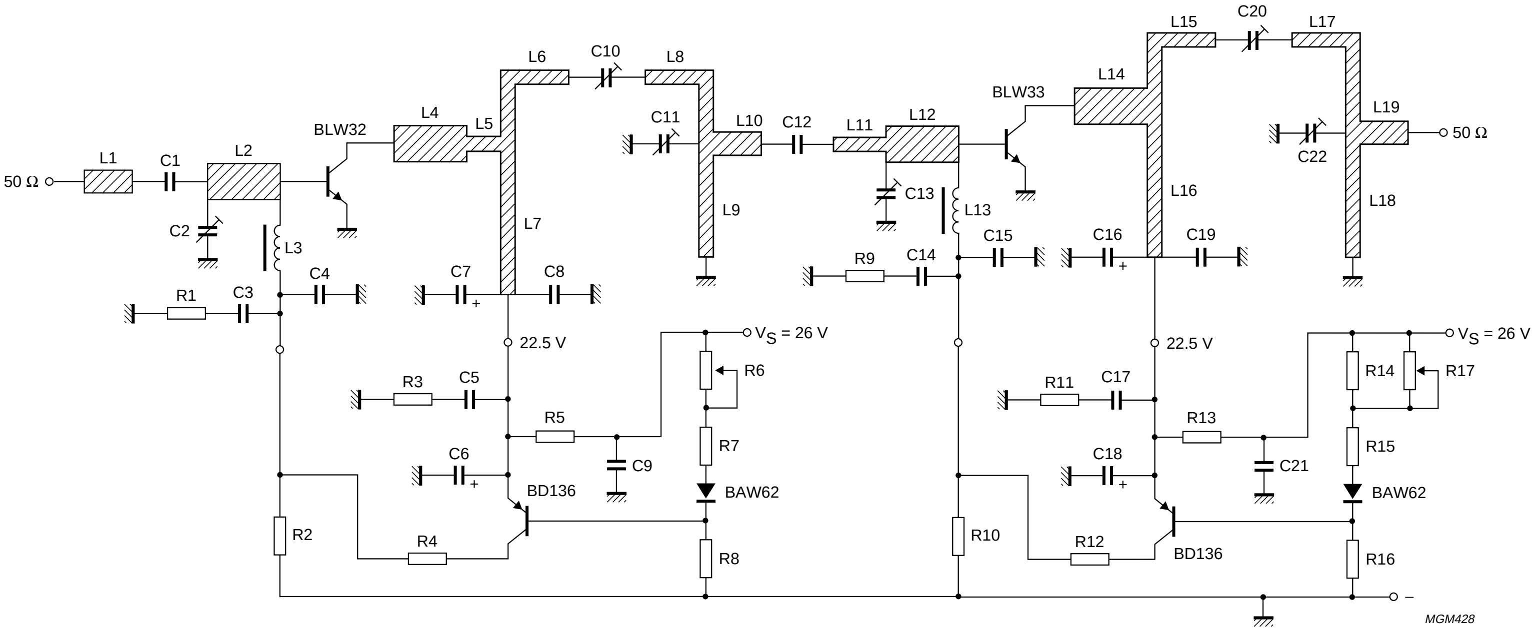

Fig.26 BLW32 and BLW33 circuit.

Table 9 List of components of the BLW32 and BLW33 amplifier (Figs 15 and 26)

