APPLICATION NOTE

A wideband linear power amplifier (1.6 − 28 MHz) for 300 W PEP with 2 MOS transistors BLF177

NCO8703

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CONTENTS

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1 SUMMARY

This report gives a description of a wideband push-pull amplifier for the frequency range 1.6 − 28 MHz.
The amplifier has been designed around 2 MOS transistors BLF177 which operate in class-AB at V_{DS} = 50 V and I_{DQ} = 0.5 A/transistor.
The main properties at P_O = 300 W are:

Powergain: 22 to 23 dB

Efficiency: 52.5 to 61%

Return losses input: $\leq$ −15.5 dB

2nd harmonics: $\leq$ −25 dB

3rd harmonics: $\leq$ −16 dB

IMD at P_O = 300 W PEP: $\leq$ −33 dB.

2 INTRODUCTION

The BLF177 is an RF Power MOS transistor for the HF and VHF range in a 4 leads flange SOT121 encapsulation. For the frequency range 1.6 − 28 MHz a wideband push-pull power amplifier has been developed with 2 × BLF177 having an output power of 300 W PEP at an intermodulation distortion level below −30 dB.
The transistors operate in class-AB at V_{DS} = 50 V and a quiescent current of 0.5 A each.

3 DESIGN OF THE AMPLIFIER
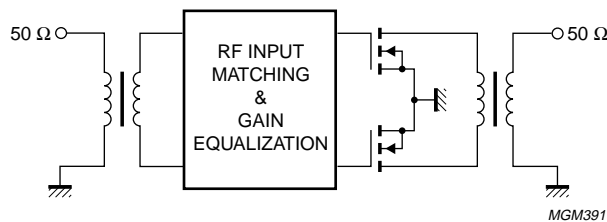
3.1 General

The schematic set-up is given in Fig.1.

Fig.1 Schematic set-up 2 × BLF177 amplifier.

The two balance to unbalance transformers are applied to split the single ended input into 2 out of phase driving ports and to add the 2 out of phase output ports into one single ended output. The transformers have an impedance transformation ratio of 4 : 1 and match the low-ohmic in- and output impedance of the transistors to the 50 Ω system impedance. At the input a special circuit takes care of a good input matching and a flat powergain over the whole bandwidth.

3.2 Output circuit

3.2.1 LOAD IMPEDANCE

The output impedance of each transistor can be represented as a combination of the output capacitance C_{oss} and the optimum load resistance. Because of the larger drain voltage swing the effective output capacitance C_O is appr. 15% higher than the value of C_{oss}. So $C_O = 1.15 \times 190 \approx 220$ pF. The optimum load resistance for class-AB can be determined with formula:

$R_L = (0.85 \times V_{DS})^2/(2 \times P_O)$

For $V_{DS} = 50$ V and $P_O = 150$ W we get $R_L = 6 \, \Omega$. To keep the transformer simple a transformation ratio of 4 or 9 is preferable. A ratio of 4 gives a load impedance of 50/4 = 12.5 $\Omega \rightarrow 6.25 \, \Omega$ for each transistor. This is very near to the optimum load resistance.

3.2.2 OUTPUT TRANSFORMER

The output transformer has to transform the 50 Ω asymmetrical impedance to the $2 \times 6.25 = 12.5 \, \Omega$ symmetrical load impedance. The reactance (ωL) of the shunting inductance at 1.6 MHz has been chosen at 4 times 50 Ω = 200 Ω. So the inductance is 20 μH. The transformer has been wound on a ferrite toroid of 4C6 material. Dimensions: $36 \times 23 \times 15$ mm ($D \times d \times h$) which gives a volume (A.1) = 8.97E-6m³.
Because the power handling of one toroid is critical two transformers in parallel with an inductance of 40 μH each have been chosen.

$n_{sec} = SQR((L.1)/(\mu o \times \mu r \times A))$
$n_{sec} = SQR((40E − 6 \times 9.2E − 2)/(4\pi E − 7 \times 120 \times 97.6E − 6)) = 15.8$ turns.
So $n_{pr} = 8$ turns and $n_{sec} = 16$ turns.

For each transformer V_{max} depends on the power over 100 Ω.
$V_{max} = SQR(2 \times P_O \times R_L) = SQR(2 \times 150 \times 100) = 173.2$ V

B_{max} depends on the parallel loss resistance at 1.6 MHz; for a power loss of 1%: $B_{max} = 1.3E − 2T$.
The volume A.1 needed per core is:
$A.1 = (V_{max}/(\omega \times B_{max}))^2(\mu_o \times \mu_r)/L$.

$A.1 = (1.73.2/(2\pi \times 1.6E + 6 \times 0.013))^2(4\pi E − 7 \times 120)/40E − 6 = 6.62E − 6$ m³.

Each of the toroids has a volume of 8.97E − 6 m³. Figure 7 shows one of the two parallel connected output transformers. On each toroid the primary winding has 8 turns of copperfoil (width 5 mm and thickness 0.05 mm). The secondary winding has 16 turns of 2 enamelled copper wires (0.6 mm) in parallel.
So each primary turn has been covered with 2 secondary turns which means 4 wires of 0.6 mm. Both windings are isolated with PTFE-foil (thickness 0.1 mm). To reduce the stray-inductance the transformer has been wound as follows:

1. The primary has been wound evenly around the periphery of the toroid

2. With the secondary the same has been done with the first 8 turns; the second part of 8 turns has been wound in between the first part. So the secondary has been wound twice around the core.

The measured secondary inductance of each transformer is 38 μH and $L_{str} = 300$ nH.
With the aid of a network analyser the parallel combination of these 2 transformers has been corrected. For the higher frequencies at the low-ohmic side a parallel capacitor of 240 pF and for the lower frequencies at the high-ohmic side a series capacitor of 10 nF give return losses below −21 dB over the whole frequency range (see Fig.2).
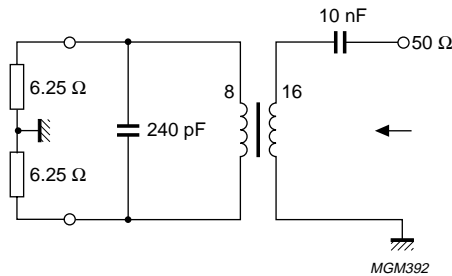
Fig.2 Output transformer with correction.

Replacing the transistors by resistors of 6.25 Ω the return loss can be measured at the 50 Ω side. Figure 11 gives the return losses of the parallel combination of the two transformers before and after the correction.

3.2.3   THE TAPPED CHOKE

The chokes in the drain circuits are wound around a common ferrite rod of 4B1 material. Dimensions: 50 × 10 mm (l × d). Figure 3 gives a schematic electrical circuit of the output.
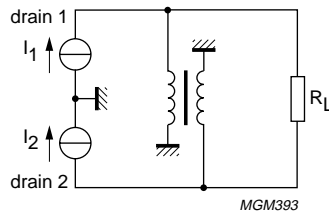

Fig.3 Output with the tapped choke.

