

APPLICATION NOTE

A wide-band linear power amplifier (470 – 860 MHz) with two transistors BLW34

ECO7901

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1 ABSTRACT

For application in driver or final stages of TV-transposers in band IV/V (470-860 MHz) a linear wideband power amplifier has been designed with 2 transistors BLW34 coupled by means of 3 dB –90° hybrids. Each transistor is adjusted in class-A at $V_{CE} = 25\text{ V}$ and $I_c = 0.6\text{ A}$. The peak sync output power for a 3-tone I.M. distortion of –60dB varies between 3.6 and 5.4 W. The power gain is $9.1 \pm 0.3\text{ dB}$. Input and output VSWR are below 1.3.

2 INTRODUCTION

This report describes the realisation of a wide-band UHF power amplifier for TV transposer service in band IV and V (470 – 860 MHz).

The amplifier is designed with the BLW34 transistor being developed for ultra linear applications operating in class A. Each device is able to deliver at least 1.8 W peak sync output. The BLW34 forms a series with the smaller devices BLW33 (1.0 W) and BLW32 (0.5 W).

The power gain at 860 MHz is at least 9 dB.
The BLW34 is encapsulated in a 1/4 inch capstan envelope with ceramic cap.

3 THEORETICAL CONSIDERATIONS

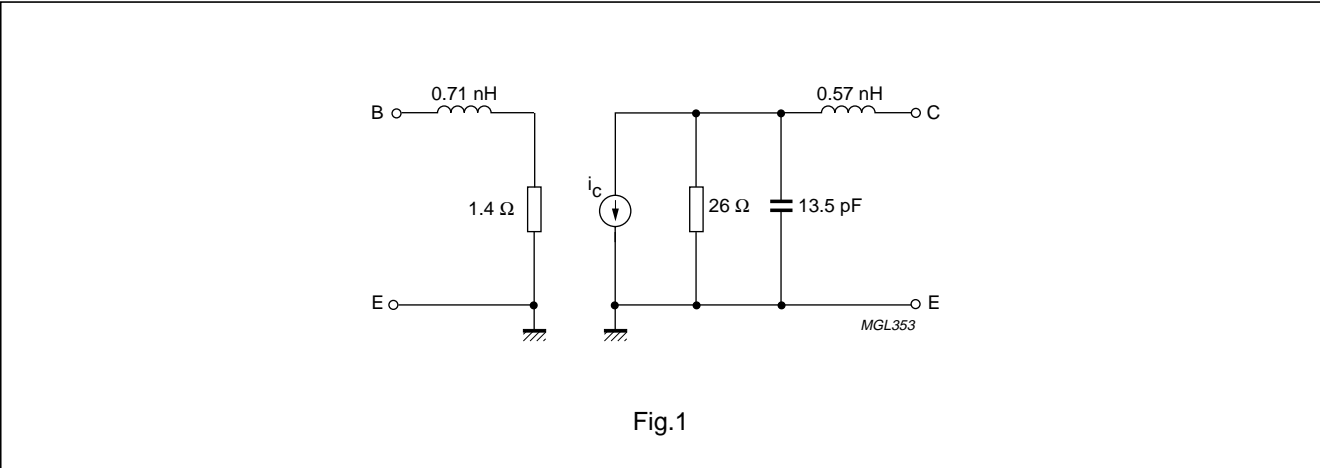
3.1 The equivalent circuit of the BLW34

For class A operation the BLW34 is specified at $V_{CE} = 25\text{ V}$; $I_c = 600\text{ mA}$.
The corresponding typical gain, input and load impedance according to the Data sheets are given in Table 1.

Table 1

f (MHz)	GAIN (dB)	R _i (SERIES) (Ω)	X _i (SERIES) (Ω)	R _L (SERIES) (Ω)	X _L (SERIES) (Ω)
470	15.2	1.46	1.93	12.8	11.0
636	12.6	1.39	2.84	8.85	9.97
860	10.1	1.27	4.00	5.36	7.67

To facilitate calculations an approximate equivalent circuit for the transistor input and output impedance can be given. It is shown in Fig.1.



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3.1.1 THE OUTPUT NETWORK

The circuit will be designed on printed circuit board with PTFE fibre glass as a dielectric having an $\epsilon_r = 2.74$ and thickness of 1/32 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 1/32 inch the characteristic resistance is 21 Ω . The length for the collector lead amounts to 3 mm, but the base lead is different in length.

The first step in the matching is to tune out the output capacitance of the transistor by means of the collector R.F. choke.

This choke is executed as a stripline with a width of 1 mm corresponding with a characteristic resistance of 72 Ω to keep the parallel capacitance at this point as low as possible.

For practical reasons the choke is connected to the main transmission line at a distance of 3 mm from the transistor edge. For the design procedure one is referred to Part 1 of this handbook (SC19B).

The results of the calculations before and after computer optimization are given in Fig.2 and Table 2.

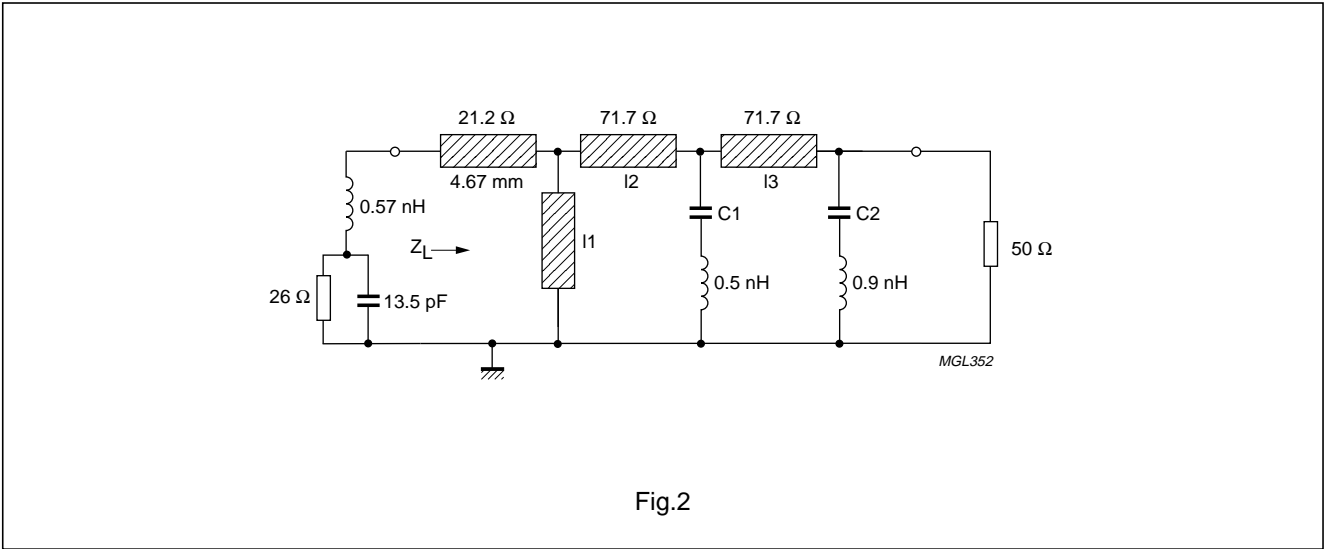


Table 2

ELEMENT	BEFORE OPTIM.	AFTER OPTIM.	UNIT
l_1	16.8	22.1	mm
l_2	18.2	20.8	mm
C_1	8.58	9.76	pF
l_3	38.2	43.5	mm
C_2	3.91	3.09	pF
$S_{max.}$	2.48	1.23	—

S_{max} is the maximum VSWR of the network.

The lenghts given hold for air lines. The actual lengths on the printed-circuit board are shorter; the reduction factors are: 1.445 for the 72 Ω lines and 1.556 for the 21 Ω line.

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The predicted minimum output power of the complete amplifier with two transistors BLW34 is:

$$Po_2 = \frac{2Po_1}{S_{max}} \cdot 0.95 = \frac{2 \cdot 1.8}{1.23} \cdot 0.95 \cong 2.8W$$

Po₁ is the minimum output power of one transistor in a narrow band circuit, S_{max} is as specified above and the factor 0.95 represents the power loss of the hybrid coupler at the output.

3.1.2 THE INPUT NETWORK

The design procedure can be found in Part 1 of this handbook (SC19B).

The results of the calculations are summarized in Fig.3 and Table 3.