$C_1 = C_{12}$	8.2 pF	multilayer ceramic chip capacitor, ATC type 100A-8R2-J-Px-50
$C_2 = C_{11} = C_{13} = C_{22}$	1 to 3.5 pF	film dielectric trimmer (cat.no. 2222 809 05001)
$C_3 = C_9 = C_{14} = C_{21}$	100 nF	polyester capacitor
$C_4 = C_8 = C_{15} = C_{19}$	100 pF	multilayer ceramic chip capacitor (cat.no. 2222 852 13101)
$C_5 = C_{17}$	470 nF	polyester capacitor
$C_6 = C_7 = C_{16} = C_{18}$	6.8 µF	63 V, electrolytic capacitor
$C_{10} = C_{20}$	1 to 4 pF	micro thin-trim, Tekelec Airtronic part no. AT 9401-4-SL1
$L_1 = L_{10} = L_{19}$	stripline (Z_C = 50 Ω)	width 4.0 mm
L_2	stripline (Z_C = 37.6 Ω)	11.5 × 6.0 mm^2
$L_3 = L_{13}$	470 nH	microshoke
$L_4 = L_{14}$	stripline (Z_C = 37.6 Ω)	3.0 × 6.0 mm^2
L_5	stripline (Z_C = 102 Ω)	4.8 × 1.0 mm^2
$L_6 = L_8$	stripline (Z_C = 102 Ω)	20.8 × 1.0 mm^2
L_7	stripline (Z_C = 102 Ω)	19.3 × 1.0 mm^2
L_9	stripline (Z_C = 102 Ω)	23.4 × 1.0 mm^2
L_{11}	stripline (Z_C = 102 Ω)	9.6 × 1.0 mm^2
L_{12}	stripline (Z_C = 37.6 Ω)	5.3 × 6.0 mm^2
$L_{15} = L_{17}$	stripline (Z_C = 102 Ω)	21.7 × 1.0 mm^2
L_{16}	stripline (Z_C = 102 Ω)	16.2 × 1.0 mm^2
L_{18}	stripline (Z_C = 102 Ω)	20.4 × 1.0 mm^2
$R_1 = R_3 = R_9 = R_{11}$	10 Ω (±5%)	carbon resistor CR25 type
$R_2 = R_{10}$	33 Ω (±5%)	carbon resistor CR25 type
$R_4 = R_{12}$	220 Ω (±5%)	power metal film resistor PR37 type
R_5	18 Ω (±5%)	power metal film resistor PR52 type
$R_6 = R_{17}$	220 Ω	cermet preset potentiometer
$R_7 = R_{15}$	150 Ω (±5%)	carbon resistor CR25 style
$R_8 = R_{16}$	1.8 kΩ (±5%)	carbon resistor CR25 style
R_{13}	10 Ω (±5%)	enamelled wire-wound resistor WR 0617E style
R_{14}	1 kΩ (±10%)	carbon resistor CR25 type
$D_1 = D_2$		BAW62

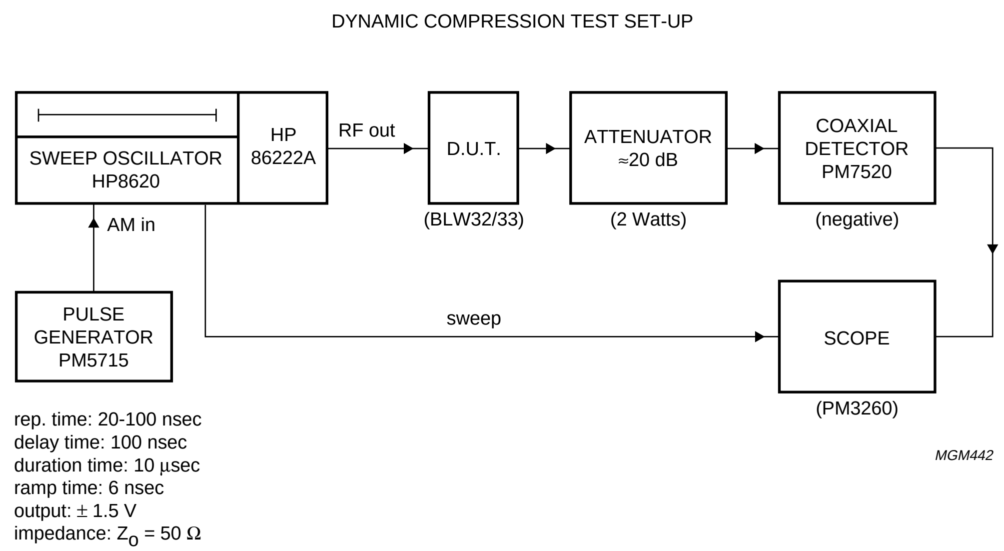

Fig.27 Dynamic compression test set-up.