Between both drains the impedance for the even harmonics depends on the coupling factor between both windings. If the coupling factor amounts to 1 both drains will be short circuited for the even harmonics.
Because the voltage over one winding is equal to half of the voltage between both drains, the total inductance between both drains is 4 times the inductance of one winding.
The reactance of the shunting inductance at 1.6 MHz has been chosen at 4 times 12.5 Ω = 50 Ω.

So the inductance between both drains is 5 μH, this means for one winding an inductance of 1.25 μH. According to the Philips Data Handbook *"MA01 on soft ferrites of 1996, the effective permeability"* of a rod with 1/d = 5 and μ_r = 250 is appr. 20.

The number of turns can be calculated with:

$n = SQR(L.1/(\mu_o \times \mu_r \times A))$

$n = SQR(1.25E − 6 \times 50E − 3/(4\pi E − 7 \times 20 \times \frac{1}{4}\pi(10E − 3)^2) = 5.6$ turns. In practice 6 twisted turns of the primary and secondary windings have been wound around the rod. Figure 8 shows the tapped choke. To increase the coupling factor each winding consists of 2 enamelled copper wires (0.8 mm) in parallel. The measured inductance is 1.275 μH.

3.2.4 TUNING OF THE OUTPUT CIRCUIT

For an optimum alignment of the output circuit the 2 transistors have been replaced by dummies consisting of the parallel connection of a resistance and a capacitance. The resistance is equal to the optimum load resistance and the capacitance to the output capacitance (see Section 3.2.1).

Tuning of the output circuit has been carried out by measuring the return losses at the output with a network analyser under swept conditions (see Fig.4).
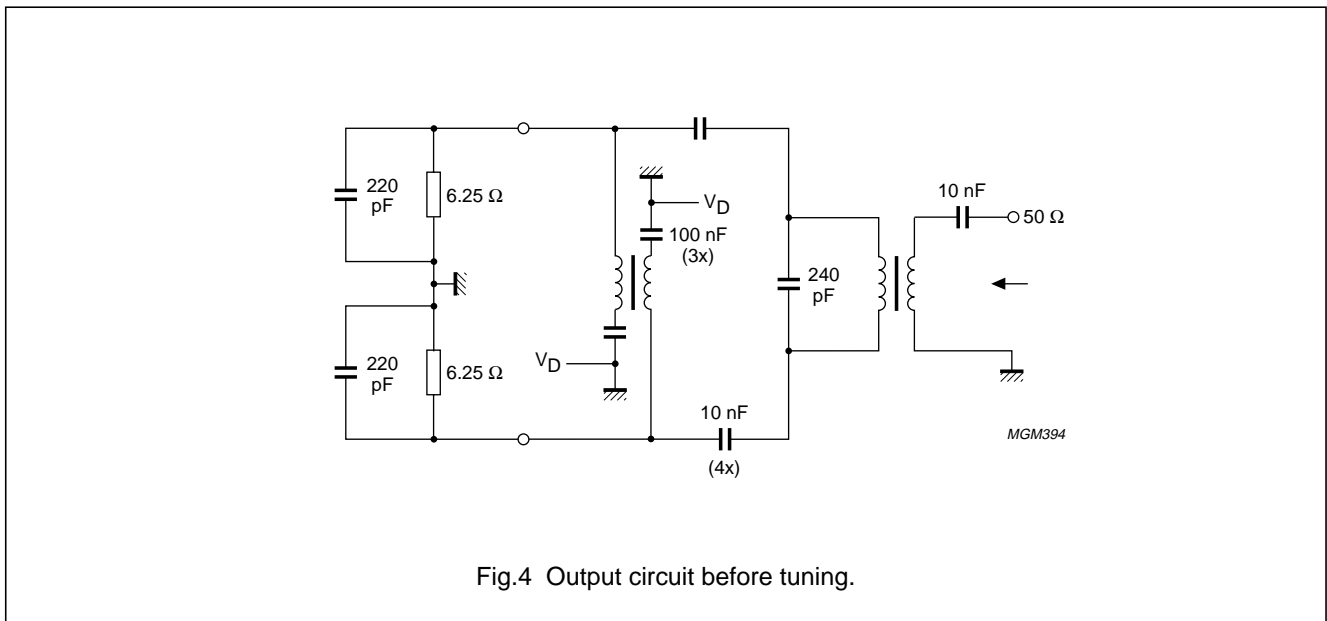

Fig.4 Output circuit before tuning.

The measured return losses should be as low as possible by changing the correction capacitors. Figure 12 shows the return losses of the output before and after tuning. For optimum results the capacitance across the primary winding of the output transformer has been reduced from 240 to 150 pF and the low frequency correction capacitor of 10 nF at the output has been changed to an inductance of 100 nH. The last change can be explained as follows:

1. The low frequency compensation is taken over by the coupling capacitors between the drain choke and the impedance transformer

2. The function of the transformer is not only impedance matching but also transfer from balanced to unbalanced. The latter makes that the interwinding capacitance has more influence. This is so much that a series inductance at the output is needed for high frequency compensation.

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3.3 Input circuit

3.3.1 INPUT IMPEDANCE

The input impedance and gain of the transistor can be determined with the aid of a computer model of the BLF177. Table 1 shows the calculated gain and impedances for the frequency range 1.6 to 28 MHz.

BLF177 Vds = 50 V Po = 150 W Class-AB.

Table 1 Calculated gain and impedances of the BLF177

f (MHz)	G (dB)	INP.IMP. (Ω)	LOAD IMP. (Ω)
1.6	54.70	2.29 − j133.58	6.23 + j.07
2.5	50.82	2.29 − j85.50	6.23 + j.12
3.5	47.90	2.29 − j61.09	6.23 + j.16
5.0	44.80	2.29 − j42.78	6.22 + j.23
7.0	41.88	2.29 − j30.58	6.21 + j.32
10.0	38.78	2.29 − j21.45	6.18 + j.46
14.0	35.85	2.29 − j15.37	6.13 + j.64
20.0	32.75	2.29 − j10.84	6.03 + j.89
24.0	31.17	2.29 − j9.09	5.95 + j1.05
28.0	29.83	2.29 − j7.85	5.85 + j1.20

By adding a gate-source resistor of 6.25 Ω the power gain reduces from 29.8 to 23.3 dB at 28 MHz.

