APPLICATION NOTE

100 – 450 MHz 250 W Power Amplifier with the BLF548 MOSFET

AN98021

Philips
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1 INTRODUCTION

In this report the design procedures and measurement results are given of a two octave wideband amplifier (covering both the civil and military airbands between 100 and 450 MHz), equiped with two MOSFET devices, which is capable of generating 250 W of output power.

In order to achieve a good broadband capability one has to use devices with the output capacitance reduced to the utmost minimum. While applying PHILIPS' BLF548 MOSFETs it was possible to obtain a respectable powergain of more than 10 dB, throughout the whole band.

The BLF548 is a balanced N-channel enhancement mode vertical D-MOS transistor in a SOT262 package, especially designed for use in wideband amplifiers up to 500 MHz. The transistor is capable to deliver 150 W nominal outputpower at a supply voltage of 28 Volts. Due to the low output capacitance the attainable bandwidth will exceed 300 MHz.

2 DESIGN CONSIDERATIONS

While designing broadband amplifiers, one has to take several things into account:

- To select the right manufacturer, able to supply the products with a good reliability, gives a good support and offers a complete range of transistors e.g. for driverstages.
- To select the right active components, capable to fulfill the desired wishes, such as; high reliability, high powergain, high efficiency, excellent mismatch capabilities, right loadpower, good long-life properties and last but not least; good broadband capability.
- To terminate the transistor with the right load impedance, with other words, to determinate the right output matching network.
- To eliminate the 6 dB/octave gain slope throughout the band of operation, in order to achieve an acceptable gainflatness.
- To find the right input matching network; the input VSWR has to be low in order to achieve a good termination for the driverstage.
- To design the matching networks in such a way that they are capable to handle the, at some points very high, R.F. currents.

A balanced transistor was chosen in order to reduce the second harmonic (due to the push-pull effect) and to reduce the number of required components.

The criteria for chosen MOSFETs over bipolar transistors are; high powergain, high load mismatch capabilities, low noise and easy biasing.

Nowadays three major MOSFET suppliers are involved when Pl = 150 W is needed at f = 500 MHz and Vds = 28 V. Available are; BLF548, industry type A and industry type B. Table 1 gives an overview of the characteristics of these 3 types.

Table 1

	BLF548	TYPE A	TYPE B	UNIT
f	500	400	500	MHz
Gp	>10	>10	>8	dB
ηd	>50	>50	>55	%
Ciss	105	180	140	pF
Coss	90	200	100	pF
Crss	25	20	32	pF
BW	300	133	266	MHz

With: BW = $1/(2\pi * Rload * Co)$; Rload = $Vds^2/(2 * Pl)$ and Co = 1.15 * Coss.

BW = bandwidth, Rload = loadresistance, Co = outputcapacitance.

In order to achieve the best possible broadband results, the BLF548 is a very good choice.

Other Philips MOSFETs in the 500 MHz series are, followed by nominal loadpower:

12.5 Volts – single ended

BLF521	2 W
BLF522	5 W

28 Volts – single ended

BLF542	5 W
BLF543	10 W
BLF544	20 W

28 Volts – push-pull

BLF544B	20 W
BLF545	40 W
BLF546	80 W
BLF547	100 W
BLF548	150 W

3 AMPLIFIER CONCEPT

The amplifier concept described in this paper is based upon two identical modular units, each containing one BLF548 MOSFET. Both units are combined by means of two 3 dB – 90° hybrid couplers, which is shown in Fig.1. The main advantage is that the input VSWR will be very good; since it is independent of the mismatch introduced by the units, the 50 Ω termination will cause a good load for the driver stage, e.g. equipped with BLF544.

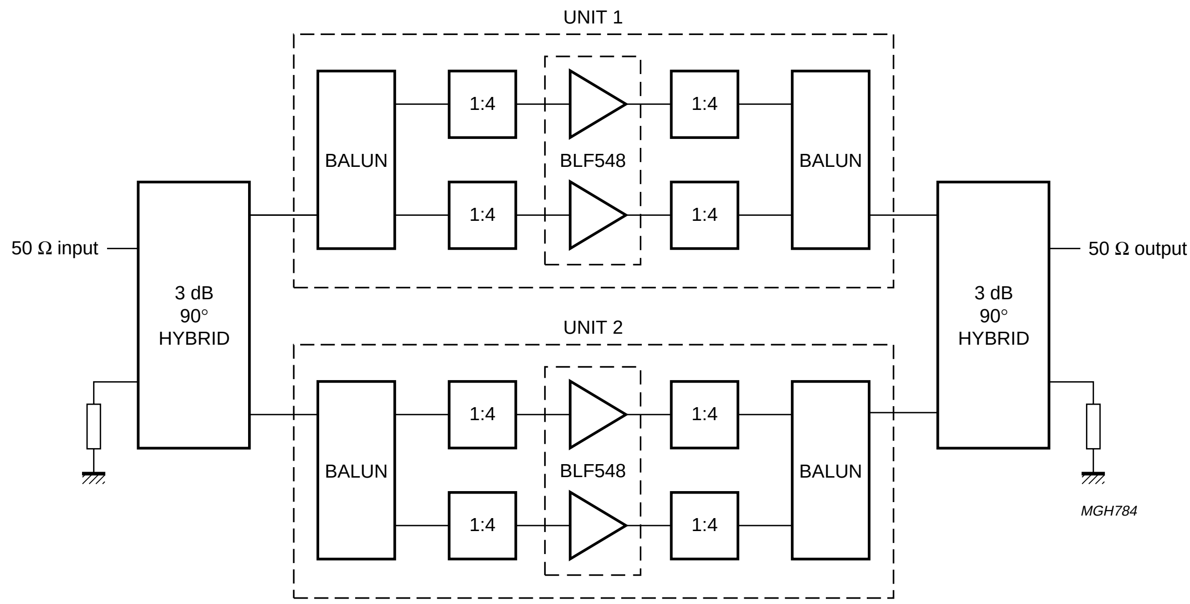


Fig.1 Schematical representation of the concept of the eventual amplifier.

