APPLICATION NOTE

A wideband power amplifier (25 – 110 MHz) with the MOS transistor BLF245

NCO8602

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CONTENTS

1 SUMMARY

For military communication purposes a wideband class-AB power amplifier has been designed around the BLF 245 with the frequency range 25 to 110 MHz.

The DC-setting is V_D = 28 V and I_{DQ} = 200 mA.

In the input and output matching networks asymmetrical 1 : 4 transformers on 4C6 ferrite core material have been applied.

Table 1 The main properties are:

		UNIT
gain at P_O	17.7 ± 0.5	dB
bandwidth	25 – 110	MHz
V_D	28	V
I_{DQ}	200	mA
efficiency	55 – 67	%
input VSWR	≤1.6	

2 INTRODUCTION

The BLF245 is an RF power MOS transistor for the VHF frequency range in a SOT123 encapsulation.

For application in military communication equipment a wideband power amplifier has been developed with a frequency range from 25 to 110 MHz. The transistor operates in class-AB at V_{DS} = 28 V and a quiescent current I_{DQ} = 200 mA. The useful output power is in the range of 25 – 30 W.

3 DESIGN OF THE AMPLIFIER

3.1 General remarks

The amplifier has been developed with 1 : 4 impedance transformers in the input as well as in the output circuit. These transformers of the transmission line type with a ferrite core transform the 50 Ω system impedance at the input and output to about 12.5 Ω. An LC compensation circuit has been applied to transform this 12.5 Ω to the optimum load impedance of the transistor. At the input a circuit matches the 12.5 Ω to the gate impedance of the transistor and also takes care of a flat gain over the whole bandwidth.

3.2 Output circuit

For an optimum alignment of the output circuit the transistor has been replaced by a dummy. This dummy consists of a resistor of 12 Ω parallel with a capacitor of 82 pF. The real part of the dummy has been determined by the available drain voltage and the required output power.

$$R_L = \frac{V_D^2}{2P_O} \rightarrow R_L = \frac{28^2}{2.30} = 13.1\ \Omega$$

This is near to the value of 12.5 Ω mentioned in Section 3.1. The capacitor is about 15% higher than the output capacitance of the transistor. The RF choke at the drain side must have a sufficient high reactance at the lower end of the frequency range. Choosing this reactance appr. a factor 5 higher than the transistor loadresistance we get an inductance of 455 nH for L_4.

The output capacitance of the transistor can be compensated according to the Appendix. The result is: L_6 = 18.6 nH and C_{11} = 82 pF. To transform the achieved 12.5 Ω to the 50 Ω system impedance an asymmetrical 1 : 4 transformer has been used. Information about this kind of transformation can be found in Refs 1 and 2. For the transformer a toroid of 4C6 material has been used. Dimensions: 23 × 14 × 7 mm. On this toroid 5 turns of two 0.7 mm twisted enamelled Cu-wires are uniformly distributed and connected as shown in Fig.1.

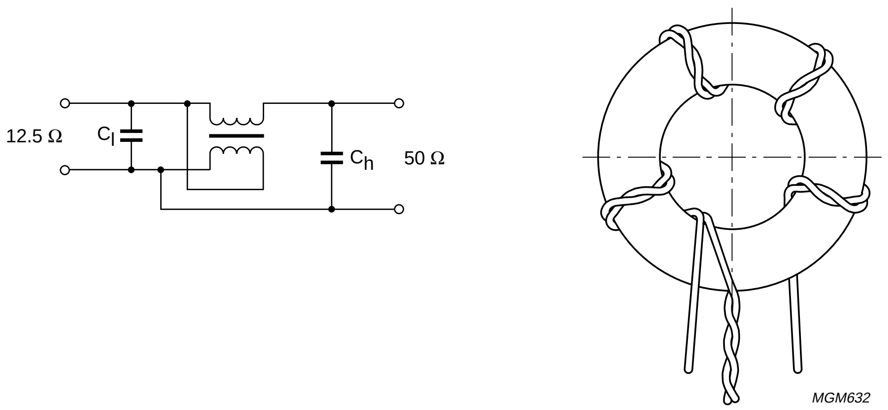

Fig.1 Output transformer.