3.3.2 INPUT MATCHING CIRCUIT

As mentioned in Section 3.1 a special circuit matches the input impedance of each transistor to the 6.25 Ω of the input transformer. The matching network chosen can be treated as the half of a double π-section as described in Ref.1. Removing the in- and output capacitance the circuit changes in a T-section with C_i as capacitor and 2 inductances with a value of half the inductances of the double π-section (see Fig.5).
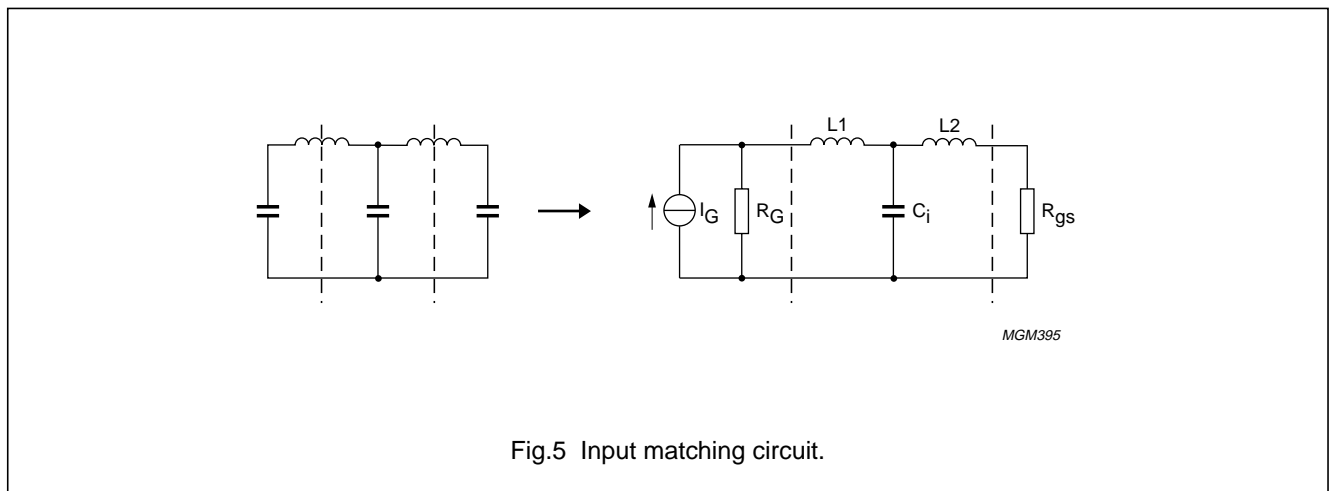

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Fig.5 Input matching circuit.

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C_i represents the input capacitance of the BLF177 and can be calculated from the input impedance of Table 1.

For 7 MHz: $C_i = 1/(2\pi \times 7E + 6 \times 30.58) = 744$ pF.

Across this capacitor a constant voltage versus frequency from 1.6 up to 28 MHz has to be developed. Provided C_i is an ideal capacitance the dimensioning of this network is as follows:

$R_G = R_{gs}$ must be appr. 6 Ω to obtain low I.M. distortion and good stability. This appeared during the development of the narrow band testcircuit as given in the BLF177 publication data. To judge whether this value is also acceptable for wideband operation we calculate the product:

$W_c \times C_i \times R_{gs}$ in which ω_c is the maximum angular frequency.

Doing so we find:

$2\pi \times 28E + 6 \times 744E - 12 \times 6.25 = 0.818$

R_{gs} has been chosen 6.25 Ω for the ease of transformation. Comparing the value of this product with the one given in Ref.2 we see that with a double π-section we can easily reach a bandwidth of 50 MHz. Therefore we have simplified the network as described above. Continuing the calculation we find:

$L = 0.997 R_G/\omega_c = 35.4$ nH (So L1 = L2 = 17.7 nH)

With the computer model mentioned in Section 3.3.1 a gain of 22.3 dB has been calculated with $R_{gs} = 6.25$ Ω.

Starting from this 22.3 dB gain, $L_1 = L_2 = 17.7$ nH and $R_{gs} = 6.25$ Ω the input VSWR and gain deviation have been calculated (see Table 2).

Initial results
Rs = 6.250 Ω; Gs = 22.300 dB
Par.LR: L = 17.700 nH; R = 6.250 Ω
Ser.Ind.: L = 17.700 nH

Table 2 Results before optimization

f (MHz)	VSWR	dG (dB)
1.6	1.010	1.440
2.5	1.016	1.435
3.5	1.023	1.431
5.0	1.032	1.419
7.0	1.046	1.402
10.0	1.068	1.354
14.0	1.100	1.265
20.0	1.161	1.075
24.0	1.214	.910
28.0	1.281	.703

Before optimization the maximum VSWR = 1.28 and the gain = 22.7 dB ± 0.37 dB. To achieve a maximally flat gain and a low input VSWR a computer optimization program has been used. This optimization results in a gain of 23.3 dB with a maximum ΔGain = ±0.09 dB and a VSWR ≤1.09, see Table 3. For these results L_1 has been changed from 17.7 nH to 9 nH and L_2 from 17.7 nH to 21.1 nH. the R_{gs} has been decreased from 6.25 Ω to 5.7 Ω.

Final results
Rs = 6.250 Ω; Gs = 23.300 dB
Par.LR: L = 21.079 nH; R = 5.749 Ω
Ser.Ind.: L = 8.950 nH

Table 3 Results after optimization

f (MHz)	VSWR	dG (dB)
1.6	1.087	.072
2.5	1.087	.071
3.5	1.087	.074
5.0	1.086	.076
7.0	1.085	.084
10.0	1.082	.087
14.0	1.076	.086
20.0	1.059	.056
24.0	1.047	.006
28.0	1.044	−.087

3.3.3 INPUT TRANSFORMER

The input transformer is simular to the output transformer. It transforms the asymmetrical system impedance to the $2 \times 6.25 \, \Omega = 12.5 \, \Omega$ symmetrical source impedance. However the lower power handling (<3 W) justifies a toroid of 4C6 material with smaller dimensions: $14 \times 9 \times 5$ mm ($D \times d \times h$) which gives a volume A.1 = 0.445E − 6 m^3. As described in Section 3.2.2 the primary winding can be calculated for L = 20 µH.

$n_{pr} = SQR((L.1)/(\mu_o \times \mu_r \times A))$

$n_{pr} = SQR((20E − 6 \times 3.55E − 2)/(4\pi E − 7 \times 120 \times 12.54E − 6)) = 19.4$ turns. With $n_{pr} = 20$ turns and a transformation ratio of 4 : 1 the $n_{sec} = 10$ turns.
$V_{max} = SQR(2 \times P_i \times R_S) = 17.3$ V and $B_{max} = 0.013$T.

The needed core volume A.1 is:

$A.1 = (V_{max}/(\omega \times B_{max}))^2 \times (\mu_o \times \mu_r)/L$

$A.1 = (17.3/(2\pi \times 1.6E + 6 \times 0.013)^2 \times (4\pi E − 7 \times 120)/20E − 6 = 0.14E − 6$ m^3

The core used has a volume of 0.445E − 6m^3.
Figure 9 shows the input transformer. The secondary winding has 10 turns of copperfoil (width 2 mm, thickness 0.05 mm). The primary winding has 20 turns of enamelled copper wire (0.5 mm). Each secondary turn has been covered with 2 primary turns with a PTFE foil of 0.1 mm thickness as isolation between the 2 windings. The method of winding is the same as described for the output transformer in Section 3.2.2. The measured inductance is 20.95 µH and $L_{str} = 250$ nH.
The correction method used for the input transformer is the same as described already in Section 3.2.2 (see Fig.6).