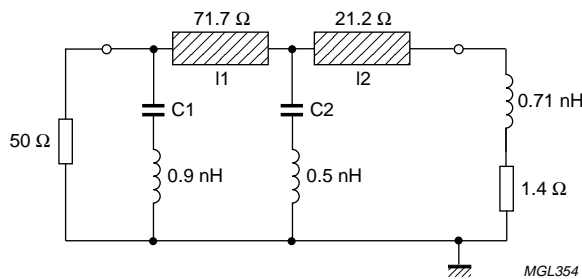


Fig.3

Table 3

ELEMENT	BEFORE OPTIM.	AFTER OPTIM.	UNIT
C1	4.78	4.41	pF
l1	25.9	32.1	mm
C2	21.9	19.9	pF
l2	13.3	10.5	mm
ΔG	±1.93	±0.12	dB

ΔG is the resulting power gain variation caused by the transistor and the input network over the frequency band. The lengths of the lines hold for air as a dielectric. Transformation to striplines on a printed-circuit board is done in the same way as in the previous section.

The minimum power gain of the complete amplifier with two transistors BLW34 is expected to be:

$$G_0 = G_t - 2A_H - A_1 - A_2$$

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in which: G_t = minimum power gain of BLW34 in a narrow band circuit.

A_H = power loss of one hybrid coupler

A_1 = reflection loss of input network

A_2 = reflection loss of output network.

In practical figures this means:

$$G_o = 9.0 - 2 \times 0.2 - 0.15 - 0.05 = 8.4 \text{ dB.}$$

The input VSWR of a single amplifier was calculated to vary from 11 at 470 MHz down to 1.45 at 860 MHz.

4 THE HYBRID COUPLED AMPLIFIER

4.1 Practical considerations

On previous pages the theoretical approach of a single amplifier has been discussed.

In practice, it was the intention to realize a small compact amplifier on a printed-circuit board with the input and output terminal ($R_c = 50 \Omega$) in line for easy cascading of several amplifiers.

Besides the wide-band properties it is the intention to obtain a higher output power, so two BLW34 branches are connected in parallel.

At the same time it is of course rather unacceptable that the amplifier loads a driver stage with a mismatch causing a VSWR of 11 at 470 MHz.

Both problems may be solved sufficiently by applying two BLW34 branches in parallel with the aid of two wide-band 3 dB -90° C. coaxial hybrids on a 50Ω basis.

In this configuration the output power will be nearly doubled; reduction of the input VSWR of the complete system to a value of around 1.2 may be explained from the properties of the coupler applied. The reflected power will be absorbed in the resistor matching the isolated port. This resistor is 50Ω and consists of two 100Ω power metal film resistor in parallel. The same has been done on the output side.

The printed-circuit board needs to be double copper clad with a PTFE fibre glass dielectric ($\epsilon_r = 2.74$) for low losses at UHF.

A thickness of 1/32 inch has been chosen; so the 72Ω lines are 1 mm wide.

Figure 4 shows the circuit diagram of the complete $2 \times$ BLW34 class A amplifier. The biasing circuit is drawn in Fig.5.

The printed-circuit board and the amplifier lay-out in Fig.6.

For a correct earthing the upper earth sheet parts are connected to the lower sheet by soldering copper straps at the edges of the printed-circuit board. The black parts in Fig.6 are the soldered copper straps.

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

Both transistors are screwed to a water-cooled brass heatsink. So, several heatsink temperatures can be applied by means of a TAMSON unit supplying water with a thermostatically controlled temperature.

The tuning capacitors in the circuit are of the film dielectric type with three tags of which both earth terminals are fed through small holes and soldered to the upper as well as the lower plane. Fixed capacitors in the r.f. path are of the multi-layer ceramic chip type.

The coaxial connectors are of the SMA 50Ω type, being soldered to upper and lower sheet.

4.2 Practical optimization

We started with optimization on a small signal basis with the circuit inserted in a network analyzer chain, having swept S-parameter facilities.

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So far, similar tuning methods have been applied as described in Ref. 2.

Because it is rather complicated to find the best compromise between an acceptable flat gain curve (S_{21}) and sufficient output power with low i.m.d. a dynamic large signal optimization method has been realized. Figure 7 shows the block diagram.

This tuning method is based on the correlation between the single tone 1dB compression point and the i.m.d figure of a linear amplifier.

In this set-up the swept output power level of the amplifier under test is kept constant and the required (detected) drive power monitored on an oscilloscope screen (PM3260).

The swept drive power is available from the sweep generator HP8620C in combination with RF plug-in unit HP86222A. Because the output power of this system is too low viz. appr. 20 mW (+13 dBm) a combined amplifier with BLW32 and BLW33 (Ref. 2) has been added. The latter combination shows an overall gain of approx. 20 db in the range 470 – 860 MHz. When the circuit under test is inserted in the chain of Fig.7, the input power measured on port C of hybrid 1 corresponds in principle with the gain curve (S_{21}) being measured with the network analyzer; in fact one is the inverse of the other.

When the drive level is slowly increased, the shape of the gain curve changes somewhat when compression starts. By careful retuning of the amplifier the shape of the gain curve can be corrected again in the direction of the original smaller signal curve.

The actual single tone output power has been measured with the aid of the calorimetric watt meter HP435A when the action of the sweep is stopped. Also it is possible to examine the output signal itself by means of a spectrum analyzer being loosely coupled via a 50 Ω pick-off device.

Resuming it can be said that the advantage of applying this high level tuning system is characterized by the fact that the output power is leveled and so compression does not start earlier due to gain fluctuations of the amplifier. This makes the judgement of the compression level easier.

4.3 Intermodulation, VSWR and gain measurements

For i.m.d. measurements on television systems the post offices advice and apply the 3-tone test method (vision carrier –8 dB, sound carrier –7 dB, sideband signal –16 dB; zero dB corresponds to peak sync level).

For this reason a wideband test set-up has been realized. The block diagram is in Fig.8. In this set-up first the sound and vision carriers are joined in the wide-band coaxial hybrid H_1 . Then the sideband signal from the (smaller) generator SMLU is added in hybrid H_2 .

The complete 3-tone signal passes a low-pass filter (700 or 1000 MHz cut-off depending on input frequency) a continuously variable attenuator and a circulator (three different types needed to cover at least the range 470 – 860 MHz).

The output power is measured with a HP435A calorimetric watt meter and the i.m.d. observed with a spectrum analyzer (HP8558B).

Figure 9 shows the 3-tone i.m.d. results measured on the complete hybrid coupled $2 \times$ BLW34 amplifier. To get an idea of the P_{o_sync} drop for different heatsink temperatures, measurements have been done for $T_{HS} = 23^\circ\text{C}$ and $T_{HS} = 70^\circ\text{C}$. The maximum drop in P_{o_sync} amounts to 1.1 dB.

Besides the 3-tone test the peak envelope power has been measured in a two-tone way for a third order i.m.d of $d_3 = -47$ dB.

It can be proved that there is a correlation between the afore described –60 dB 3-tone and the –47 dB two-tone test method.

It is known that the advantage of video precorrection on i.m.d. amounts to values up to 10 dB. Calculating with an average of 8 dB, the two-tone i.m.d. for $d_3 = 39$ dB is interesting. From Fig.10 it may be seen that the output power is more than doubled. Both tests have been done for heatsink temperatures of resp. 23 and 70 $^\circ\text{C}$.

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Finally input and output VSWR and gain figures were measured only under small signal conditions. The results of amplifier branches 1 and 2 for a heatsink temperature of 70 °C are shown in Figs 11 to 13. The VSWR figures of the input and output are expressed in reflection damping (resp. S_{11} and S_{22}) on a 50 Ω basis.

According to Fig.11 the minimum reflection damping of the input amounts to $S_{11} = -1$ dB (VSWR = 17.4) at 470 MHz and to $S_{11} = -10$ dB (VSWR = 1.93) at 860 MHz.

On the output side the worst reflection damping amounts to $S_{22} = -5.5$ dB (VSWR = 3.26) in the middle of the passband (Fig.13).

The power gain, represented by S_{21} , is appr. 9.5 dB (Fig.12).

Resuming it can be said that, as Figs 11 to 13 show, the differences in gain, input and output VSWR between both branches are rather small.

Figs 14 to 16 show the final results of the complete hybrid coupled amplifier. Examining the S_{11} and S_{22} figures it appears that the minimum input reflection damping (S_{11}) amounts to appr. 18 dB, what corresponds with a maximum VSWR of 1.29.

The latter value may be mainly explained from the specified maximum VSWR = 1.25 of the applied Anaren hybrids. If more attention is paid to matching of the isolated ports the results may be improved. The matching consists of two metal film power transistors of 100 Ω in parallel. They have low-frequency tolerances of 5%.

The small signal gain of the complete amplifier amounts to 9.1 ± 0.3 dB for $T_{HS} = 23$ °C, whilst the dashed line shows the results for $T_{HS} = 70$ °C being 8.9 ± 0.3 dB.