The following seven steps have been followed in order to develop a first prototype of one unit.

1. Determine the BLF548’s 150 W output power load impedance between 100 and 500 MHz (50 MHz interval steps) by measurement techniques or simulations. At the moment of writing it was not possible to perform full automatic measurements at frequencies lower than 500 MHz with transistors build in a balanced SOT262 header. Therefore the load impedances have been calculated by means of the electrical equivalent diagram shown in Fig.2.
2. Find the correct output matching network which transforms the 50 Ω termination to the required load impedance for the frequency range 100-500 MHz.
3. Optimize the outputmatching network of step 2 with help of linear simulation software, such as Touchstone (EESOF).
4. Since the matching network will not have an ideal behaviour, it is necessary to determine the actual load impedance of the selected output matching network, again in 50 MHz steps between 100-500 MHz.
5. Calculate (or even better, determine by means of load-pull measurements) both the powergain and input impedance of the transistor by presenting the load impedences, found at step 4, to it. This is very important to investigate the behaviour of the transistor while terminating it with the selected output matching network.
6. Choose the right input matching network which has a minimum returnloss (Rl) at the highest frequency (450 MHz) and a declining Rl for lower frequencies in a way that the gain increase effect for lower frequencies is equalized. Other possibilities, as feedback or frequency dependent damping at the gate side (by means of low Rgs), can be taken into consideration.
7. Optimize the input network for gainflatness by means of linear simulation software (Touchstone, EESOF). Remember the input VSWR throughout the band is taken care of by the use of 90° hybrids, which combine the two modular units.

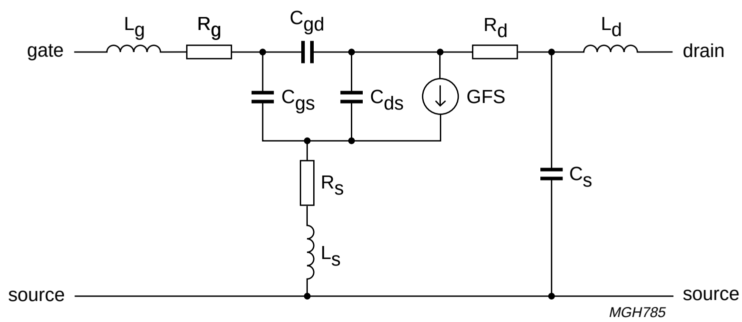


Fig.2 RF Power MOSFET equivalent diagram (one BLF548 section).

At the following pages the design steps are presented which were followed at PHILIPS' laboratories in order to design a 150 W unit. Using the diagram shown in Fig.2, powergain and impedances have been calculated first, using the data given in Table 2.

Table 2

Lg	0.58 nH	
Ls	0.11 nH	
Ld	0.50 nH	
Rg	0.09 Ω	
Rs	0.08 Ω	
Rd	0.19 Ω	
Cgd	29 pF	1.15 × Crss
Cgs	120 pF	1.5 × (Ciss-Crss)
Cds	72 pF	1.15 × (Coss-Crss-Cs)
Cs	2.4 pF	2.4 pF
Gfs'	1.6 S	0.5 × Gfs (for Class B)

Rg, Rd, Rs are derived from Rdson measurements, Gfs and Cs are measured, Cgs, Cds, Cgd derived from measured Ciss, Coss, Crss respectively. Lg, Ls and Ld are calculated.

Some of the assumptions are based on empirical rules and have proven to be correct in the past.

Gp and Zin can now be calculated:

$Zin = Ri + jXi$

$Gp = 10 * {}^{10}\log (Gfs' \times Rload/\omega^2 \times Ls \times Ci)$

with;
$Xi = \omega \times Li - 1/(\omega \times Ci)$
$Ri = (Gfs' \times Ls)/Ci$
$Li = Lq + (Ls \times Cgs)/Ci$
$Ci = Cgs + Cgd\ (1 + Gfs' \times Rload)$
$w = 2\ \pi f$

Zload is chosen for maximum broadband capability.

Table 3 Calculated powergain, Zin and required Zload (series components)

F (MHz)	PL (W)	Gp (dB)	ZIN (Ω)	ZLOAD (Ω)
100	78.8	26.7	0.43 – j4.1	4.7 + j1.5
150	78.8	23.3	0.43 – j2.5	4.0 + j1.0
200	78.8	20.8	0.42 – j1.7	3.4 + j2.0
250	78.8	18.4	0.43 – j1.1	2.8 + j1.9
300	78.8	17.2	0.43 – j0.7	2.3 + j1.7
350	78.8	15.8	0.43 – j.0.3	1.9 + j1.4
400	78.8	14.5	0.44 – j0.0	1.6 + j1.1
450	78.8	13.4	0.44 + j0.2	1.3 + j0.7
500	78.8	12.4	0.45 + j0.5	1.1 + j0.4

The data is also given in datahandbook *“RF power MOS transistors”* – Philips Components. It can be noticed that without any gaincompensation the powergain difference between 100 and 500 MHz will exceed 10 dB.

To terminate the transistor with the required loadimpedance, with respect to the broadband capability, the unbalanced 50 Ω load has to be transformed as close as possible to the loadimpedance as shown in Table 3. (Note: the impedances shown are based on one section, since the transistor is of a balanced type, Zin and Zload are related to virtual ground).