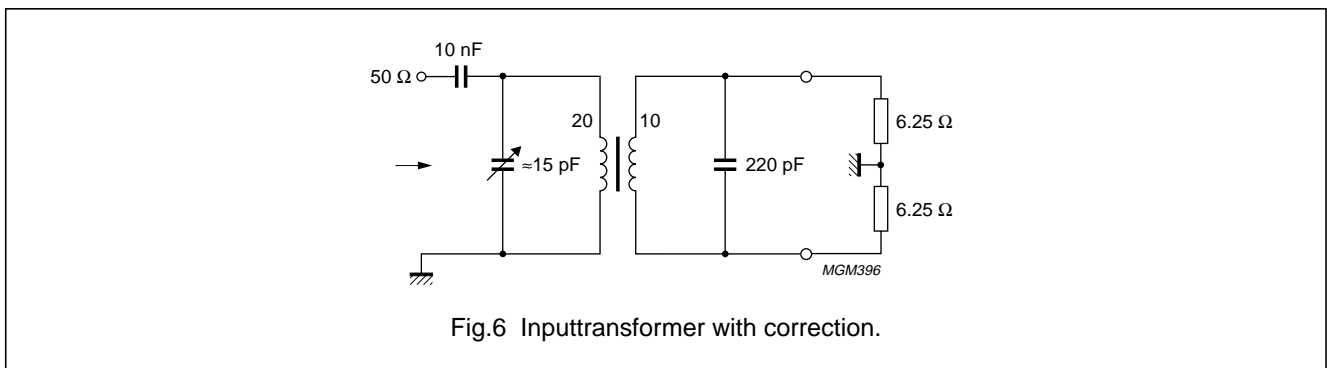Fig.6 Inputtransformer with correction.

The transformer has been corrected with parallel capacitors for the higher frequencies and a series capacitor for the lower frequencies. Figure 13 gives the return losses before and after the correction.

3.3.4   TUNING OF THE INPUTCIRCUIT

For the practical tuning of the inputcircuit each transistor has been adjusted at V_{DS} = 50 V and a quiescent current of 0.5 A.

The gain and input return losses have been measured in the frequency range 1.6 up to 35 MHz. The best results have been achieved by changing the secondary correction capacitor of the inputtransformer from 220 to 30 pF and the primary correction trimmer from ≈ 15 pF to ≈ 20 pF. The low frequency correction capacitor at the input has been removed. The inductance in serie with R_{gs} has been increased from 21.6 to 35 nH. Figure 14 gives the complete circuit diagram of the wide band amplifier with 2 BLF177 transistors. Table 5 gives the corresponding parts list.

4   CONSTRUCTION OF THE AMPLIFIER

For the printed circuit board double Cu-clad epoxy fibre glass has been used with a thickness of 1/16" and ε_r = 4.5. The position of the components is on one side and the other side serves as a groundplane. Connections to the groundplane have been made with rivets and with straps under the source leads and at the edges of the PC-board on the in- and output side.

The printed circuit board has been attached to a solid copper plate (145 × 120 × 10 mm) which functions as a heatsink. Around the position of both transistors a tube has been soldered in the copper plate to controle the temperature by means of a watercooling system. For a good thermal contact between heatsink and transistors heatsink compound has been used.

Figure 15 shows the lay-out of the amplifier. The transformers have been fastened above the printed circuit board by means of accessories of Delrin material.

These accessories have been attached through the PC-board in the copper plate.

5   MEASURED PERFORMANCE

5.1   Single tone measurements

Figures 16 to 20 show at a constant outputpower of 300 W at 2 heatsink temperatures the gain, efficiency, input return losses, 2nd and 3rd harmonics at the output as a function of the frequency. In the range 1.6 to 28 MHz the gain is 22 to 23 dB, the efficiency 52.5 to 61%, the input return losses are below −15.5 dB, the second harmonics better then −25 dB and the third harmonics below −16 dB.

At a heatsink temperature of 70 °C the gain decreases about 1.5 dB. The heatsink temperature has only little influence on the other parameters. Figures 21 to 23 shows at 4 frequencies the output power as a function of the input power and the gain and efficiency versus outputpower.

Above 10 MHz the efficiency decreases about 6%. At 20 MHz the gain decreases above P_O = 200 W. At other frequencies this decrease starts at P_O = 300 W.

5.2   Two tone measurements

The two tone measurements have been carried out with 2 carriers with a frequency distance of 1 KHz. Figure 24 to 27 give as a function of the frequency the gain, efficiency, 3rd order distortion and 5th order distortion at 4 output levels. Over the whole frequency range the gain variation is less than 1 dB at each power level. At P_O = 300 W PEP the efficiency is at least 40%, the 3rd order distortion ≤−33 dB and the 5th order distortion ≤−38 dB.

Figures 28 and 29 give the 3rd and 5th order distortion versus output power at 4 frequencies.

To verify the choice of I_{DQ} = 1 A the 2nd and 3rd order distortion have been measured versus I_{DQ}. These measurements have been carried out at the most critical frequency and output level of 20 MHz and 30 W PEP resp.

Figure 30 shows that I_{DQ} = 1A for both transistors together was a good choice.

6   BALANCED CIRCUIT

As shown in Table 4 there is a certain amount of unbalance between both drain currents at RF operation. It is possible to improve this by using baluns in front of the input transformer and after the output transformer.

Table 4   Drain currents at P_O = 300 W

f (MHz)	I_{D1} (A)	I_{D2} (A)
1.6	5.2	4.8
5	5.1	4.75
10	5.25	4.8
15	5.3	4.95
20	5.7	5.3
25	6.2	5.25
30	5.85	5.15

7   CONCLUSIONS

This report shows that it is possible to design a wideband push-pull amplifier with 2 BLF177 MOS transistors having a very good performance.
The main properties are:

- Bandwidth: 1.6 to 28 MHz

- V_{DS}: 50 V

- I_{DQ}: 1 A

- Gain at P_O = 300 W: 22 to 23 dB

- Efficiency at P_O = 300 W: 52.5 to 61%

- Return losses input at P_O = 300 W: $\leq$−15.5 dB

- 2nd harmonics output at P_O = 300 W: $\leq$−25 dB

- 3rd harmonics output at P_O = 300 W: $\leq$−16 dB

- IMD at P_O = 300 W PEP: $\leq$−33 dB.

8   REFERENCES

G. Lukkassen
Application report NCO8602
A wideband power amplifier (25 to 110 MHz) with the MOS transistor BLF245.

n_1 = 8 turns
copper foil

PTFE foil

n_2 = 16 turns
enam. copper wire

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Fig.7 Output transformer.