With the aid of a network analyser the transformer has been corrected for higher frequencies. With C_l = 68 pF and C_h = 12 pF the return losses in the range 20 – 140 MHz are better than –30 dB (VSWR < 1.07). Optimization of the complete output circuit has been carried out by measuring the return losses at the output with the network analyser under swept condition (see Fig.2).

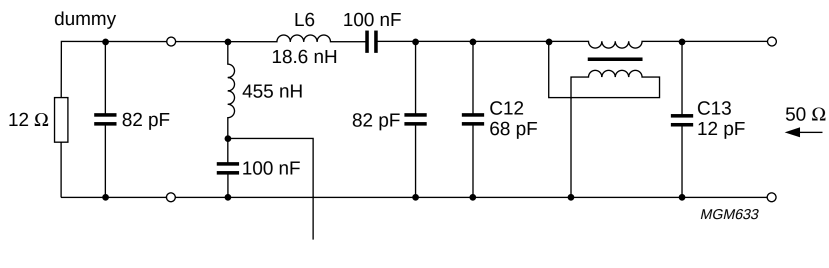

Fig.2 Output circuit before optimization.

Figure 4 shows the return losses of the output circuit before and after practical optimization. By decreasing L_6 to 10 nH and C_{12} to 43 pF the return losses improved about 10 dB in the frequency range 20 to 140 MHz to –20 dB (VSWR = 1.22).

3.3 Input circuit

As mentioned in Section 3.1 a special circuit matches the input impedance of the transistor to 12.5 Ω and also takes care of a sufficient flat gain over the whole bandwidth. To determine the gate-source impedance and the gain of the transistor in combination with the output circuit described in Section 3.2, narrow band input circuits have been used at several frequencies. By tuning such an auxiliary input circuit the gain of the transistor in combination with the output circuit can be measured directly. In case the input circuit has been tuned the output impedance of this circuit is the conjugate complex of the input impedance of the transistor.

Figs 5 to 7 give the input impedance and the gain of the transistor in combination with the output circuit. The matching network chosen at the input of the transistor is depicted in Fig.3.

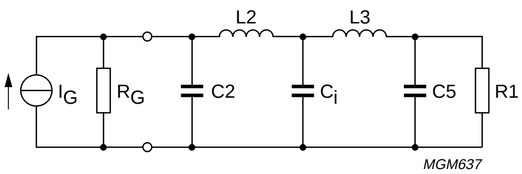

Fig.3 Input matching circuit.

C_i represents the input capacitance of the BLF245 which is appr. 220 pF (see Fig.6). Across this capacitor a constant voltage versus frequency from 25 up to 110 MHz has to be developed. Provided Ci is an ideal capacitance the optimum dimensioning of this network is as follows:

$R_G = R_1 = 1.6/(\omega_c \times C_i) = 10.5\ \Omega$

$C_2 = C_5 = 0.386\ C_i = 85$ pF

$L_2 = L_3 = 0.997\ R_1/\omega_c = 15.1$ nH.

in which ω_c is the maximum angular frequency. The calculated voltage variation across C_i is ±0.36 dB and the maximum VSWR seen by the generator is 1.36. Deviating from this calculation, for the ease of transformation, R_G and R_1 have been chosen 12.5 Ω. Further the resistive component of C_i is substantial especially at higher frequencies.

Therefore the values of the components have been changed in a computer optimization program for a maximally flat gain and a low input VSWR. This optimization results in a gain of 17.5 dB with a variation of ±0.17 dB and a maximum VSWR = 1.177. These results have been achieved by changing the components of Fig.3:

$C_2 = 97$ pF, $C_5 = 102$ pF, $L_2 = 17.6$ nH, $L_3 = 29$ nH and $R_1 = 12\ \Omega$.

The remaining part of the transformation from 12.5 Ω to the 50 Ω system impedance has been accomplished with a transformer similar to the output transformer. However the input transformer has been wound on a core consisting of 2 small toroids of 4C6 material (6 × 4 × 2 mm).