To reduce the number of components which would be needed in case of a lumped element solution, a coaxial semi-rigid balun is used to transform the unbalanced 50 Ω load into two 25 Ω sections that are 180° apart in phase and 90° away from virtual ground. This is followed by a coaxial 4 : 1 transformer, with a characteristic impedance of 25 Ω.

The result of this is: $Rp = (\sqrt{25 \times 25})/4 = 6.2\ \Omega$, which is close to the required Rload of the transistor.

In order to give a good description of the outputnetwork, it will be described as a 3-port: one port terminated with 50 Ω unbalanced, the other two terminated with the transistor’s outputimpedance (the complex conjugate of loadimpedance). A computer listing of the outputnetwork is given in “Appendix A”. After optimizing the network to minimum returnloss (S11), while checking S13, the optimized return loss (in dB) of this network has been determined, see Fig.3. As a next step now the difference between the required and the network related loadimpedance can be (re-) calculated. The result on powergain (Gp) and imputimpedance (Zin) is given in Table 4.

Table 4 Result on Zin and Gp as a result of the presented outputmatching network

F (MHz)	PL (W)	Gp (dB)	ZIN (Ω)	ZLOAD (Ω)
100	78.6	32.7	0.11 – j4.1	4.3 + j1.0
150	78.8	19.7	1.04 – j2.9	4.1 + j0.2
200	78.8	17.3	1.02 – j2.2	3.3 + j0.1
250	78.8	16.0	0.87 – j1.5	2.9 + j0.1
300	78.8	14.9	0.79 – j0.9	2.8 + j0.3
350	78.8	13.5	0.80 – j.0.4	2.8 + j0.4
400	78.8	12.4	0.79 – j0.1	2.5 + j0.2
450	78.8	12.0	0.67 + j0.1	1.9 + j0.1
500	72.0	12.4	0.43 + j0.5	1.1 + j0.6

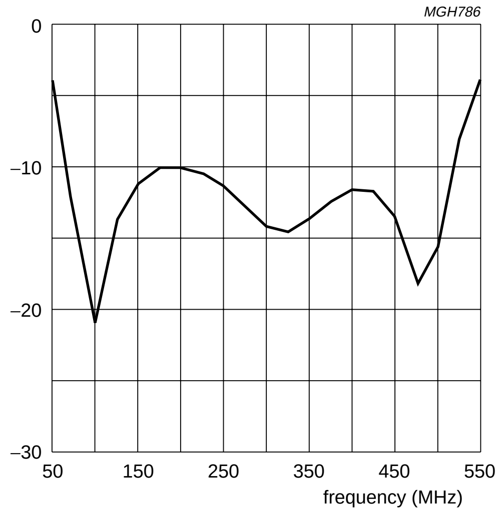


Fig.3 Simulated network response (output side).

4 INPUT CIRCUITRY

Since Zin and Gp are now determined in a accurate way, the inputcircuitry can be determined. Special attention is given to the flatness of the gain as a function of frequency. The input network also consists of a coaxial balum, followed by a 1 : 4 coaxial transformer, both made of semi-rigid coaxial cable. Since Zin is rather low the characteristic impedance of the 1 : 4 transformer was chosen to be 10 Ω.

The result of this is: Rp = $(\sqrt{25} \times 10)/4 = 3.9\ \Omega$.

In order to compensate for the 6 dB/octave slope, matching to Zin is achieved at 450 MHz. At lower frequencies a mismatch is created, resulting in a decrease of powergain inversely proportional to the increase of the gain related to the transistor's 6 dB/octave slope.

The network listing of the input circuitry, again presented as a 3-port, is given in "Appendix B". The network response (both input returnloss and predicted powergain) is given in Fig.4. Finally the schematic diagram and list of components are given in Fig.5. The unit's layout is given in Fig.6. Note: two toroidal cores around T2 and T3 are used to prevent oscillations.

5 ADJUSTMENT OF THE AMPLIFIER

5.1 Tuning the outputnetwork

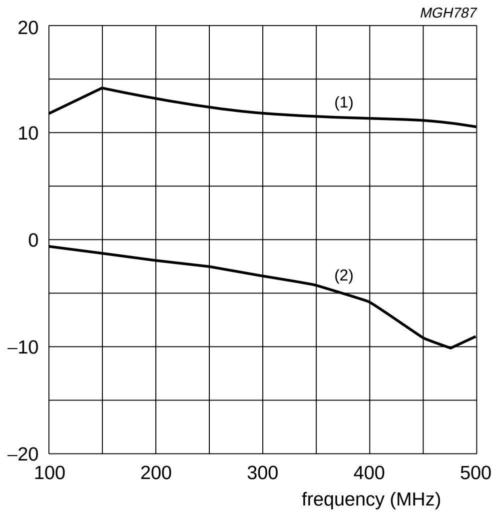
In order to terminate the transistor with the proper load impedance, first the output network has to be tuned.

The transistor was replaced by a dummyload, representing the transistors output impedance under full power conditions. The dummyload was realized after fitting the data of Table 3. To the dummyload model (roughly Rload in parallel with Coss, in series with draininductance Ld). Later the model was compensated for parasitics of both SOT262 header and network components.

Initial settings for each side of the dummyload are:

$Rload = Vds^2/2 \times Pl = 5.2\ \Omega$
$C = 1.15 \times Coss = 104\ pF$
$L = Ld = 0.5\ nH$

The network listing is given in "Appendix C". The final result, the dummyload lay-out, is given in Fig.7.



(1) DB [S12].
(2) DB [S22].

Fig.4 Simulated network response of inputside (predicted Gp = f(f)).