$n_1 = n_2$ = 6 turns of 2 twisted enam. copper wires

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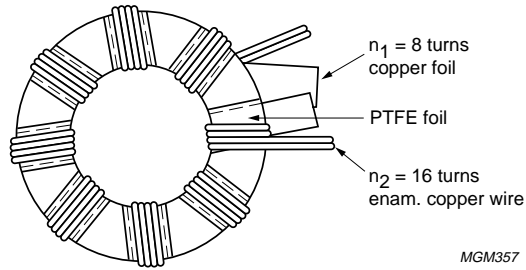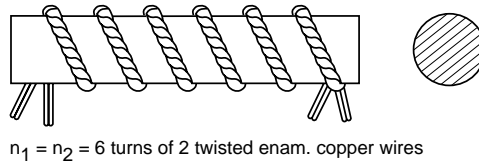
Fig.8 Tapped drain choke.

n_1 = 20 turns
enam. copper wire

PTFE foil

n_2 = 10 turns
copper foil

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Fig.9 Input transformer.

modified FXC choke
grade 3B

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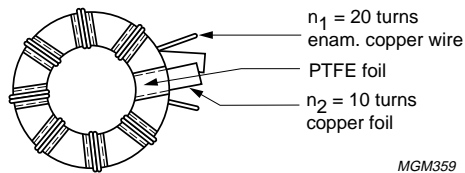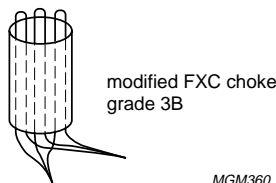
Fig.10 Decoupling choke.
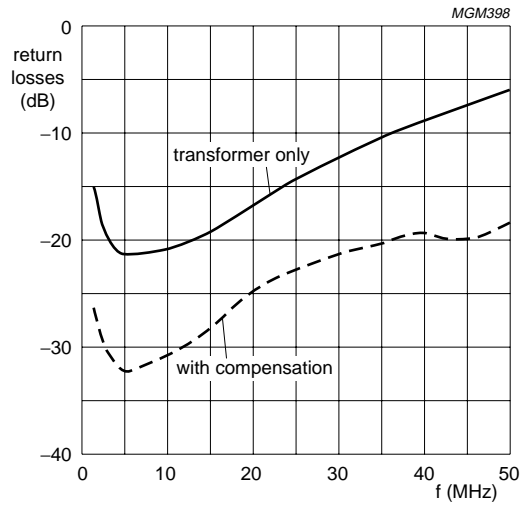
Fig.11 Output transformer correction.
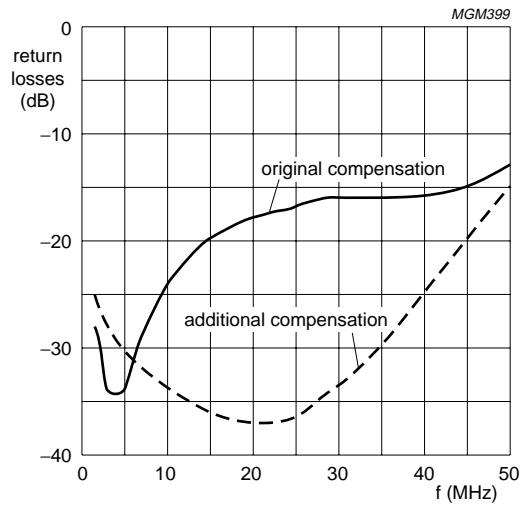
Fig.12 Return losses output circuit.
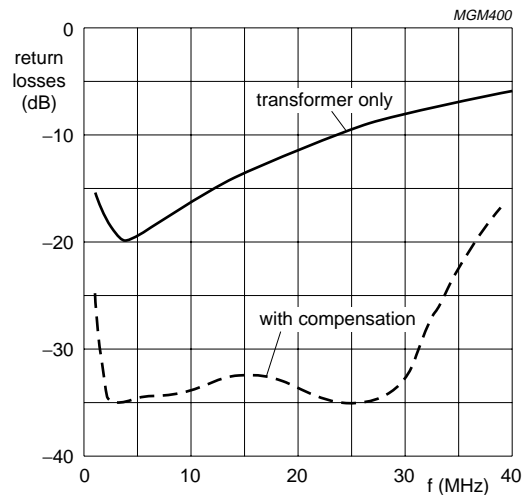
Fig.13 Input transformer correction.
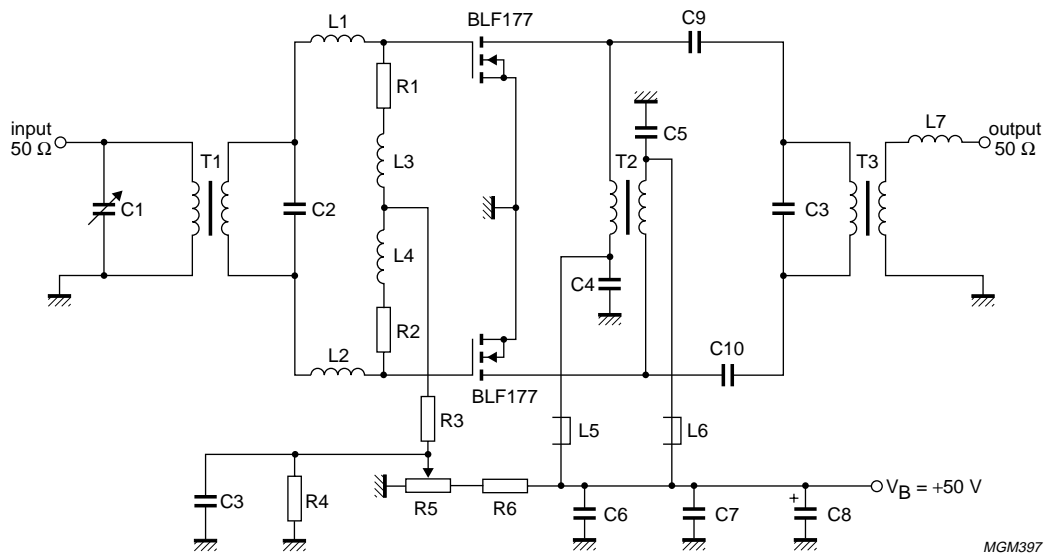
Fig.14 Circuit diagram of the 2 × BLF177 amplifier.

Table 5   Parts list of the wide band push-pull amplifier with 2 × BLF277 (1.6 to 28 MHz); note 1