The temperature influence on the S_{11} and S_{22} figures is almost negligible.

5 CONCLUSIONS

On preceding pages the theoretical and practical design has been described of the wide-band (470 – 860 MHz) high quality linear amplifier being equipped with two BLW34 transistors operating in class A.

There are some small differences between the theoretical design and the practical circuit.

- The calculated value for the chip capacitors $C_{11} = C_{12} = C_{13} = C_{14}$ in Fig.4 was 10 pF. In practice 8.2 pF appeared to be a better choice.
- Also the values of $C_{19} = C_{20} = C_{21} = C_{22}$, being calculated as 4.7 pF are changed. The new value is 3.9 pF.
- In the first instance a chip capacitor of 1.5 pF in parallel with C_{23} and C_{26} was planned. However this capacitor could be omitted in the operational circuit.

The expected results for a single amplifier: $G_{p \text{ min.}} = 8.8$ dB; and $P_{o \text{ sync}} = 1.5$ W at a 3-tone i.m.d. of -60 dB and $G_{p \text{ min}} = 8.4$ dB and $P_{o \text{ sync}} = 2.8$ W for $T_{HS} \leq 70$ °C are realized in this design.

Here, we have calculated with a total insertion loss of 0.4 dB for the applied hybrids. They are specified for a maximum of 0.25 dB per device (1.06 \times power).

6 REFERENCES

Ref. 1: O. Pitzalis Jr. and R.A. Gilson - Tables of impedance matching networks which approximate prescribed attenuation versus frequency slopes. IEEE Transactions on microwave theory and techniques, Vol. MTT-19, no. 4, April 1971, pp. 381-386.

Ref. 2: A.H. Hilbers and M.J. Köppen - Wide-band linear power amplifier (470 – 860 MHz) with the transistors BLW32 and BLW33. C.A.B. report ECO7806.

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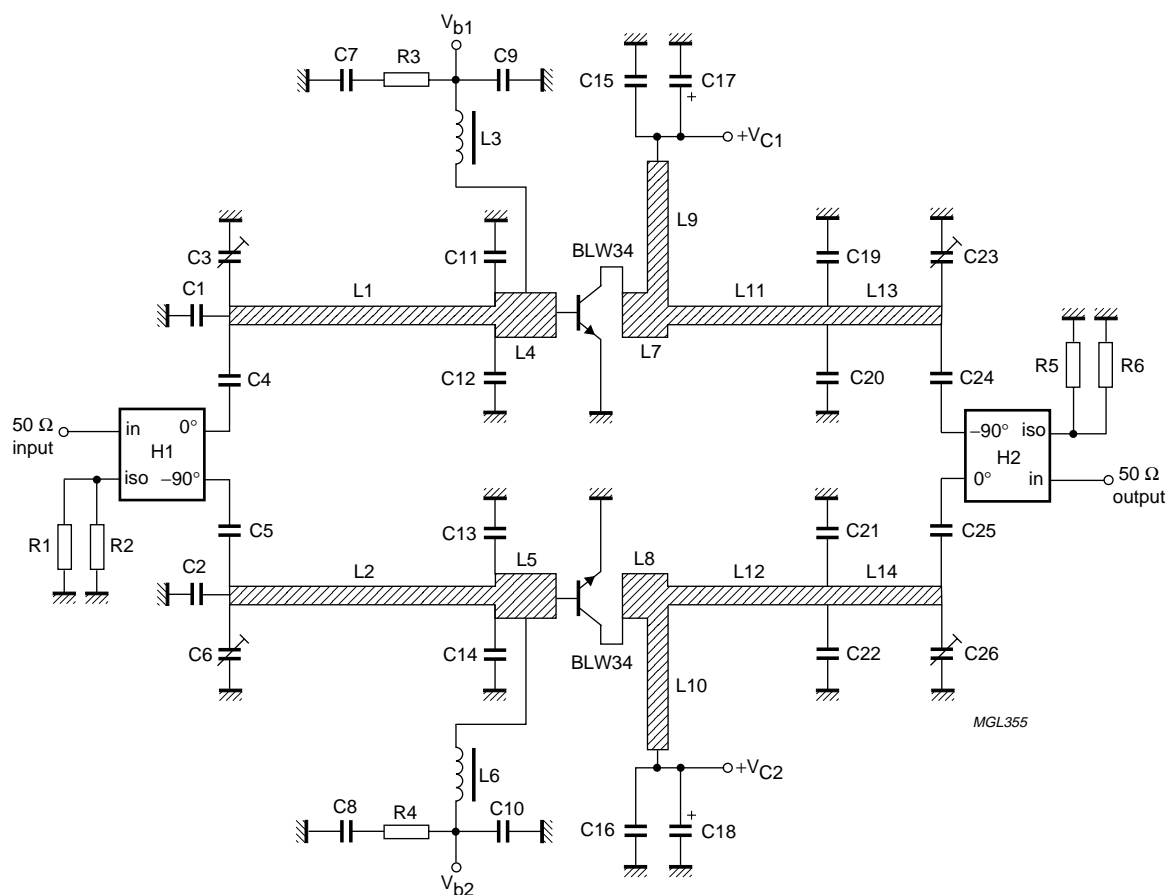


Fig.4

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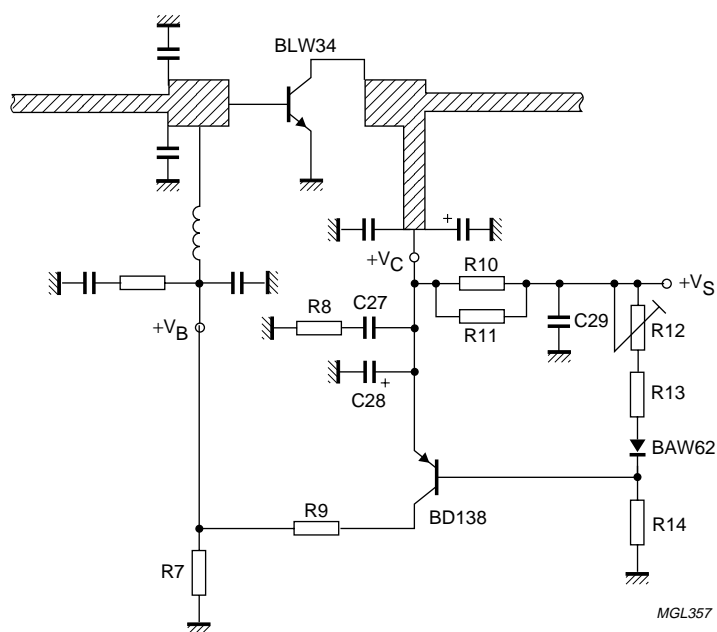