On this core 6 turns of two 0.25 mm twisted enamelled Cu-wires are uniformly distributed similar to the output transformer described in Section 3.2. (see Fig.1). With correction capacitors at the high ohmic and the low ohmic side of respectively 8.2 and 47 pF the return losses in the range 20 – 140 MHz are better than –27 dB (VSWR ≤ 1.1).

For the practical optimization of the complete inputcircuit the transistor has been adjusted at V_D = 28 V and a quiescent current I_{DQ} = 200 mA. The gain and input return losses have been measured in the frequency range of 20 up to 110 MHz.

The best results have been achieved by changing the correction capacitor C_3 from 47 to 62 pF and by executing R_1 as a parallel connection of 5 resistors of 61.9 Ω.

Figure 8 gives the complete circuit diagram of the BLF245 wideband amplifier and Table 3 gives the corresponding parts list.

4 MEASURED PERFORMANCE

4.1 Constant input power

Figs 9 to 11 give the gain, efficiency and output power versus the frequency at a constant input power (P_i = 0.5 W).

In the frequency range of 25 to 110 MHz the gain is 17.2 to 17.9 dB, the efficiency 55 to 70% and the output power 26.5 to 30.5 W.

4.2 Constant output power

Figs 12 and 13 give the gain and efficiency versus the frequency at a constant power (P_O = 27.5 W) and heatsink temperatures of 25 and 70 °C.

Figs 14 and 15 give the input return losses and the 2e and 3e harmonics of the output signal also versus the frequency. The return losses have been measured at a heatsink temperature of 25 and 70 °C. The harmonics have been measured at 25 °C. By increasing the heatsink temperature from 25 to 70 °C the gain decreases about 1.2 dB. The heatsink temperature has no influence on efficiency and return losses. At 25 °C the gain of the amplifier varies from 17.2 to 18.2 dB, the efficiency from 55 to 67% and the return losses at the input are at least –14 dB (VSWR ≤ 1.6). Also the 2e and 3e harmonics are at least 14 dB down.

4.3 Constant frequency

Figs 16 to 18 give the output power versus input power and the gain and efficiency versus power at 4 frequencies.

4.4 Stability

Applying an R&S PTU low pass filter at the output of the amplifier stability measurements have been carried out. Choosing a low pass frequency as close as possible above the measuring frequency the amplifier was stable through the whole frequency range of 25 to 110 MHz.

4.5 Mismatch

The amplifier has been tested for load mismatch at all phase angles. Up to VSWR = 10 : 1 the amplifier is stable. At VSWR = 20 : 1 the amplifier is only stable below 70 MHz. However also at higher frequencies degradation of the RF performance did not occur.

5 CONCLUSIONS

Based on the results presented in this report it may be concluded that it is quite possible to design a wideband amplifier from 25 to 110 MHz with a very good performance using the MOS transistor BLF 245.

Table 2 The main properties are:

		UNIT
Bandwidth	25 – 110	MHz
V_D	28	V
I_{DQ}	200	mA
Gain (P_O = 27.5 W)	17.7 ± 0.5	dB
Efficiency	55 – 67	%
Input VSWR	≤1.6	

6 REFERENCES

Ref.1.
A.H. Hilbers
Application report ECO6907: Design of HF wideband Power Transformers.

Ref.2.
A.H. Hilbers
Application report ECO7703: Power Transformers for the Frequency Range 30 – 80 MHz