C1	5 – 60 pF film dielectric trimmer (cat.nr.: 2222 809 08003)
C2	30 pF multilayer chip capacitor; note 2
C3	2 × 100 nF multilayer chip capacitor (cat.nr.: 2222 852 47104)
C4 = C5	3 × 100 nF metallized film capacitor (car.nr.: 2222 368 21104)
C6 = C7	100 nF multilayer chip capacitor (cat.nr.: 2222 852 47104)
C8	10 μF (63 V) electrolytic capacitor (cat.nr. 2222 030 28109)
C9 = C10	4 × 10 nF metallized film capacitor (cat.nr. 2222 368 51103)
C11	2 × 75 pF multilayer chip capacitor; note 2
L1 = L2	≈ 9 nH, printed inductance; l = 47 and w = 6 mm
L3 = L4	35 nH, 3 turns enamelled Cu-wire (0.7 mm) int.dia.: 3 mm, l = 2.35 mm
L5 = L6	2.2 μH, 1 turns through modified Ferroxcube choke grade 3B (cat.nr.: 4312 020 36642); see Fig.10
L7	100 nH, 5 turns enamelled Cu-wire (0.8 mm) int.dia.: 5 mm, 1 = 6.1 mm
R1 = R2	5.9 Ω; 4 metal film resistors of 23.7 Ω (0.4 W) in parallel (cat.nr.: 2322 151 72379)
R3	1 kΩ, metal film resistor (0.4 W) (cat.nr.: 2322 151 71002)
R4	1 MΩ, metal film resistor (0.4 W) (cat.nr.: 2322 151 71005)
R5	500 Ω, Cermet potentiometer (0.75 W)
R6	5.6 kΩ, metal film resistor (1 W) (cat.nr.: 2322 153 55622)
T1	input transformer: n_{pr} = 20 turns enamelled Cu-wire (0.5 mm) n_{sec} = 10 turns copper foil (width 2 mm), thickness 0.05 mm) wound around toroidal core, grade 4C6, dimensions: 14 × 9 × 5 mm (cat.nr. 4322 020 97181) see Fig.9
T2	drain choke: 6 turns of twisted pairs of 0.8 mm Cu-wires (each winding consists of 2 wires in parallel) wound on a Ferroxcube rod, grade 4B1, dimensions 10 × 50 mm, see Fig.8
T3	n_{pr} = 8 turns copper foil (width 6 mm, thickness 0.05 mm) n_{sec} = 16 turns of 2 enamelled Cu-wires (0.6 mm) in parallel wound around toroidal core, grade 4C6, dimensions: 36 × 23 × 15 mm (cat.nr. 4322 020 97201) see Fig.7; 2 of these transformers in parallel form the complete outputtransformer

Notes

1. PC-board: double Cu-clad, 1/16" epoxy fibre glass (ε_r = 4.5)

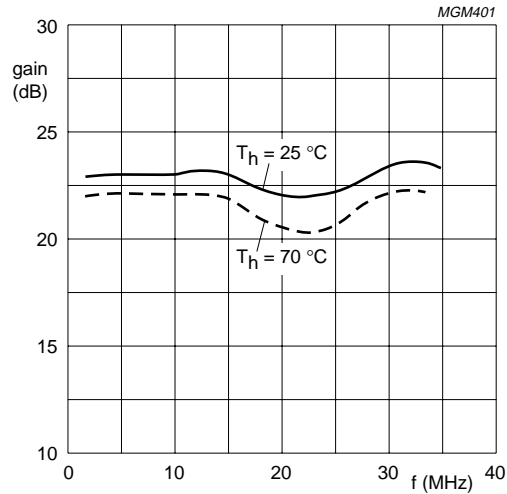2. American Technical Ceramics type 100B or capacitor of same quality.
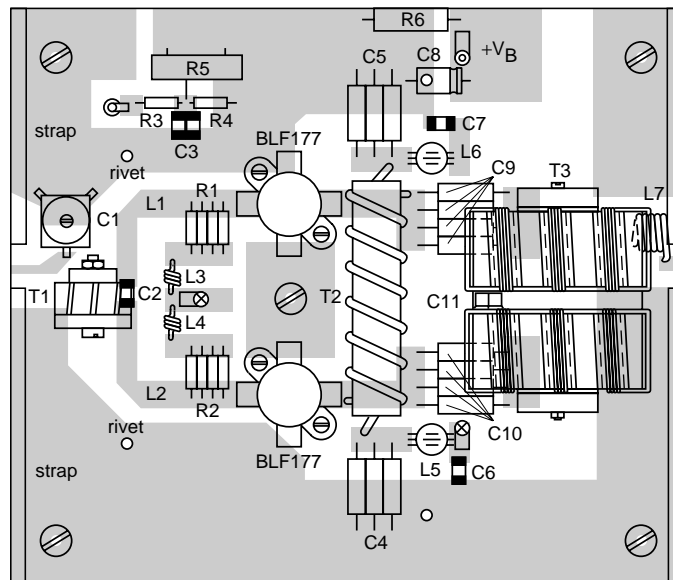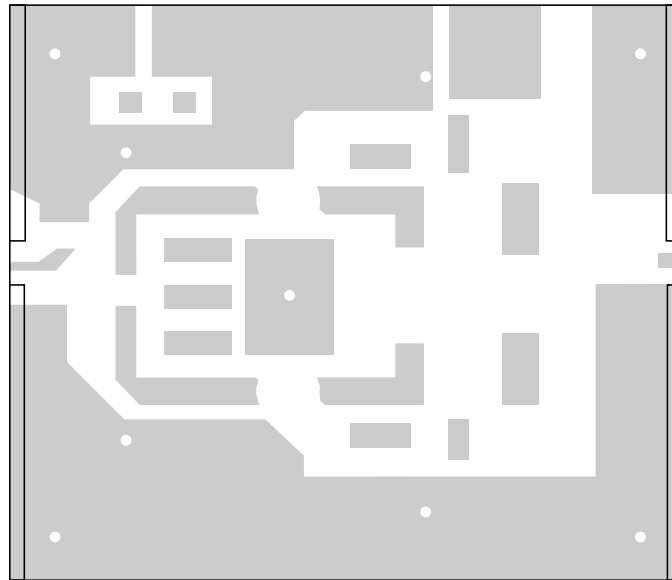
Fig.15 Lay-out of the 2 × BLF177 amplifier.
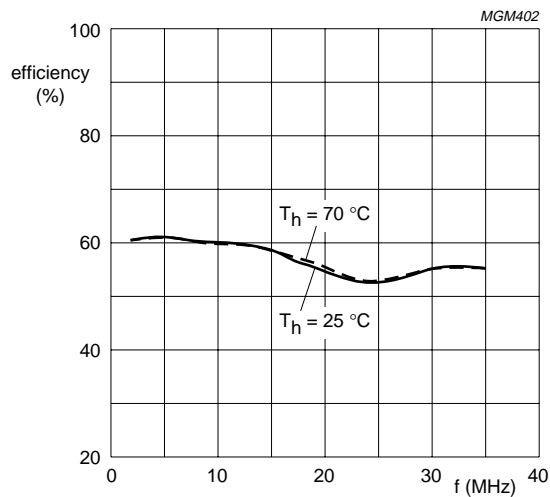
A wideband linear power amplifier (1.6 − 28 MHz)
for 300 W PEP with 2 MOS transistors BLF177

MGM401

2 × BLF177
V_{DS} = 50 V
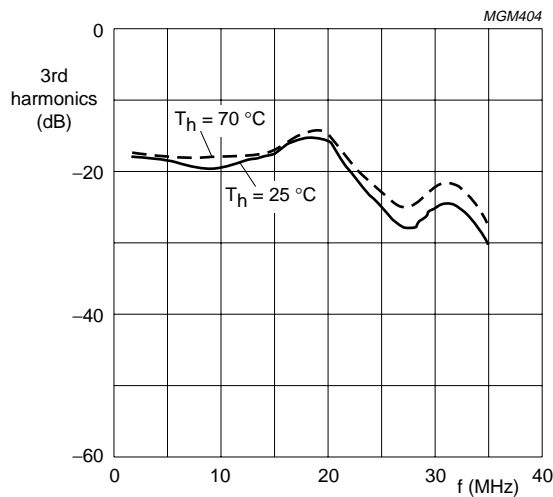I_{DQ} = 1A
P_O = 300 W

Fig.16 Gain versus frequency.