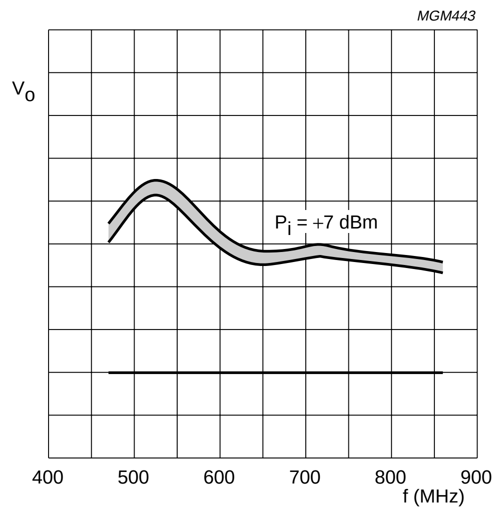

Fig.28 Compression.

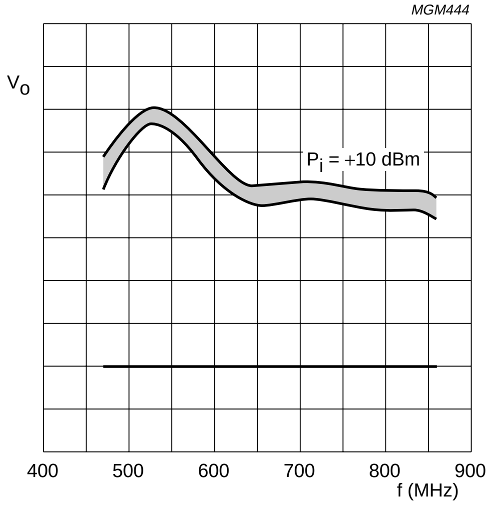

Fig.29 Compression.

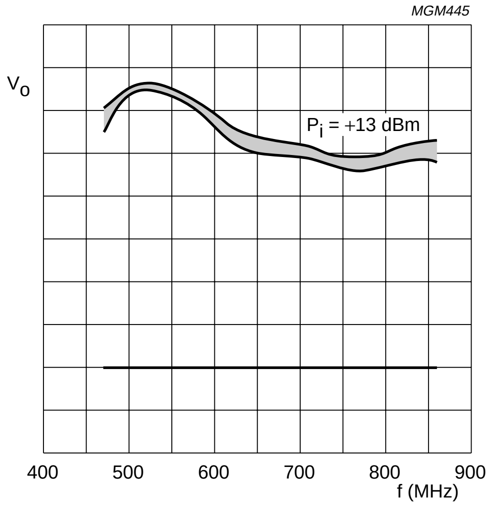

Fig.30 Compression.