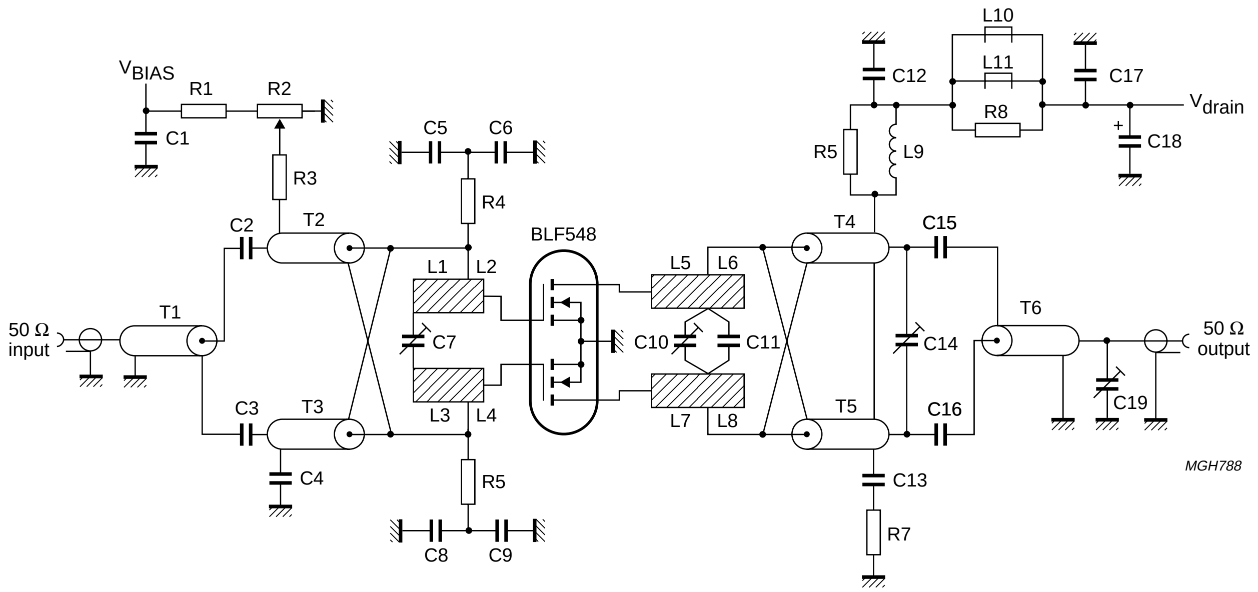


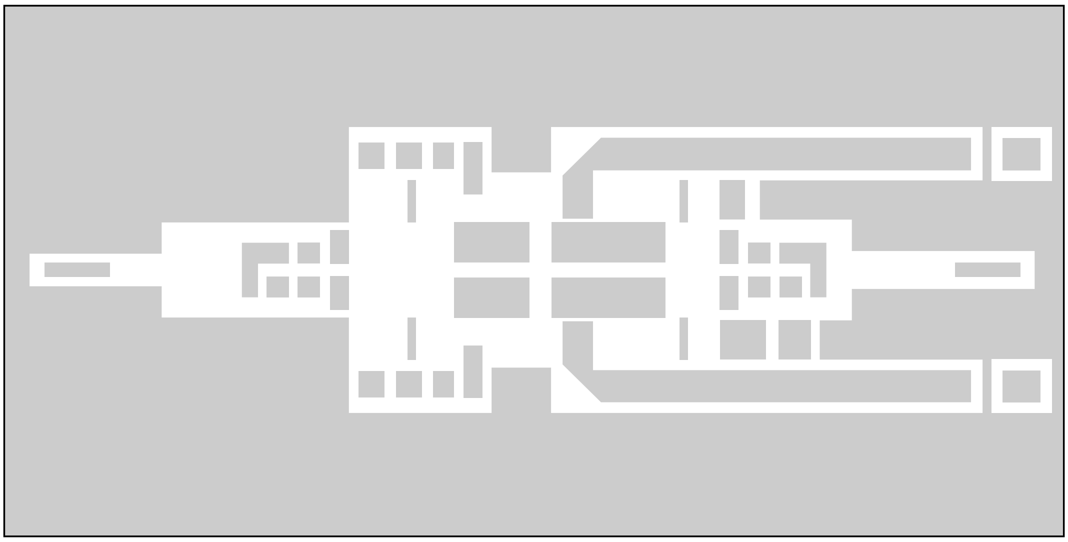
Fig.5 Schematical diagram and list of components of one unit.

List of components

DESIGNATION	DESCRIPTION	VALUE	DIMENSIONS	CATALOGUE NO.
C1, C17	multilayer ceramic chip capacitor	100 nF		2222 852 47104
C2, C3	multilayer ceramic chip capacitor (note 1)	47 pF		
C4, C5, C8	multilayer ceramic chip capacitor (note 1)	820 pF		
C6, C9	multilayer ceramic chip capacitor (note 1)	300 pF		
C7	film dielectric trimmer	2-18 pF		2222 809 09006
C10, C14	film dielectric trimmer	2-9 pF		2222 809 09005
C11	multilayer ceramic chip capacitor (note 2)	39 pF		
C12	capacitor	22 nF		
C13	capacitor	100 nF		
C15, C16	multilayer ceramic chip capacitor (note 1)	120 pF		
C18	63 V electorlytic capacitor	1 µF		2222 685 78108
C19	film dielectric trimmer	1-5 pF		222 808 09004
L1, L3	stripline (note 3)	20 Ω	5 × 8 mm	
L2, L4	stripline (note 3)	20 Ω	2.5 × 8 mm	
L5, L7	stripline (note 3)	20 Ω	11.5 × 8 mm	
L6, L8	stripline (note 3)	20 Ω	4 × 8 mm	
L9	5 turns enamelled Cu wire on R6		1.4 mm	
L10, L11	grade 3B Ferroxcube wideband RF choke			4330 030 36642
T1	semi-rigid coax (note 4)	50 Ω	length 54 mm	
T2, T3	semi-rigid coax (note 4)	10 Ω	length 44 mm	
T4, T5	semi-rigid coax	25 Ω	length 53 mm	
T6	semi-rigid coax	50 Ω	length 74 mm	
R1	0.4 W metal film resistor	19.6 kΩ		2322 151 11963
R2	10 turn potentiometer	5 kΩ		2122 362 00725
R3, R4, R5	0.4 W metal film resistor	2.05 kΩ		2322 151 12052
R6, R7, R8	1.0 W metal film resistor	10 Ω		2322 153 71009