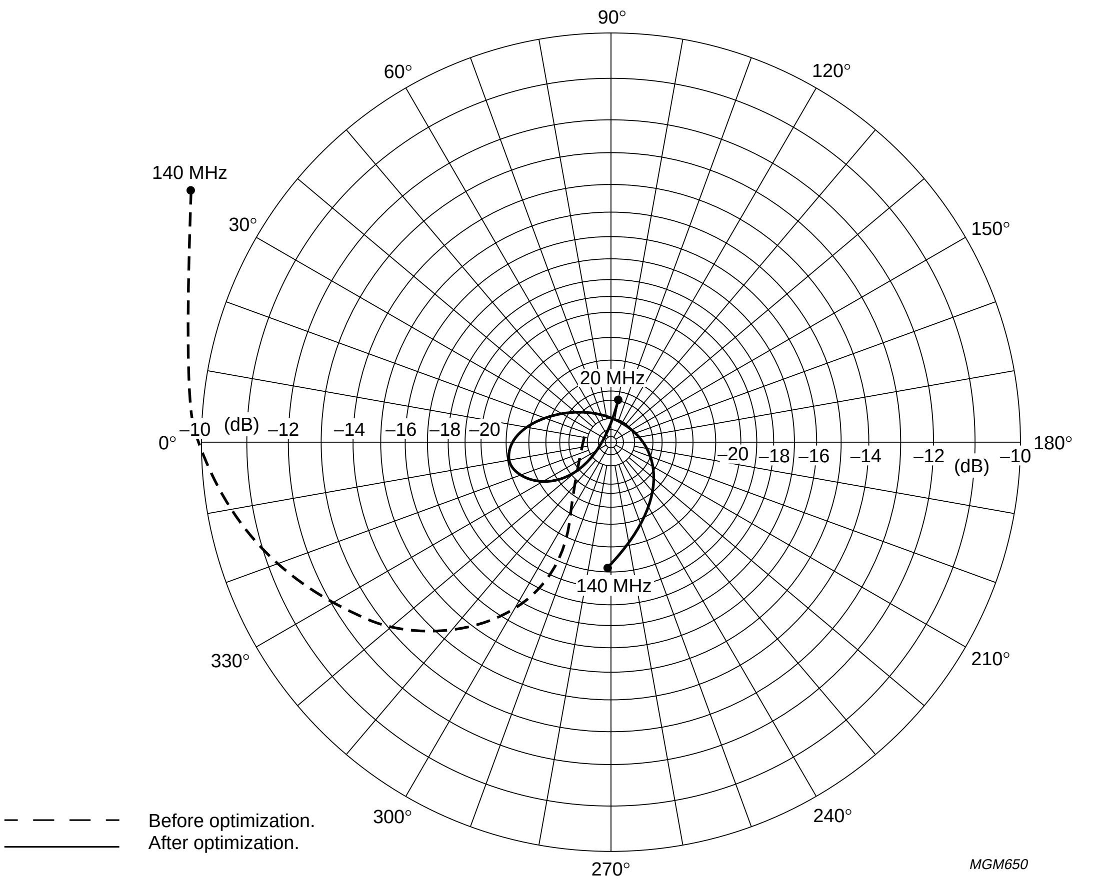

Fig.4 Return losses output circuit.

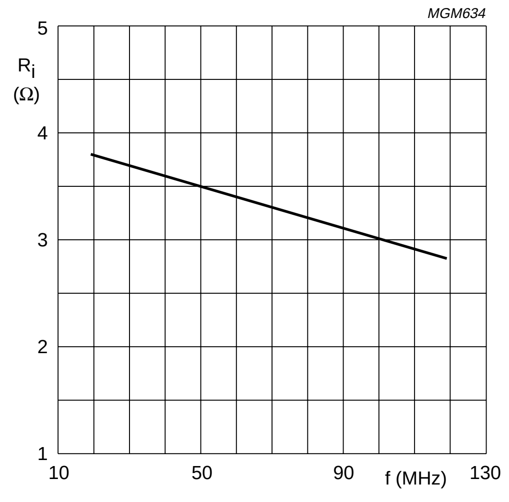

Fig.5 Real part of input impedance of loaded transistor.