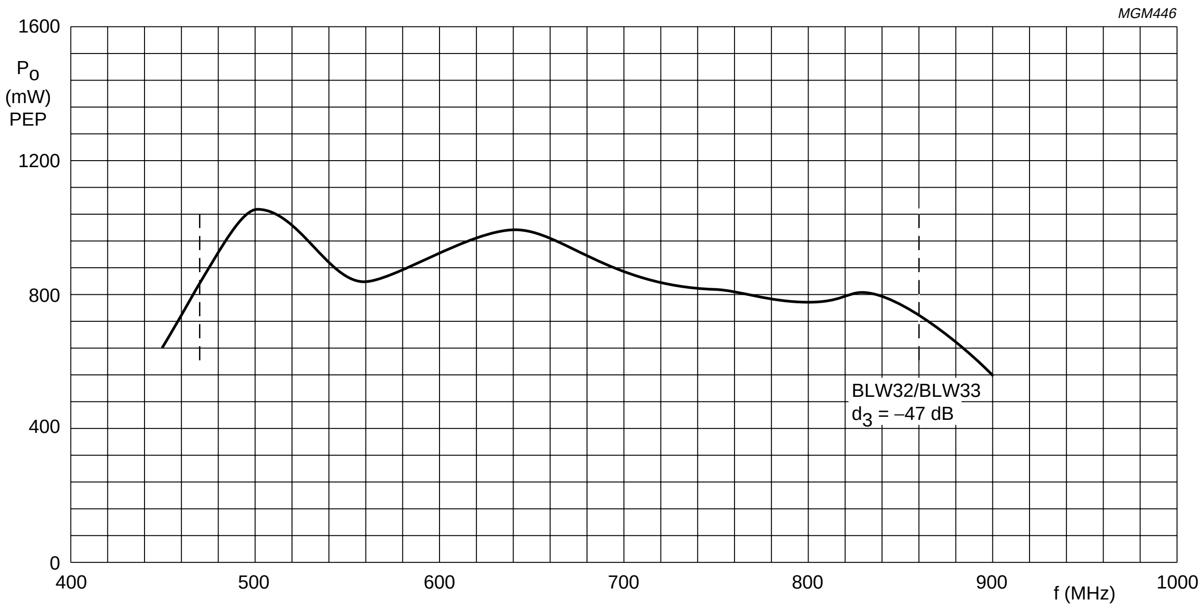

Fig.31 Ampl. no.1.

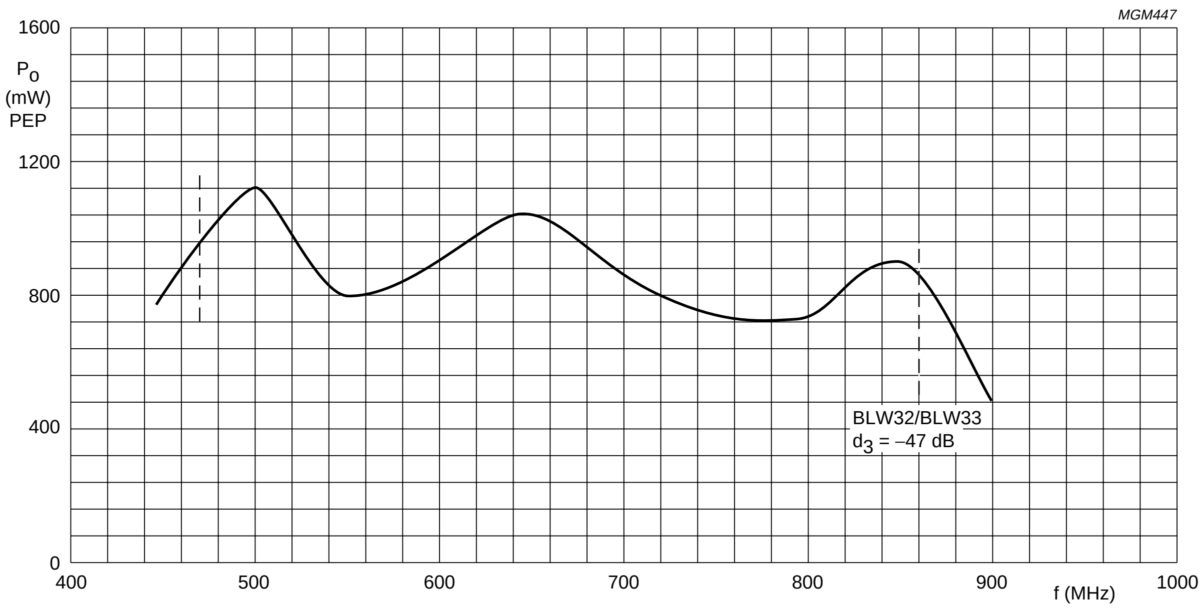

Fig.32 Ampl. no.2.