Fig.5

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Table 4

LIST OF COMPONENTS	
$C_1 = C_2 = 2.2 \text{ pF}$	multilayer ceramic chip capacitor Tekelec-Airtronic part no. 500 R15 N2R2 BA
$C_3 = C_6 = C_{23} = C_{26} = 1.4 \text{ to } 5.5 \text{ pF}$	film dielectric trimmer (cat. no. 2222 809 09001)
$C_4 = C_5 = C_{24} = C_{25} = 100 \text{ pF}$	multilayer ceramic chip capacitor (cat. no. 2222 851 13101)
$C_7 = C_8 = C_{29} = 100 \text{ nF}$	polyester capacitor
$C_9 = C_{10} = C_{15} = C_{16} = 100 \text{ pF}$	multilayer ceramic chip capacitor (cat. no. 2222 852 13101)
$C_{11} = C_{12} = C_{13} = C_{14} = 8.2 \text{ pF}$	multilayer ceramic chip capacitor, ATC (American Technical Ceramics) type 100A-8R2-J-Px-50
$C_{17} = C_{18} = C_{28} = 6.8 \text{ }\mu\text{F}, 63 \text{ V}$	electrolytic capacitor
$C_{19} = C_{20} = C_{21} = C_{22} = 3.9 \text{ pF}$	multilayer ceramic chip capacitor, Tekelec-Airtronic part no. 500 R15 N3R9 CA
$C_{27} = 470 \text{ nF}$	polyester capacitor
$R_1 = R_2 = R_5 = R_6 = 100 \text{ }\Omega (\pm 5\%)$	power metal film resistor PR37 type (cat. no. 2322 191 31001)
$R_3 = R_4 = R_8 = 10 \text{ }\Omega (\pm 5\%)$	carbon resistor; CR25 type
$R_7 = 33 \text{ }\Omega (\pm 5\%)$	carbon resistor; CR25 type
$R_9 = 220 \text{ }\Omega (\pm 5\%)$	power metal film resistor PR52 type (cat. no. 2322 192 32201)
$R_{10} = 5.6 \text{ }\Omega (\pm 5\%)$	enamelled wire-wound resistor WR0617E style
$R_{11} = 8.2 \text{ }\Omega (\pm 5\%)$	enamelled wire-wound resistor WR0617E style
$R_{12} = 100 \text{ }\Omega$	cermet preset potentiometer
$R_{13} = 120 \text{ }\Omega (\pm 5\%)$	carbon resistor; CR25 type
$R_{14} = 1.8 \text{ k}\Omega (\pm 5\%)$	carbon resistor; CR25 type
$H_1 = H_2 = \text{ultra-miniature } 3 \text{ dB } -90^\circ \text{ coupler model no. 10264-3, range } 0.5 - 1.0 \text{ GHz; Anaren Microwave Inc.}$	
$L_1 = L_2$	stripline ($Z_c = 72 \text{ }\Omega$), $22.1 \times 1.0 \text{ mm}^2$; note 1
$L_3 = L_6 = 1 \text{ }\mu\text{H}$	microchoke
$L_4 = L_5$	stripline ($Z_c = 21 \text{ }\Omega$), $6.7 \times 6.0 \text{ mm}^2$; note 1
$L_7 = L_8 = \text{stripline}$	($Z_c = 21 \text{ }\Omega$), $3.0 \times 6.0 \text{ mm}^2$; note 1
$L_9 = L_{10} = \text{stripline}$	($Z_c = 72 \text{ }\Omega$), $15.2 \times 1.0 \text{ mm}^2$; note 1
$L_{11} = L_{12} = \text{stripline}$	($Z_c = 72 \text{ }\Omega$), $14.3 \times 1.0 \text{ mm}^2$; note 1
$L_{13} = L_{14} = \text{stripline}$	($Z_c = 72 \text{ }\Omega$), $29.9 \times 1.0 \text{ mm}^2$; note 1

Note

1. These striplines are printed on double Cu-clad printed-circuit board with PTFE fibre-glass dielectric ($\epsilon_r = 2.74$); thickness 1/32".

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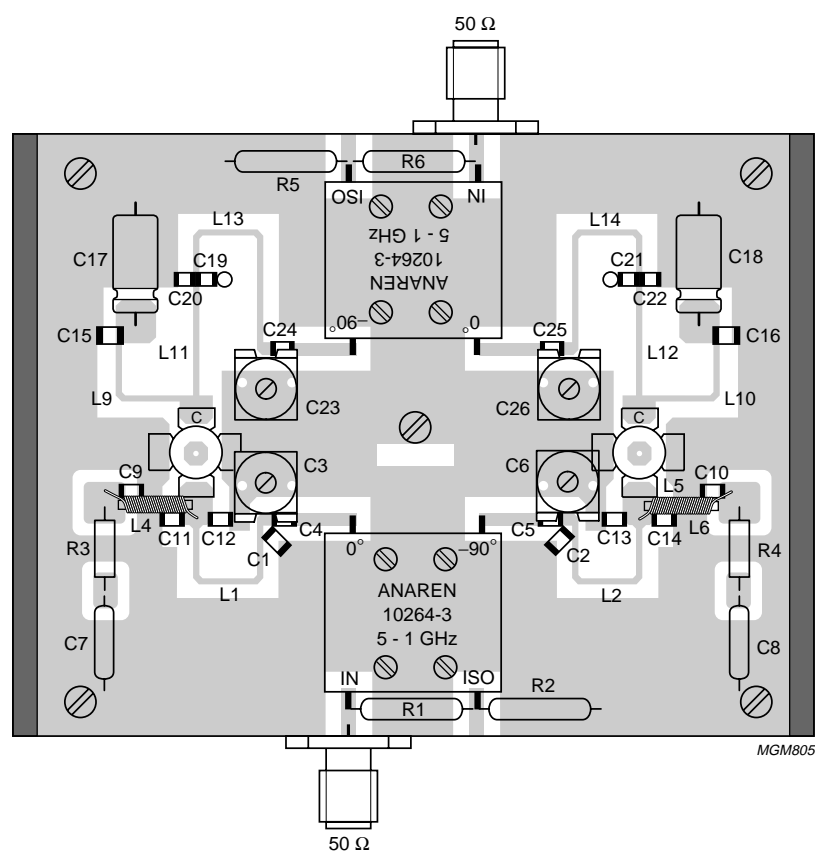
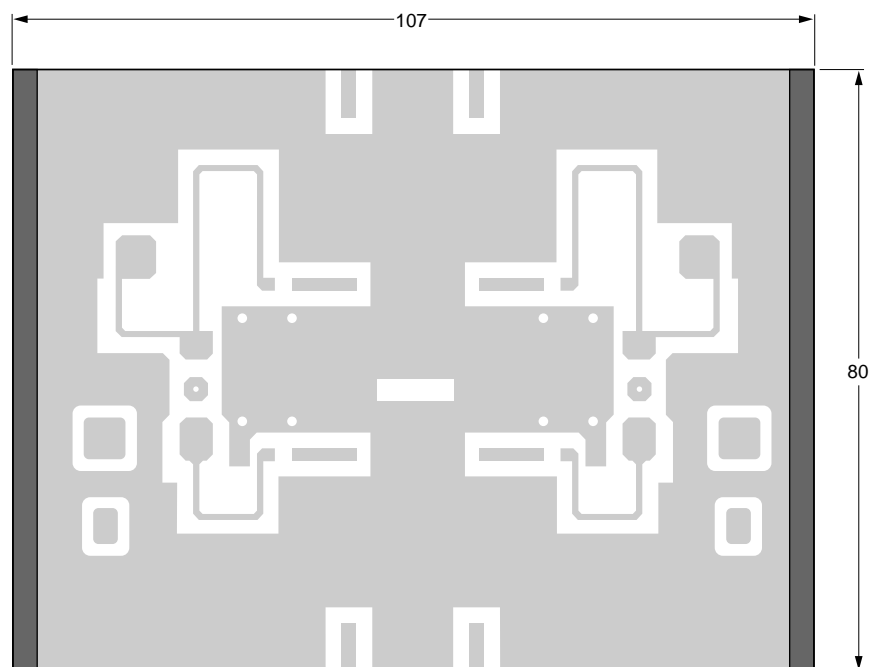


Fig.6 Printed-circuit board and amplifier lay-out.