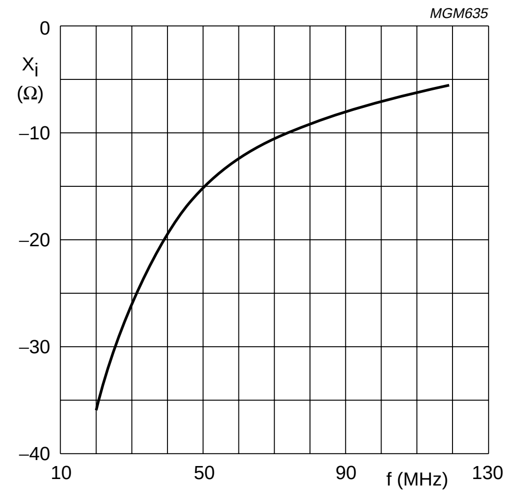

Fig.6 Imaginary part of input impedance of loaded transistor.

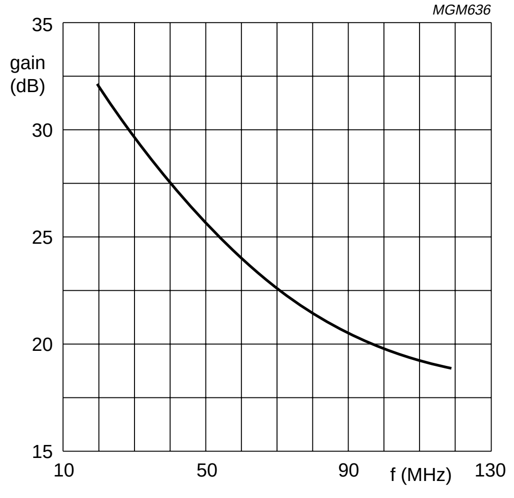

Fig.7 Gain of loaded transistor.

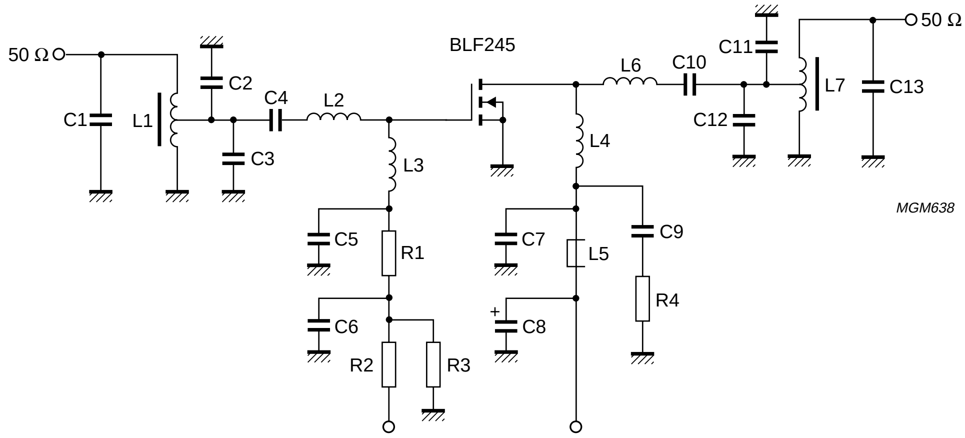

Fig.8 Circuit diagram of the BLF245 wideband amplifier.