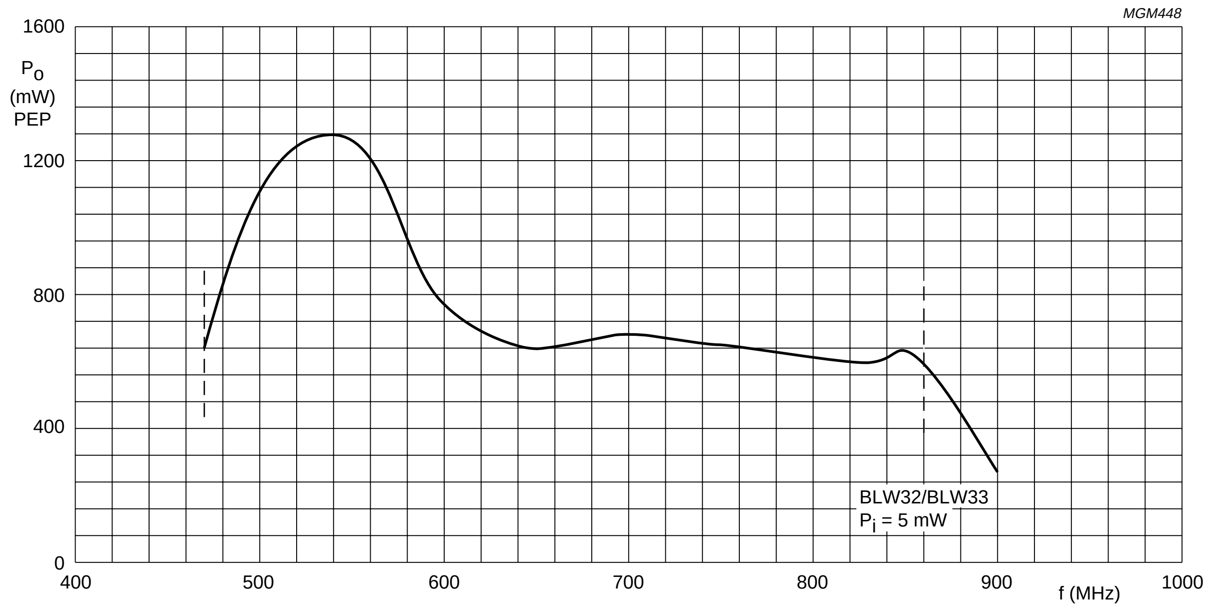

Fig.33 Ampl. no.1.

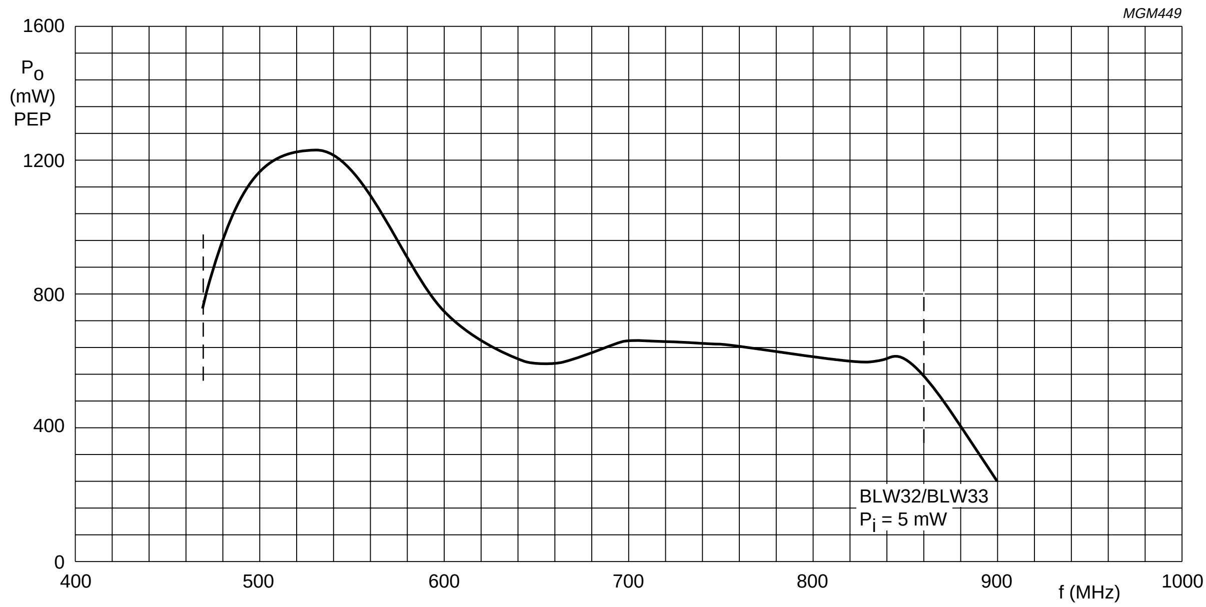

Fig.34 Ampl. no.2.