Notes

1. American Technical Ceramics type 100B or capacitor of same quality.
2. American Technical Ceramics type 175B or capacitor of same quality.
3. The striplines are on a double copper-clad PCB with P.T.F.E. fibre-glass dielectric (ε_r = 2.2); thickness 1/32 inch.
4. T2 and T3 are equipped with a Toroidal core, grade 4C6 (cat.no. 4322 020 97171).



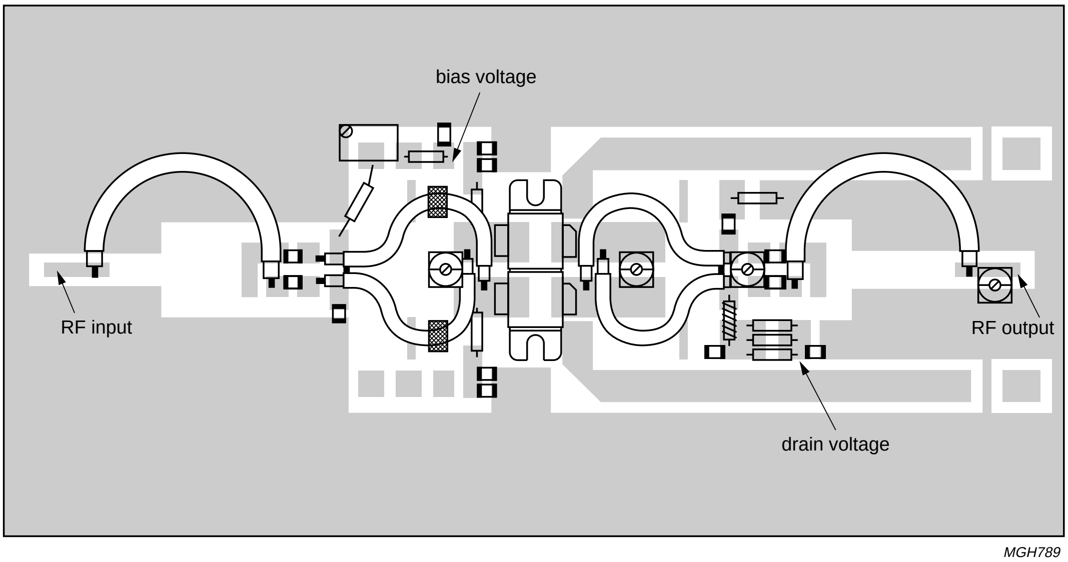


Fig.6 Lay-out of one unit.

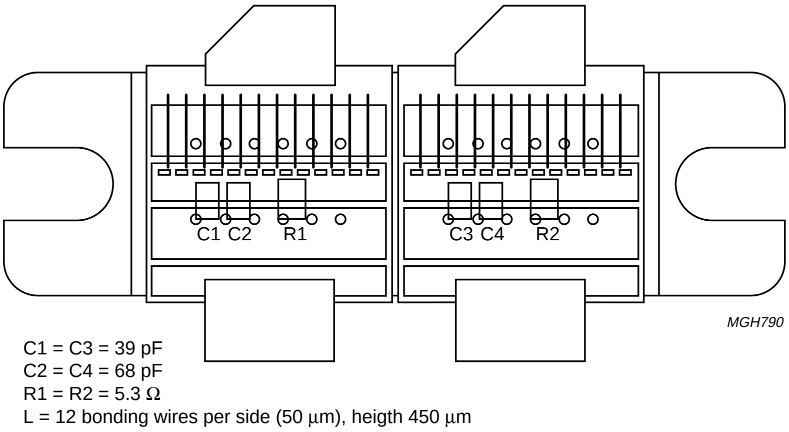


Fig.7 Lay-out of BLF548 dummyload.

By means of a RF-analyzer the predicted frequency response of the network can be reproduced in practice, while tuning C10 and C14 for optimum R1. This is presented in Fig.8. A comparison with the simulated networkresponse (Fig.3) shows a high amount of common behaviour.

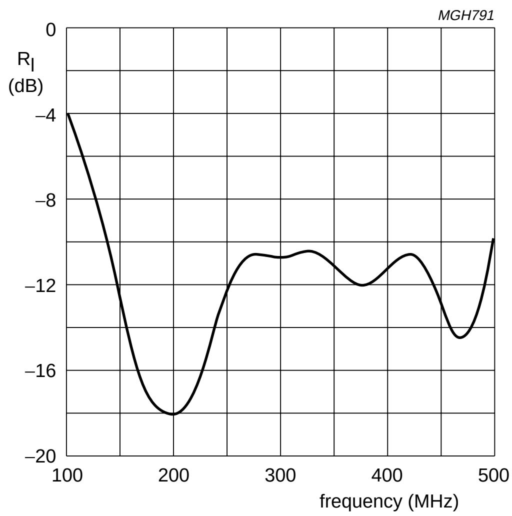


Fig.8 Measured network response of one unit after tuning C10 and C14.