MGM402

2 × BLF177
V_{DS} = 50 V
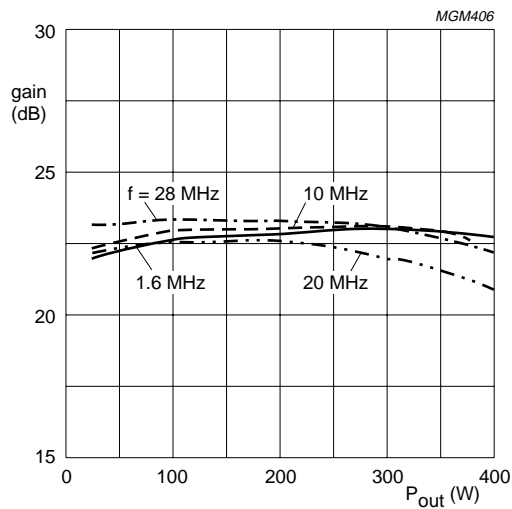I_{DQ} = 1A
P_O = 300 W

Fig.17 Efficiency versus frequency.

MGM403

return
losses
input
(dB)

$T_h = 25\ °C$ $T_h = 70\ °C$

f (MHz)
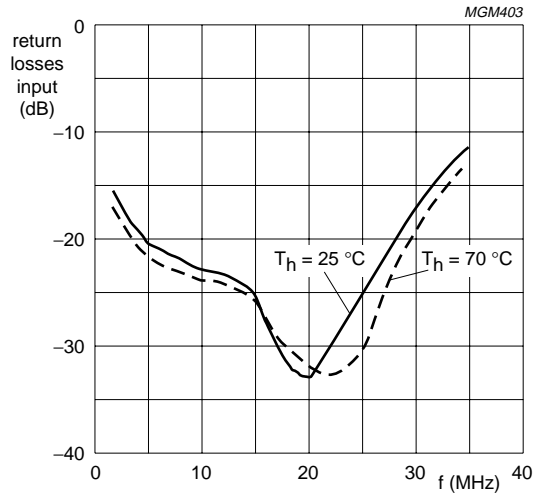
$2 \times$ BLF177.
$V_{DS} = 50$ V.
$I_{DQ} = 1A.$
$P_O = 300$ W.

Fig.18 Input return losses versus frequency.

MGM404

3rd
harmonics
(dB)

$T_h = 70\ °C$

$T_h = 25\ °C$

f (MHz)

$2 \times$ BLF177.
$V_{DS} = 50$ V.
$I_{DQ} = 1A.$
$P_O = 300$ W.

Fig.19 3rd harmonics versus frequency.

MGM405

2nd
harmonics
(dB)

T_h = 70 °C

T_h = 25 °C

f (MHz)
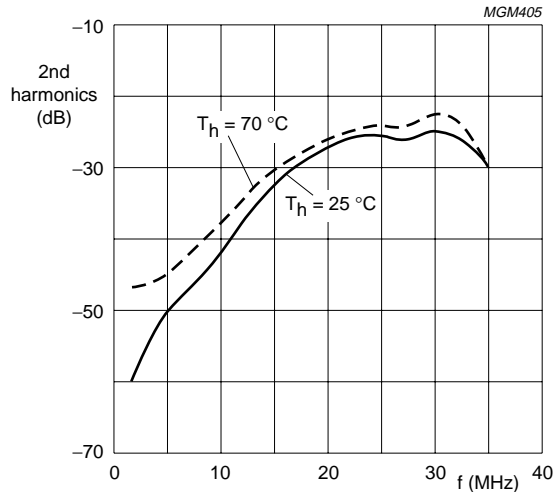
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.
P_O = 300 W.

Fig.20 2nd harmonics versus frequency.

MGM406

gain
(dB)

f = 28 MHz   10 MHz

1.6 MHz   20 MHz

P_{out} (W)

2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.

Fig.21 Gain versus outputpower.

MGM407

efficiency (%)

10 MHz

20 MHz

f = 1.6 MHz

28 MHz

P_{out} (W)
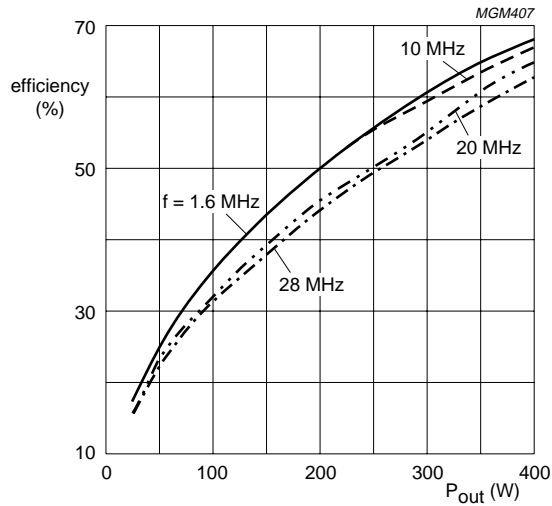
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.

Fig.22 Efficiency versus outputpower.

MGM408

P_{out} (W)

f = 1.6 MHz

28 MHz

20 MHz

10 MHz

P_{in} (W)
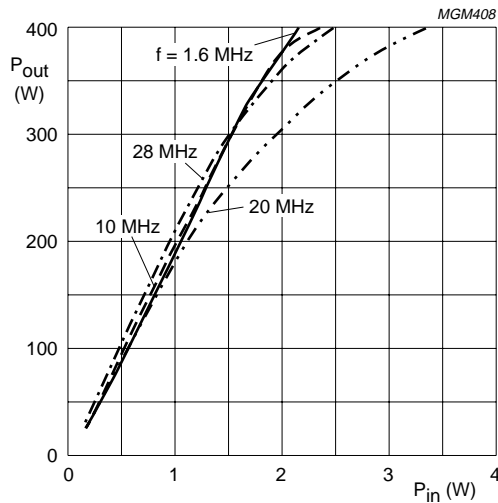
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.

Fig.23 Outputpower versus inputpower.
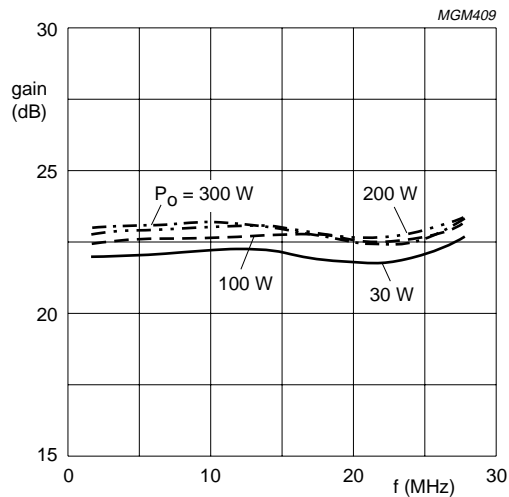
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.
$f_p − f_q$ = 1 kHz.