Table 3 Parts list of the BLF245 wideband amplifier

WIDEBAND POWER AMPLIFIER WITH BLF245 (f = 25 – 110 MHz)	
C1 = 8.2 pF multilayer ceramic chip capacitor; note 1	
C2 – C5 = 100 pF multilayer ceramic chip capacitor; note 1	
C3 = 62 pF multilayer ceramic chip capacitor; note 1	
C4 = C10 = 10 nF multilayer ceramic chip capacitor	cat. no. 2222 852 47103
C6 = C7 = 100 nF multilayer ceramic chip capacitor	cat. no. 2222 852 47104
C8 = 2.2 uF electrolytic capacitor	
C9 = 3 × 100 nF multilayer ceramic chip capacitor	cat. no. 2222 852 47104
C11 = 82 pF multilayer ceramic chip capacitor; note 1	
C12 = 43 pF multilayer ceramic chip capacitor; note 1	
C13 = 12 pF multilayer ceramic chip capacitor; note 1	
L1 = 2 Ferroxcube toroids, grade 4C6 (6 × 4 × 2 mm) with 6 turns of 2 × 0.25 mm twisted enamelled Cu-wire (see Fig.1)	cat. no. 4322 020 97160
L2 = 17.6 nH, 2 turns enamelled Cu-wire (0.6 mm) int.dia.: 3 mm, length 2.5 mm, leads 2 × 5 mm	
L3 = 28.8 nH, 3 turns enamelled Cu-wire (0,6 mm) int.dia.: 3 mm, length 3.2 mm, leads 2 × 5mm	
L4 = 455 nH, 12 turns enamelled Cu-wire (1 mm) int.dia.: 7 mm, length 16.5 mm, leads 2 × 5 mm	
L5 = Ferroxcube h.f.choke, grade 3B	cat. no. 4312 020 36642
L6 = 10 nH, 1 turn enamelled Cu-wire (1 mm) int.dia.: 3 mm leads 2 × 3 mm	
L7 = Ferroxcube toroid, grade 4C6 (23 × 14 × 7 mm) with 5 turns of 2 × 0.7 mm twisted enamelled Cu-wire (see Fig.1)	cat. no. 4322 020 97190
R1 = 12.4 Ω, parallel connection of 5 metal film resistors 61.9 Ω	cat. no. 2322 151 76199
R2 = 1 KΩ, metal film resistor	cat. no. 2322 151 71002
R3 = 1 MΩ, metal film resistor	cat. no. 2322 151 71005
R4 = 10 Ω, metal film resistor	cat. no. 2322 153 51009
Printed-circuit board: double Cu-clad, 1.6 mm epoxy fibre-glass (ε_r = 4.5)	

Note

1. American Technical Ceramics type 100B or capacitor of same quality.

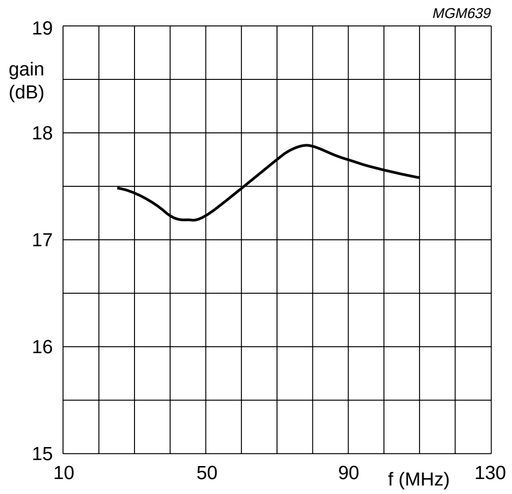

Fig.9 Gain at P_i = 0.5 W.

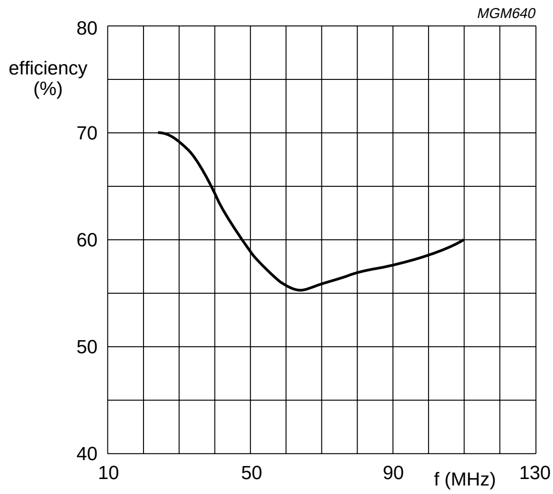

Fig.10 Efficiency at P_i = 0.5 W.

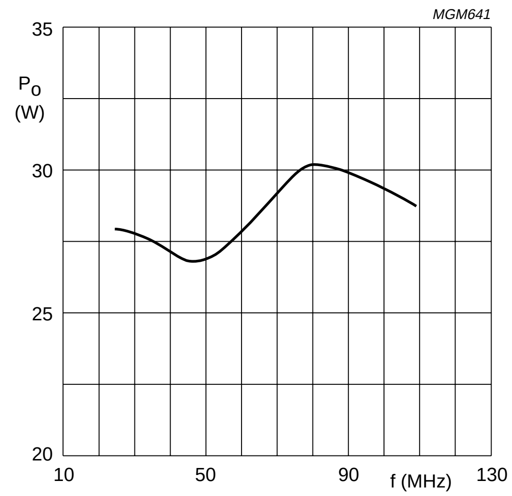

Fig.11 Output power at P_i = 0.5 W.