5.2 Testing the unit under RF conditions

After exchanging the dummyload for a BLF548, a frequencysweep under power conditions can be made with help of a network analyzer. The used measurement set-up is given in Fig.9.

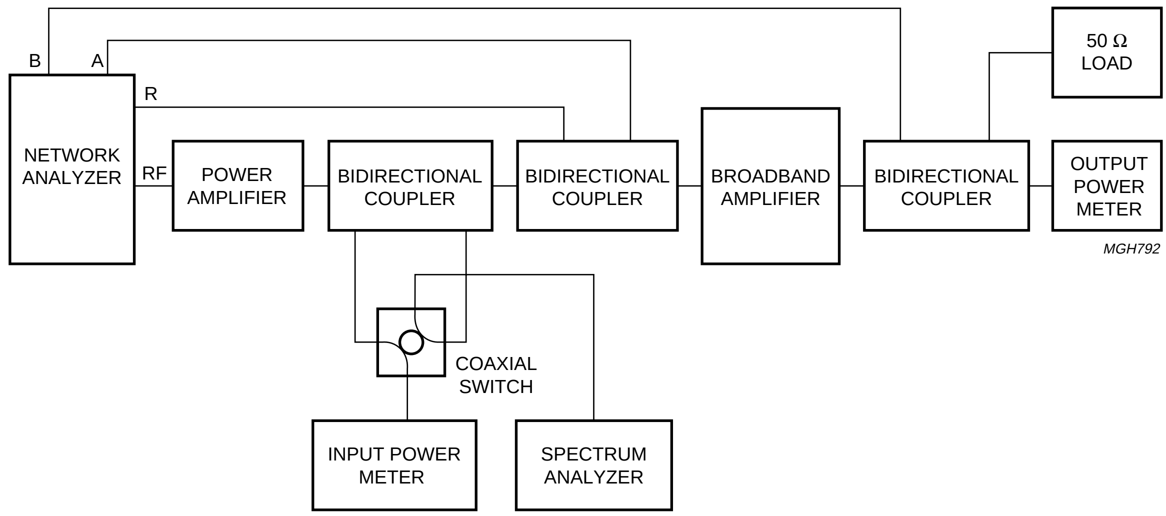
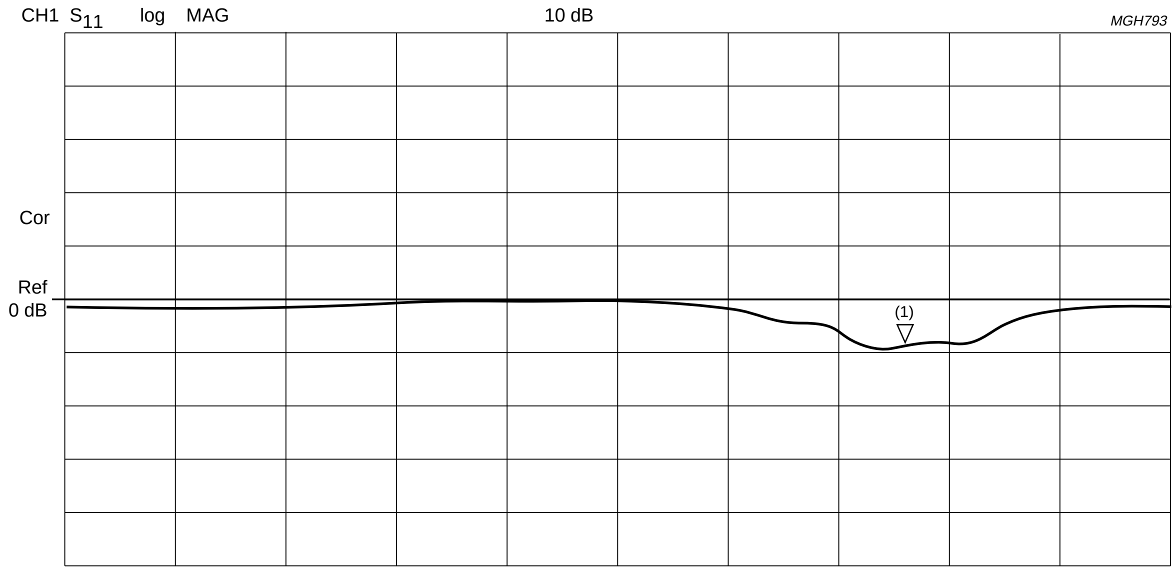


Fig.9 Measurement set-up for frequency sweep under powerconditions.