Fig.24 Gain versus frequency.
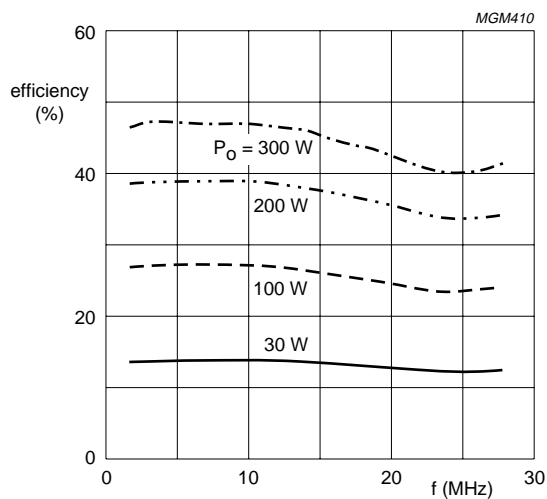
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.
$f_p − f_q$ = 1 kHz.
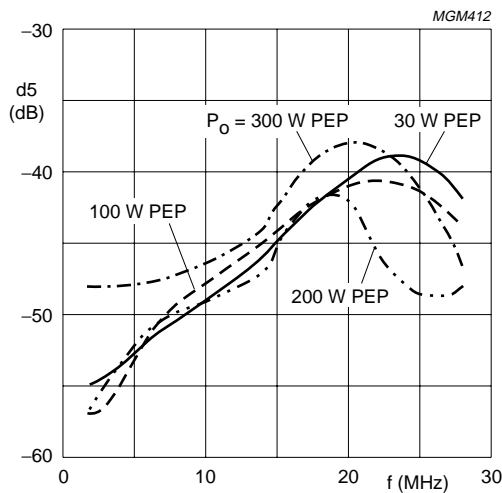
Fig.25 Efficiency versus frequency.

MGM411

d3
(dB)

P_O = 100 W PEP

300 W PEP

200 W PEP

30 W PEP

−30

−40

−50

−60

0   10   20   30
f (MHz)
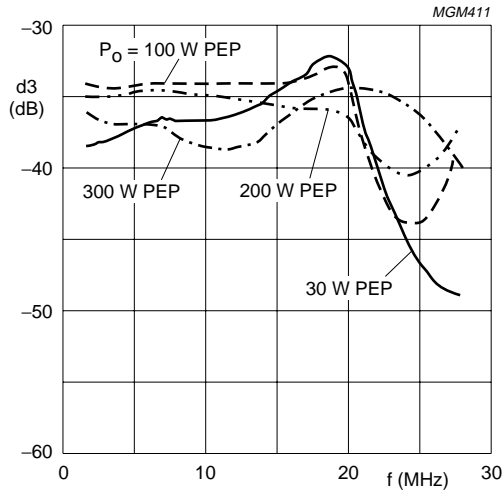
2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.
$f_p − f_q$ = 1 kHz.

Fig.26 Third order dist. versus frequency.

MGM412

d5
(dB)

P_O = 300 W PEP   30 W PEP

100 W PEP

200 W PEP

−30

−40

−50

−60

0   10   20   30
f (MHz)

2 × BLF177.
V_{DS} = 50 V.
I_{DQ} = 1A.
$f_p − f_q$ = 1 kHz.

Fig.27 Fifth order dist. versus frequency.

MGM413

d3
(dB)

f = 10 MHz

1.6 MHz

20 MHz

28 MHz

P$_O$ (W) PEP
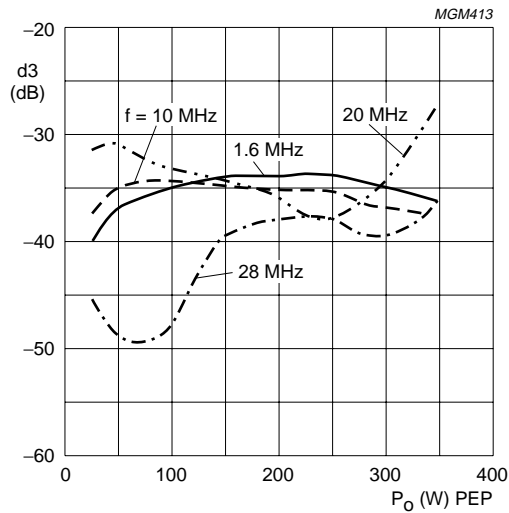
2 × BLF177.
V$_{DS}$ = 50 V.
I$_{DQ}$ = 1A.
f$_p$ − f$_q$ = 1 kHz.

Fig.28 3rd order distortion versus P$_O$.

MGM414

d5
(dB)

f = 20 MHz

10 MHz

28 MHz

1.6 MHz

P$_O$ (W) PEP
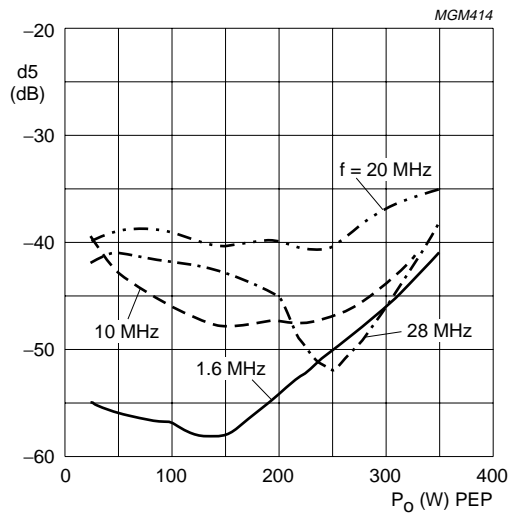
2 × BLF177.
V$_{DS}$ = 50 V.
I$_{DQ}$ = 1A.
f$_p$ − f$_q$ = 1 kHz.

Fig.29 5th order distortion versus P$_O$.

MGM415
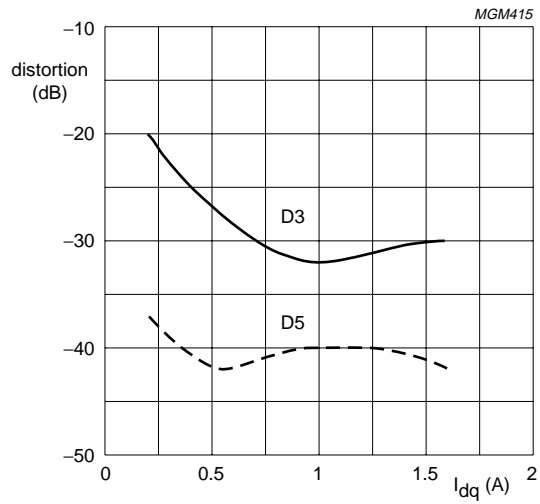
2 × BLF177.
V_{DS} = 50 V.
$f_p − f_q$ = 1 kHz.
f_p = 20 MHz.
P_O = 30 W PEP.

Fig.30 Distortion versus quiescent current.

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