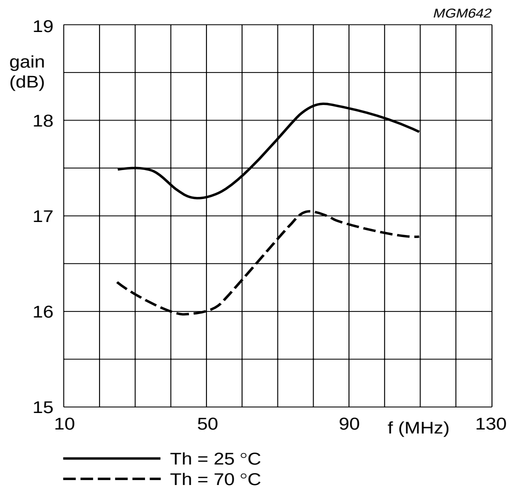

Fig.12 Gain at P_O = 27.5 W.

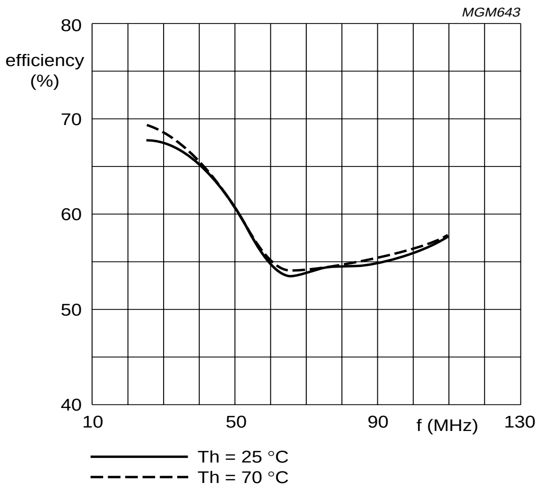

Fig.13 Efficiency at P_O = 27.5 W.

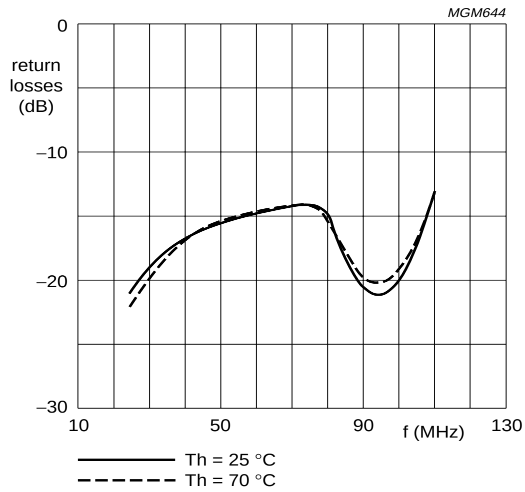

Fig.14 Input return losses at P_O = 27.5 W.

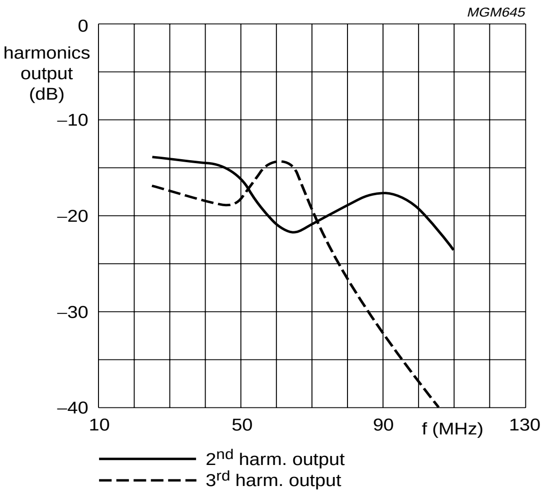

Fig.15 Output harmonics at P_O = 27.5 W.