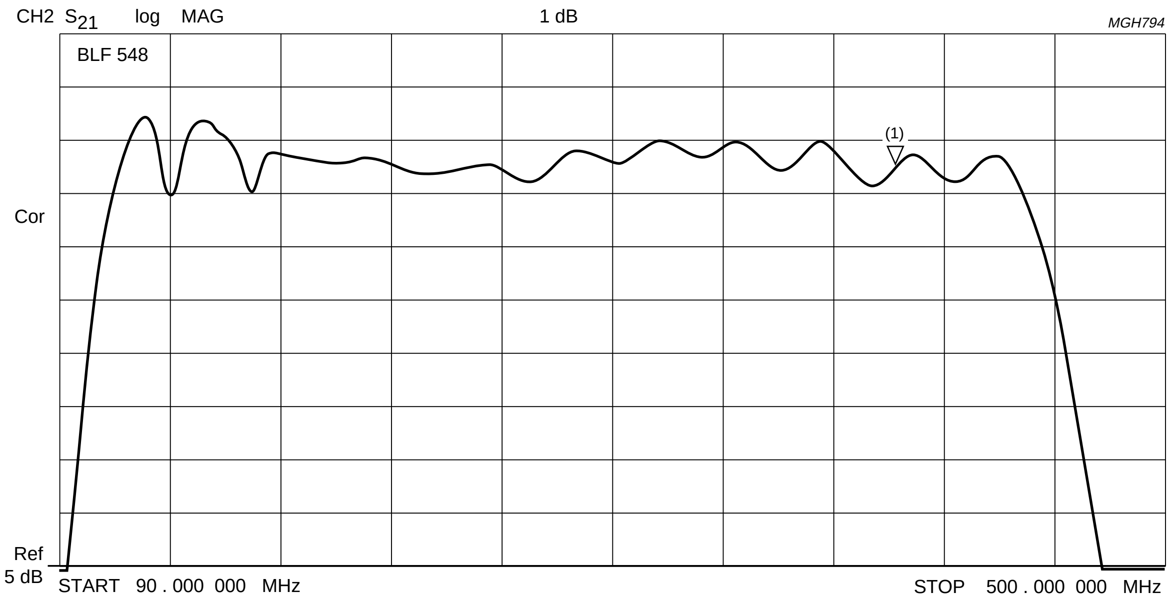
5.3 Tuning the unit’s inputnetwork

After supplying both the bias- and drain voltages to the unit and adjusting the drain quiescent current with R2 to 320 mA (160 mA per side), the inputpower (Ps) is applied. Gainflatness is optimized while tuning C7. The unit’s input returnloss is given in Fig.10. Pl versus frequency is shown in Fig.11. A comparison with the simulated network response of the inputside (Fig.4) shows a high similarity. Gainflatness within 1 dB is achieved between 100 and 450 MHz.



(1) 400 MHz; –8.3 dB.

Fig.10 Input return loss of one unit (Frequency sweep at Pl = 150 W, for f = 90 to 500 MHz.

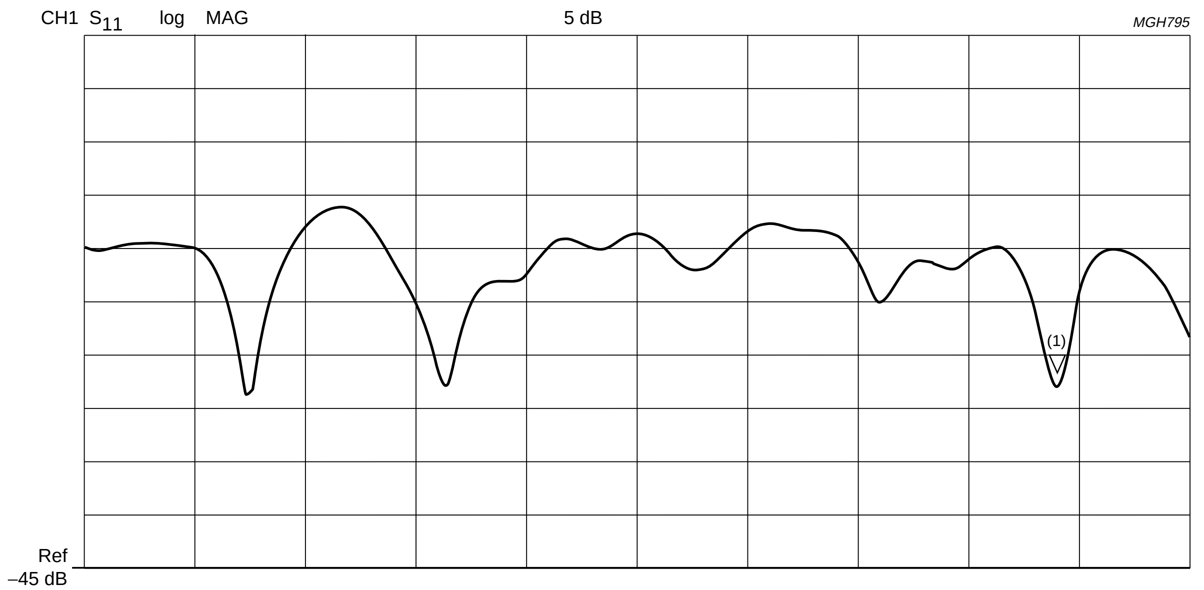


(1) 400 MHz; 12.526 dB.

Fig.11 Powergain of one unit (Frequency sweep at Pl = 150 W, for f = 90 to 500 MHz).