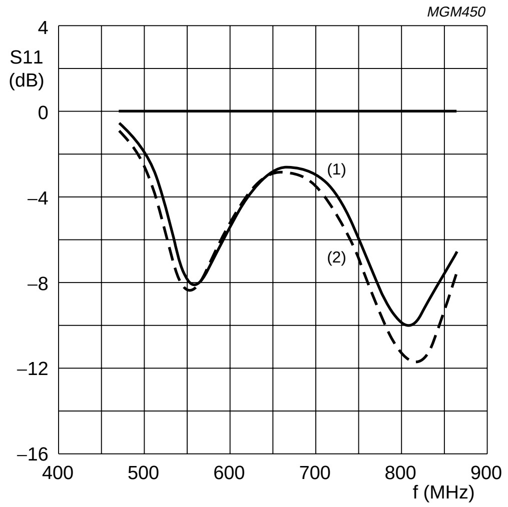

Fig.35 BLW32 and BLW33.

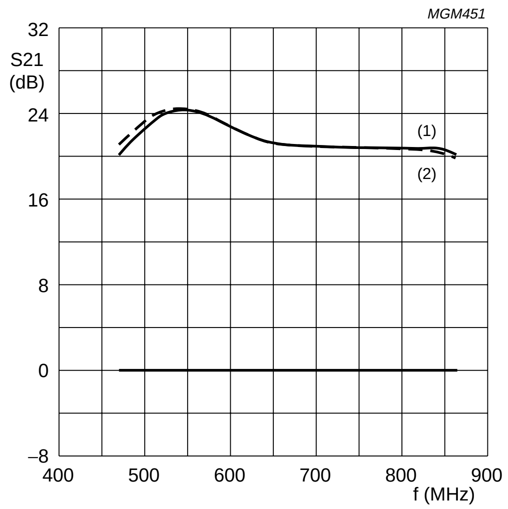

Fig.36 BLW32 and BLW33.

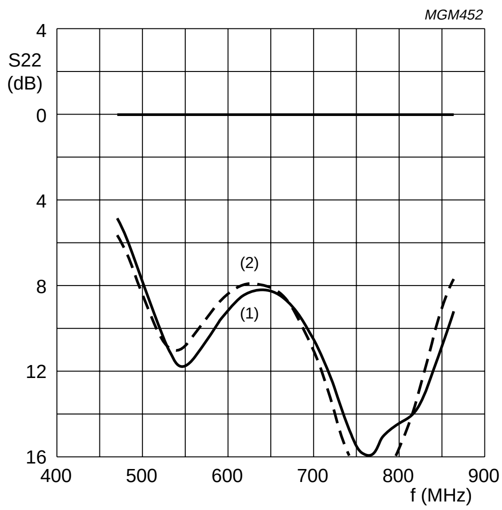

Fig.37 BLW32 and BLW33.

Table 10 BLW32

CHANNEL NO.	I.M.D. (dB)	P_O SYNC. (mW)		GAIN (dB)	
		1[1]	**2**	**1**	**2**
21	–60	462	427	12.0	12.3
21	–56	605	561		
21	–52	771	716		
39	–60	699	643	11.7	11.15
39	–56	965	937		
39	–52	1245	1189		
70	–60	487	512	10.8	11.3
70	–56	713	769		
70	–52	1007	1021		

Note

1. Corresponding amplifier number.

Table 11 BLW33

CHANNEL NO.	I.M.D. (dB)	P_O SYNC. (mW)		GAIN (dB)	
		1	2	1	2
21	–60	1132	1440	10.7	10.3
21	–56	1775	1916		
21	–52	2263	2246		
39	–60	1161	1454	10.6	10.25
39	–56	1762	1902		
39	–52	2293	2321		
70	–60	819	1063	9.5	9.3
70	–56	1342	1580		
70	–52	2014	2265		

Table 12 BLW32 and BLW33

CHANNEL NO.	I.M.D. (dB)	P_O SYNC. (mW)	GAIN (dB)
		2	2
21	–60	1045	21.65
21	–56	1706	–
21	–52	2311	–
39	–60	895	21.40
39	–56	1440	–
39	–52	2014	–
70	–60	587	19.75
70	–56	923	–
70	–52	1384	–

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Printed in The Netherlands

Date of release: 1998 mar 23

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