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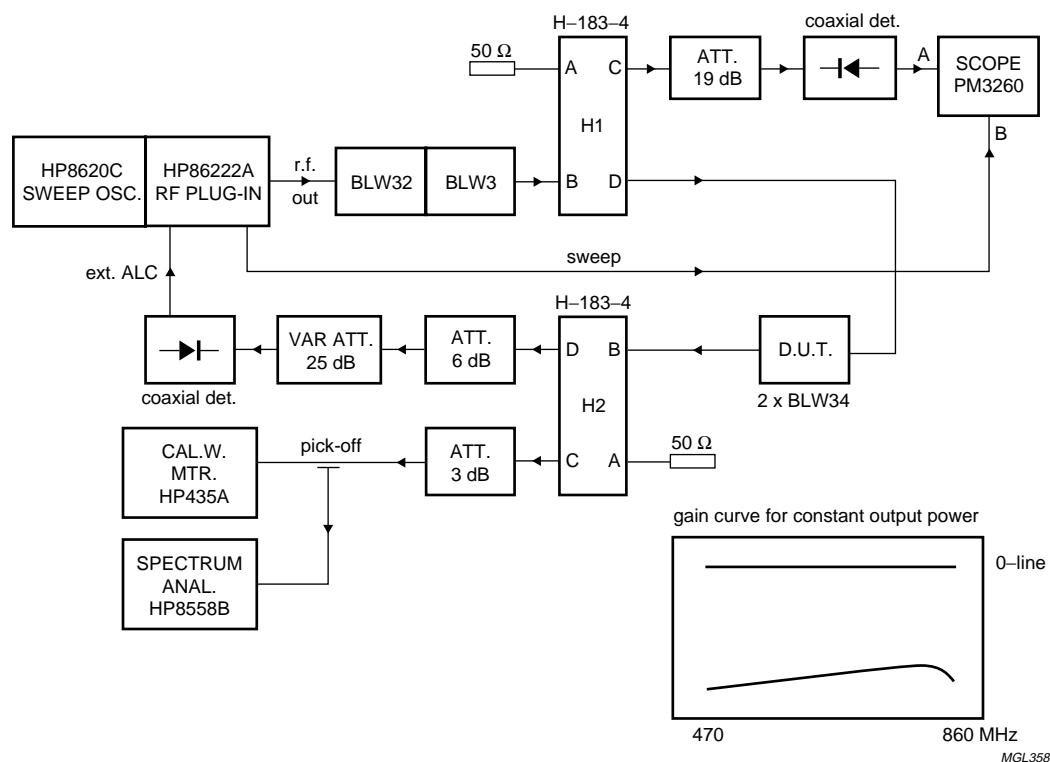
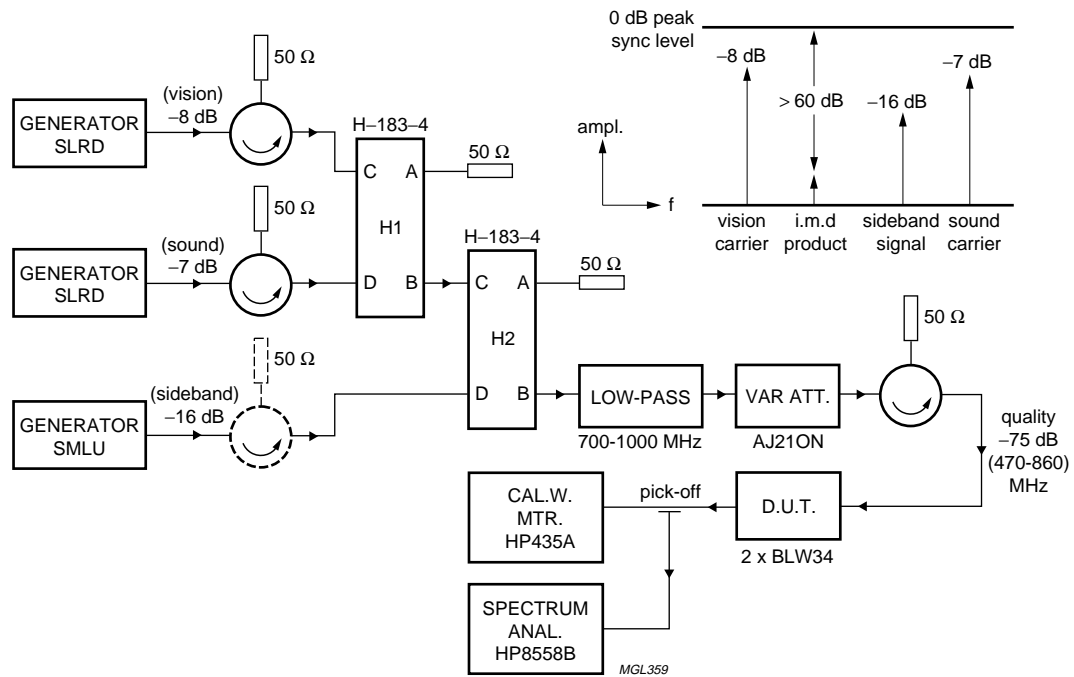


Fig.7 Dynamic compression test set-up.

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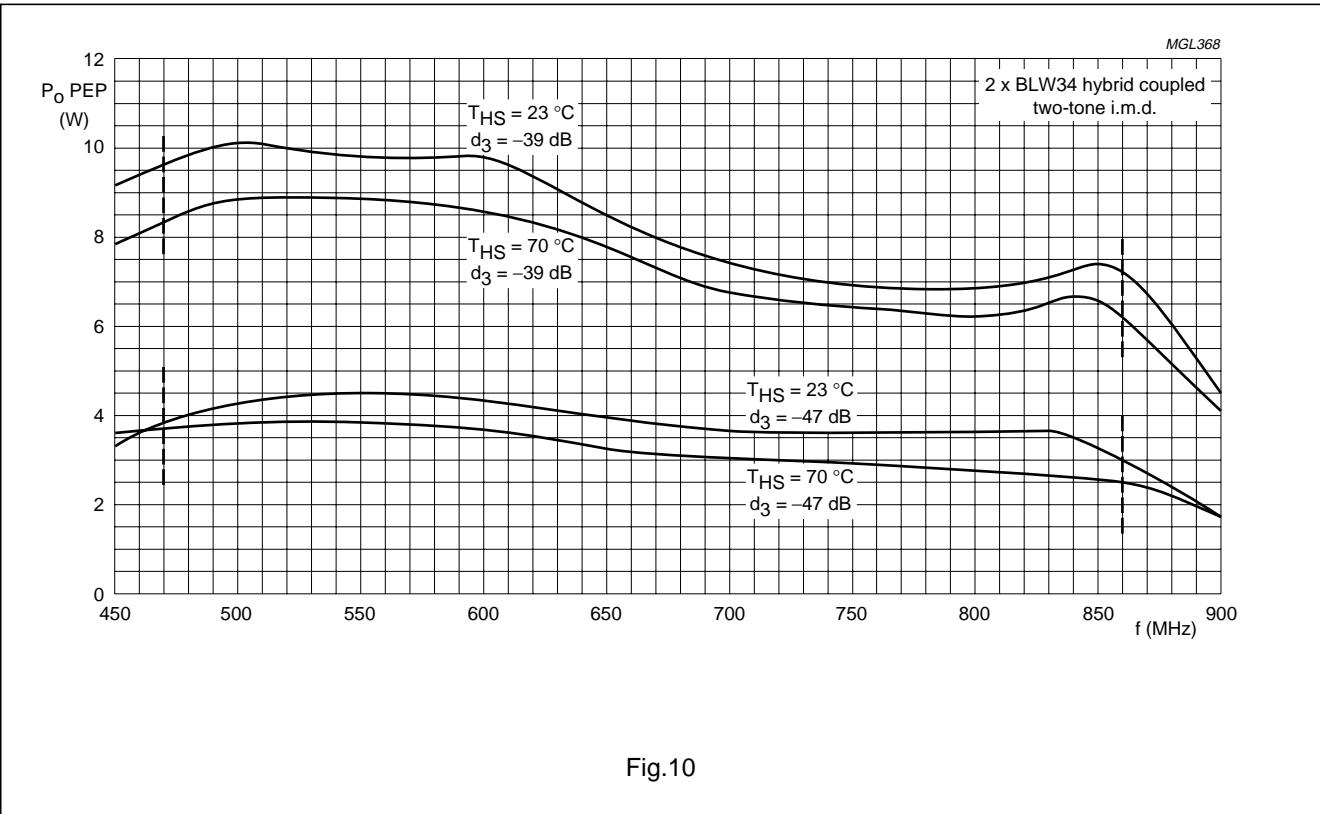
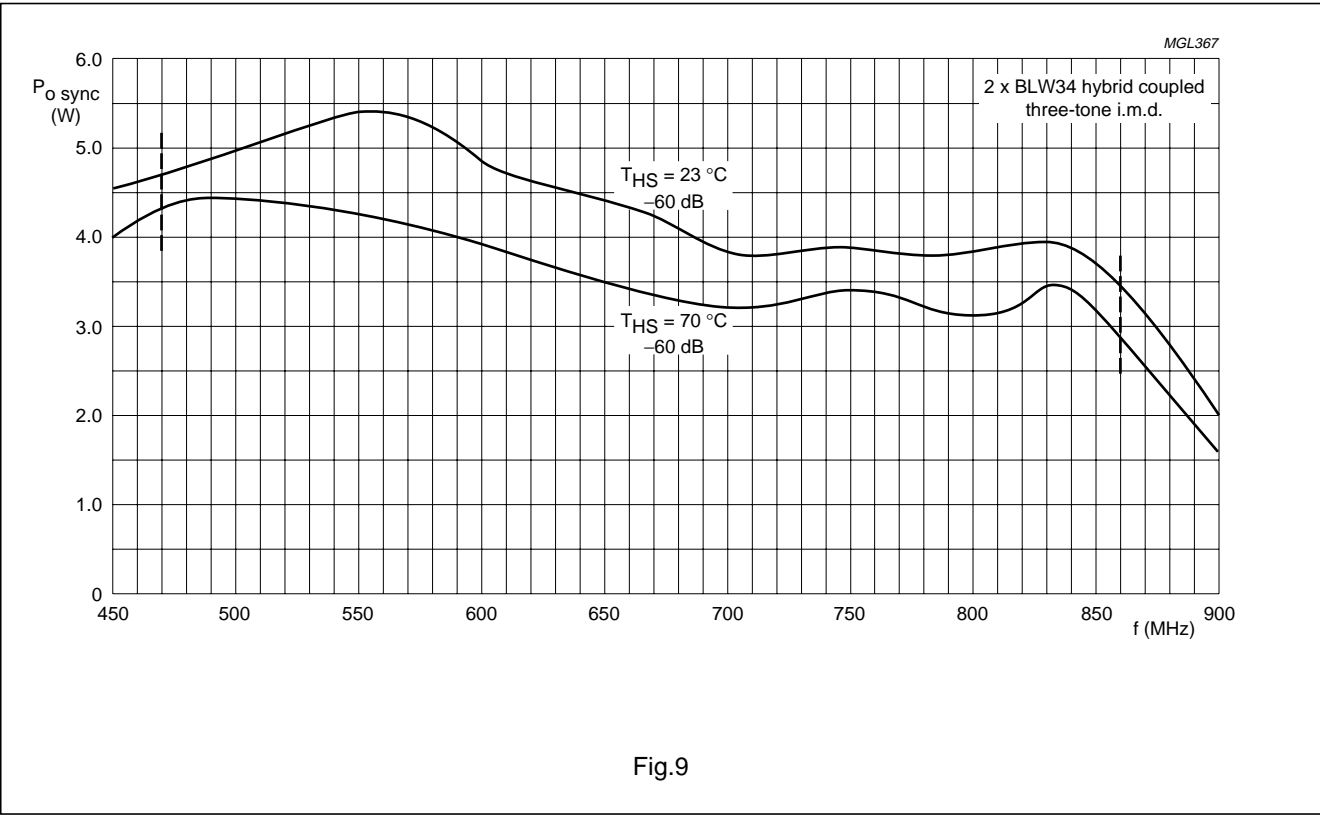
Circulator ranges:

- (1) 470 – 600 MHz.
- (2) 600 – 800 MHz.
- (3) 790 – 1000 MHz.

Fig.8 3-tone test set-up 470 – 860 MHz.

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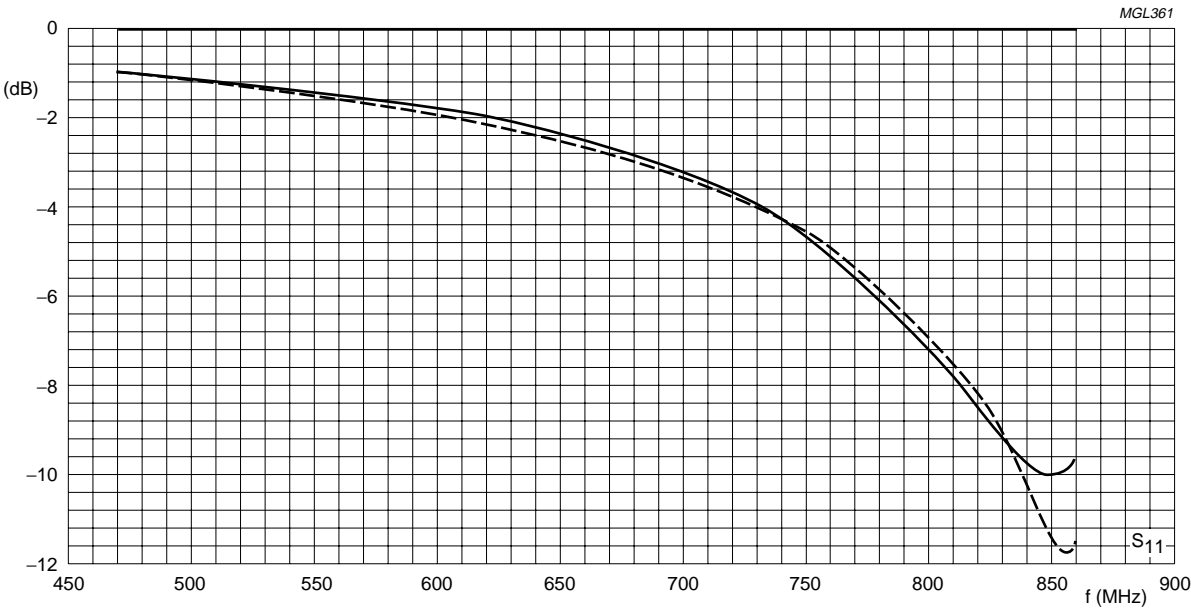
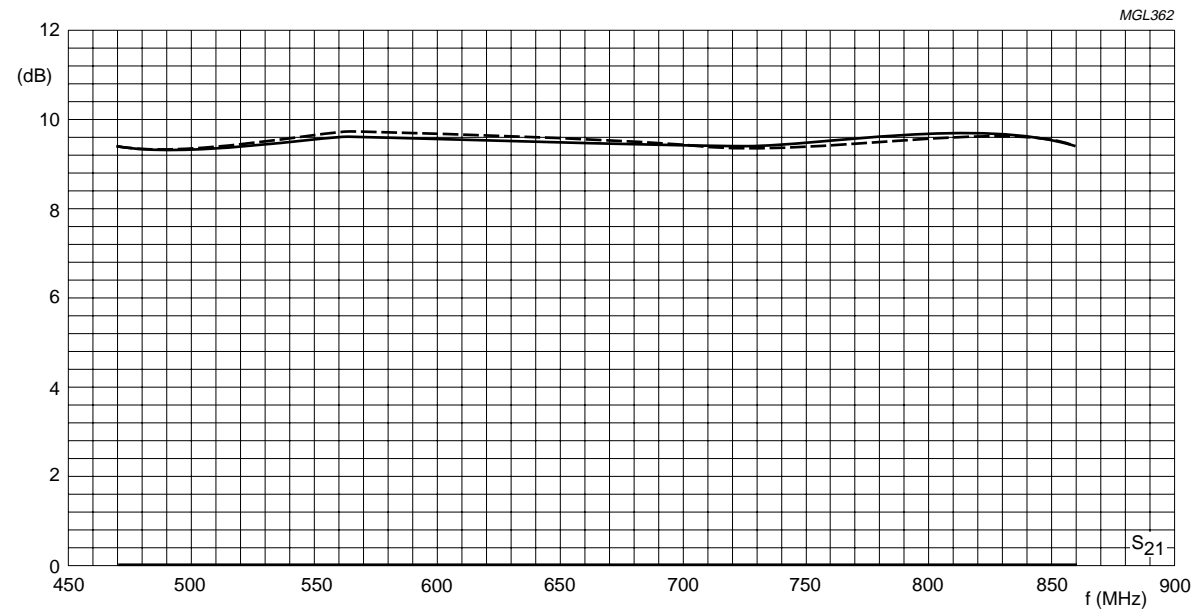