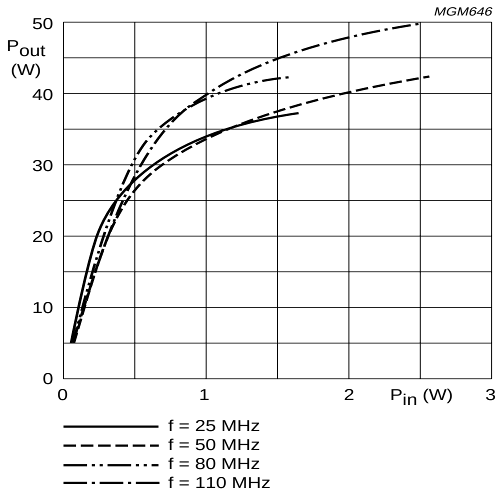

Fig.16 Output power versus input power.

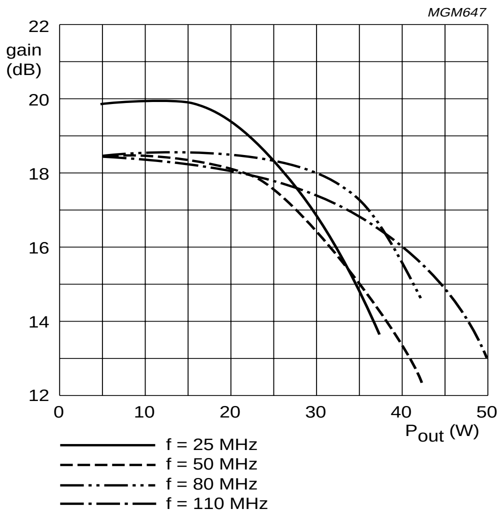

Fig.17 Gain versus output power.

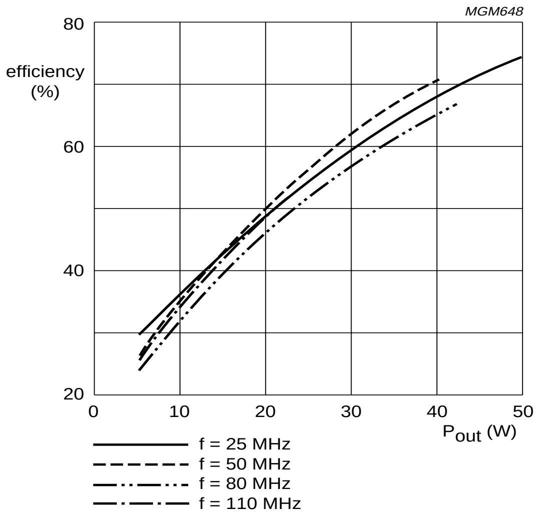

Fig.18 Efficiency versus output power.

7 APPENDIX

The output capacitance of a transistor can be compensated over a certain bandwidth by absorbing it in a low-pass Chebyshew π-section.

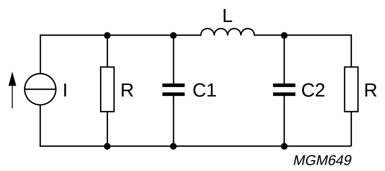

Fig.19

If C1 is the transistor output capacitance the components L and C2 must be added.
C2 = C1 = C

The normalized value of C is: $A = \omega_m CR$
In which $\omega_m = 2\pi f_{max}$

Now we can calculate the normalized value of L with:
$B = 8A/(3A^2 + 4)$
where $B = \omega_m L/R$

The maximum VSWR of this network can be calculated with the following procedure.

1. Determine $\gamma = \frac{1}{A}$

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

3. $\text{VSWR} = \left\{\frac{X^3 + 1}{X^3 - 1}\right\}^2$

In our amplifier:
R = 12.5 Ω
C = 82 pF

This gives:
A = 0.784
B = 1.029
L = 18.62 nH
γ = 1.412
X = 3.142
VSWR = 1.138

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