Fig.11



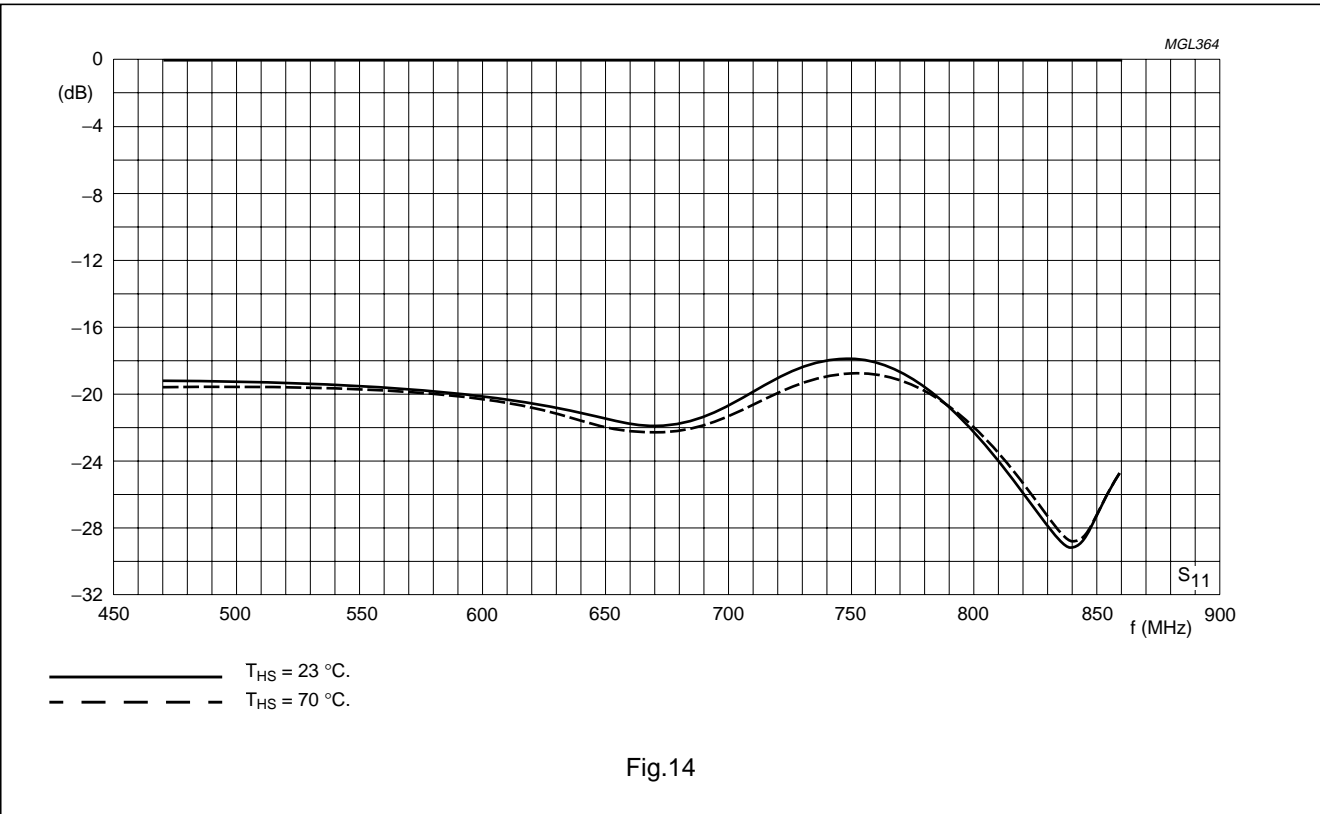
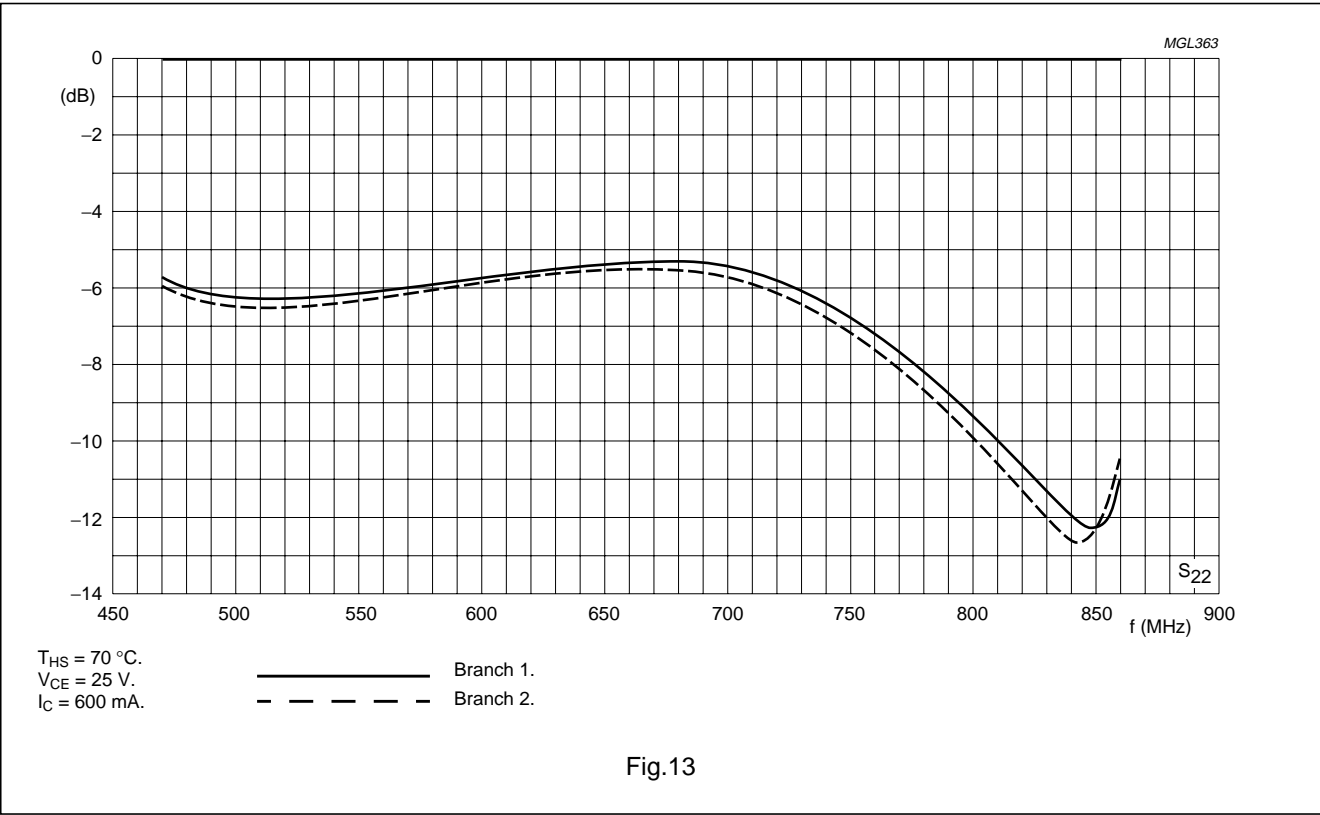
$T_{HS} = 70\text{ }^{\circ}\text{C.}$
 $V_{CE} = 25\text{ V.}$
 $I_C = 600\text{ mA.}$

— Branch 1.
- - - Branch 2.

Fig.12

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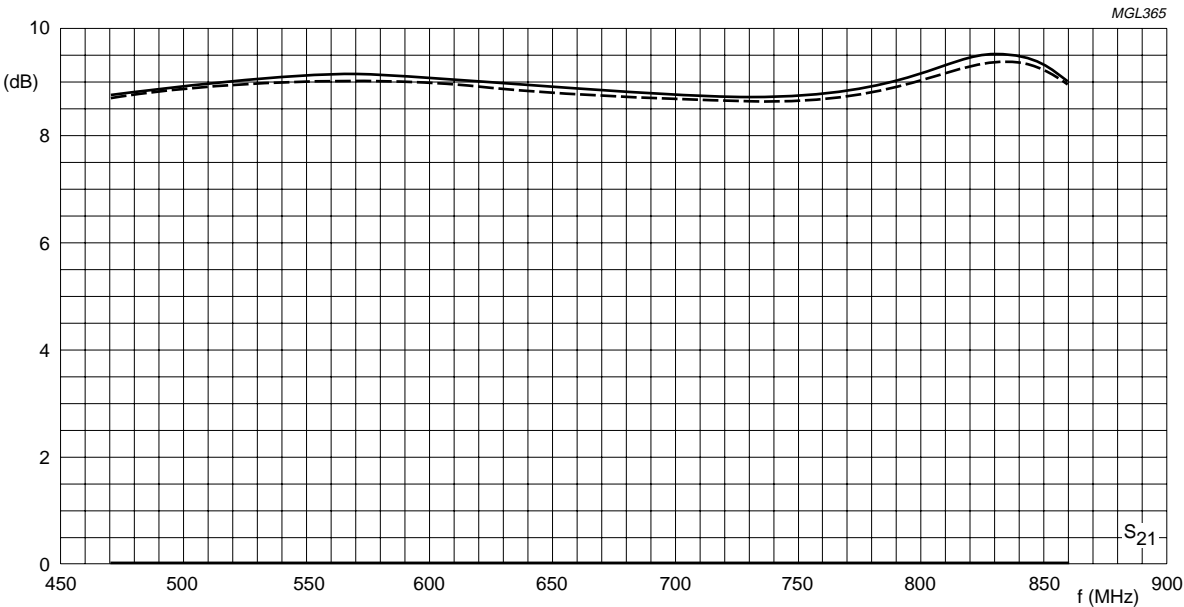


Fig.15

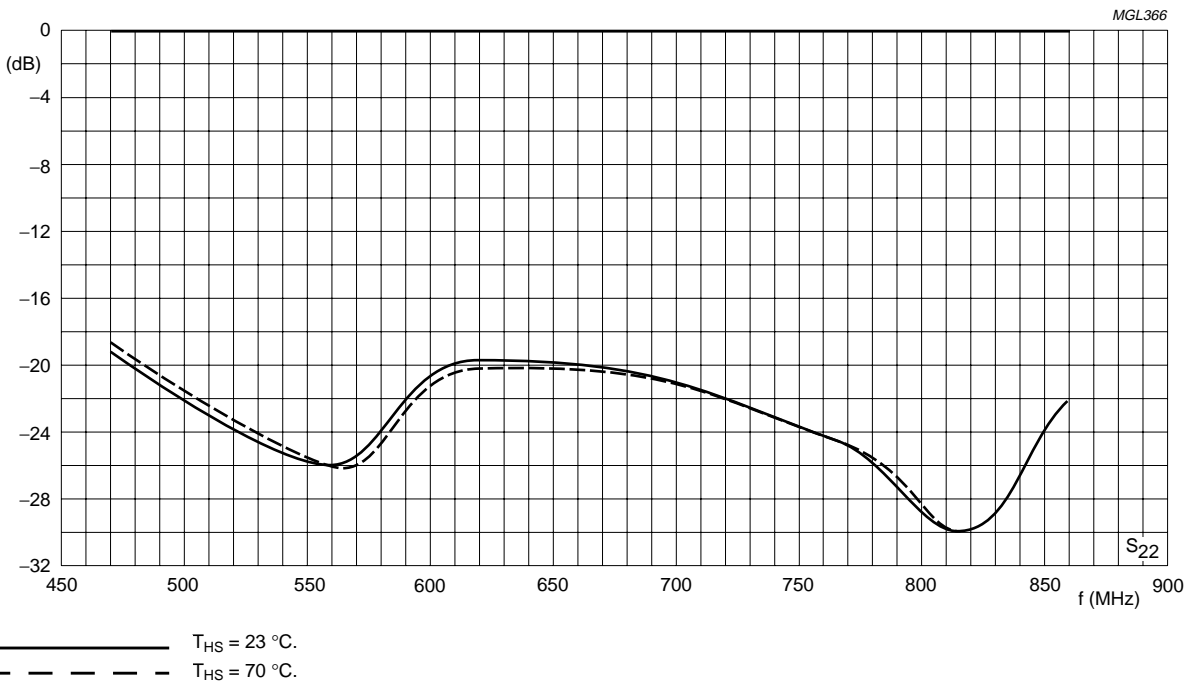


Fig.16

Philips Semiconductors – a worldwide company

Argentina: see South America

Australia: 34 Waterloo Road, NORTH RYDE, NSW 2113,
Tel. +61 2 9805 4455, Fax. +61 2 9805 4466

Austria: Computerstr. 6, A-1101 WIEN, P.O. Box 213, Tel. +43 160 1010,
Fax. +43 160 101 1210

Belarus: Hotel Minsk Business Center, Bld. 3, r. 1211, Volodarski Str. 6,
220050 MINSK, Tel. +375 172 200 733, Fax. +375 172 200 773

Belgium: see The Netherlands

Brazil: see South America

Bulgaria: Philips Bulgaria Ltd., Energoproject, 15th floor,
51 James Bourchier Blvd., 1407 SOFIA,
Tel. +359 2 689 211, Fax. +359 2 689 102

Canada: PHILIPS SEMICONDUCTORS/COMPONENTS,
Tel. +1 800 234 7381

China/Hong Kong: 501 Hong Kong Industrial Technology Centre,
72 Tat Chee Avenue, Kowloon Tong, HONG KONG,
Tel. +852 2319 7888, Fax. +852 2319 7700

Colombia: see South America

Czech Republic: see Austria

Denmark: Prags Boulevard 80, PB 1919, DK-2300 COPENHAGEN S,
Tel. +45 32 88 2636, Fax. +45 31 57 0044

Finland: Sinikalliontie 3, FIN-02630 ESPOO,
Tel. +358 9 615800, Fax. +358 9 61580920

France: 51 Rue Carnot, BP317, 92156 SURESNES Cedex,
Tel. +33 1 40 99 6161, Fax. +33 1 40 99 6427

Germany: Hammerbrookstraße 69, D-20097 HAMBURG,
Tel. +49 40 23 53 60, Fax. +49 40 23 536 300

Greece: No. 15, 25th March Street, GR 17778 TAVROS/ATHENS,
Tel. +30 1 4894 339/239, Fax. +30 1 4814 240

Hungary: see Austria

India: Philips INDIA Ltd, Band Box Building, 2nd floor,
254-D, Dr. Annie Besant Road, Worli, MUMBAI 400 025,
Tel. +91 22 493 8541, Fax. +91 22 493 0966

Indonesia: see Singapore

Ireland: Newstead, Clonskeagh, DUBLIN 14,
Tel. +353 1 7640 000, Fax. +353 1 7640 200

Israel: RAPAC Electronics, 7 Kehilat Saloniki St, PO Box 18053,
TEL AVIV 61180, Tel. +972 3 645 0444, Fax. +972 3 649 1007

Italy: PHILIPS SEMICONDUCTORS, Piazza IV Novembre 3,
20124 MILANO, Tel. +39 2 6752 2531, Fax. +39 2 6752 2557

Japan: Philips Bldg 13-37, Kohnan 2-chome, Minato-ku, TOKYO 108,
Tel. +81 3 3740 5130, Fax. +81 3 3740 5077

Korea: Philips House, 260-199 Itaewon-dong, Yongsan-ku, SEOUL,
Tel. +82 2 709 1412, Fax. +82 2 709 1415

Malaysia: No. 76 Jalan Universiti, 46200 PETALING JAYA, SELANGOR,
Tel. +60 3 750 5214, Fax. +60 3 757 4880

Mexico: 5900 Gateway East, Suite 200, EL PASO, TEXAS 79905,
Tel. +9-5 800 234 7381

Middle East: see Italy

Netherlands: Postbus 90050, 5600 PB EINDHOVEN, Bldg. VB,
Tel. +31 40 27 82785, Fax. +31 40 27 88399

New Zealand: 2 Wagener Place, C.P.O. Box 1041, AUCKLAND,
Tel. +64 9 849 4160, Fax. +64 9 849 7811

Norway: Box 1, Manglerud 0612, OSLO,
Tel. +47 22 74 8000, Fax. +47 22 74 8341

Philippines: Philips Semiconductors Philippines Inc.,
106 Valero St. Salcedo Village, P.O. Box 2108 MCC, MAKATI,
Metro MANILA, Tel. +63 2 816 6380, Fax. +63 2 817 3474

Poland: Ul. Lukiska 10, PL 04-123 WARSZAWA,
Tel. +48 22 612 2831, Fax. +48 22 612 2327

Portugal: see Spain

Romania: see Italy

Russia: Philips Russia, Ul. Usatcheva 35A, 119048 MOSCOW,
Tel. +7 095 755 6918, Fax. +7 095 755 6919

Singapore: Lorong 1, Toa Payoh, SINGAPORE 1231,
Tel. +65 350 2538, Fax. +65 251 6500

Slovakia: see Austria

Slovenia: see Italy

South Africa: S.A. PHILIPS Pty Ltd., 195-215 Main Road Martindale,
2092 JOHANNESBURG, P.O. Box 7430 Johannesburg 2000,
Tel. +27 11 470 5911, Fax. +27 11 470 5494

South America: Al. Vicente Pinzon, 173, 6th floor,
04547-130 SÃO PAULO, SP, Brazil,
Tel. +55 11 821 2333, Fax. +55 11 821 2382

Spain: Balmes 22, 08007 BARCELONA,
Tel. +34 3 301 6312, Fax. +34 3 301 4107

Sweden: Kottbygatan 7, Akalla, S-16485 STOCKHOLM,
Tel. +46 8 632 2000, Fax. +46 8 632 2745

Switzerland: Allmendstrasse 140, CH-8027 ZÜRICH,
Tel. +41 1 488 2686, Fax. +41 1 488 3263

Taiwan: Philips Semiconductors, 6F, No. 96, Chien Kuo N. Rd., Sec. 1,
TAIPEI, Taiwan Tel. +886 2 2134 2865, Fax. +886 2 2134 2874

Thailand: PHILIPS ELECTRONICS (THAILAND) Ltd.,
209/2 Sanpavuth-Bangna Road Prakanong, BANGKOK 10260,
Tel. +66 2 745 4090, Fax. +66 2 398 0793

Turkey: Talatpasa Cad. No. 5, 80640 GÜLTEPE/ISTANBUL,
Tel. +90 212 279 2770, Fax. +90 212 282 6707

Ukraine: PHILIPS UKRAINE, 4 Patrice Lumumba str., Building B, Floor 7,
252042 KIEV, Tel. +380 44 264 2776, Fax. +380 44 268 0461

United Kingdom: Philips Semiconductors Ltd., 276 Bath Road, Hayes,
MIDDLESEX UB3 5BX, Tel. +44 181 730 5000, Fax. +44 181 754 8421

United States: 811 East Arques Avenue, SUNNYVALE, CA 94088-3409,
Tel. +1 800 234 7381

Uruguay: see South America

Vietnam: see Singapore

Yugoslavia: PHILIPS, Trg N. Pasica 5/v, 11000 BEOGRAD,
Tel. +381 11 625 344, Fax. +381 11 635 777

For all other countries apply to: Philips Semiconductors,
International Marketing & Sales Communications, Building BE-p, P.O. Box 218,
5600 MD EINDHOVEN, The Netherlands, Fax. +31 40 27 24825

Internet: <http://www.semiconductors.philips.com>

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