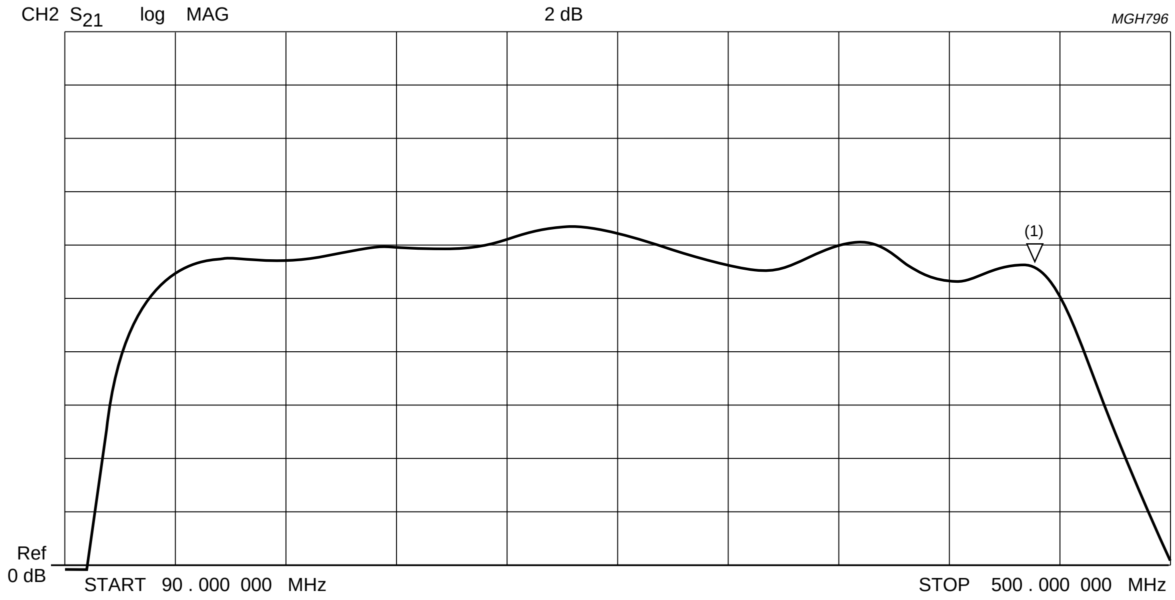
5.4 Combining the units

After tuning the second unit similar as described above, the connection was made to the 90° hybrid couplers. The couplers do contain 4 ports, one at which the input signal is applied (1). The input is devided equally into two ports (3 and 4). Between ports 3 and 4 there is a voltage lag of 90°. Mismatch at ports 3 and 4 do not effect the VSWR of port 1, since port 2 is terminated with a 50 Ω load (KDI-PPT820-75-3 flange mounted). At the output side of the amplifier the units are combined in a similar way. Both input and output hybrids and 50 Ω loads are mounted in a Brass baseplate (dimensions; 200 × 160 × 10 mm), which also serves as a heatspreader for both BLF548 devices. The baseplate is connected to a heatsink which is cooled by means of forced air. Final results are given in Fig.12. (input return loss of the amplifier) and Fig.13 (the amplifier's powergain).



(1) 450 MHz; –28.1 dB.

Fig.12 Input return loss of complete amplifier (Frequency sweep at Pl = 250 W, for f = 90 to 500 MHz).



(1) 450 MHz; 11.1 dB.

Fig.13 Powergain of complete amplifier (Frequency sweep at Pl = 250 W, for f = 90 to 500 MHz).

6 CONCLUSIONS

The described procedures shown in this paper, are a great help in designing high-power broadband amplifiers. The differences between theory and practice are relatively small.

The BLF548 is very well suited to perform in multi-octave broadband UHF-amplifiers; at a supply voltage of 28 V, between 100 and 450 MHz, 250 W of outputpower could be generated with a powergain of 11 dB (gainripple smaller than 1 dB). Drainefficiency is 45 to 55% throughout the band. The reduction of the second harmonic is more than 25 dB, with respect to the fundamental. The input returnloss is better than –12 dB.

7 REFERENCES

- Data Handbook SC08b, RF power MOS transistors – Philips Components
- Application Report Bipolar & MOS transmitting transistors – Philips Components
- A look inside those integrated two-chip amps – Joe Johnson – Microwaves feb. 1980
- Apply wideband techniques to balanced amplifiers – Lee B. Max – Microwaves apr. 1980
- Demystifying new generation silicon high power FETs – Steve McIntyre – Microwave Journal apr. 1984
- Anaren – Microwave components catalog no.17A.

8 APPENDIX A

DIM

FREQ	MHz
RES	OH
IND	NH
CAP	PF
LNG	MM

VAR

C1 = 120	!c = 270
L2 = 74	!50
L3 = 53	!50
L10 = 9	!9
L11 = 0.350559	!0.5
R11 = 4.191123	!5.1
C11 = 64.2821 4	!75
L12 = 0.409981	!0.31
W1 = 6	
L22 = 10	
L33 = 66	

CKT

RES	1 0	R = 50	
DEFIP	1	REFIMP	
S1PA	2 0	BLF548OU	
DEF1P	2	BALI	
IND	1 2	L^L11	
RES	2 0	R^R11	
SLC	2 0	L^L12; C^C11	!to determine zload
DEFIP	1	BAL3	
SLC	1 0	L = 0.5 C = 4 !13.5	
COAX	1 2 0 3	DI = 0.91; DO = 2.98; L^L3; ER = 2.03	TAND = 0.0002; RHO = 1
MSUB	ER = 2.2; H = 0.79	T = 0.035	RHO = 0.72; RGH = 1
SLC	2 4	L = 0.5; C^cl	
SLC	3 5	L = 0.5; C^cl	
COAX	4 6 0 8	DI = 1.63	DO = 2.95; L^L2 ER = 2.03; TAND = 0.0002; RHO = 1
COAX	5 8 0 6	DI = 1.63	DO = 2.95; L^L2; ER = 2.03; TAND = 0.0002; RHO = 1
SLC	6 8	L = 0.5; C = 41	!C = 41
SLC	6 8	L = 0.7; C = 8	!C = 0
MLIN	919	w^w1; 1^122	
MLIN	10 20	w^w1; 1^122	
MBEND3	19 29	w^w1	
MBEND3	20 30	w^w1	
MLIN	29 0	w^w1; 1^133	
MLIN	30 0	w^w1; 1^133	
MCLIN	6 8 9 10	W = 8; S = 2.5; L = 12	!L = 13
SLC	9 10	L = 0.5; C = 3	!C = 5
DEF3P	1 9 10	TRAFO	!OUTPUT NETWORK
BAL2	2 0		
TRAFO	3 1 2 0		
DEF2P	1 3	IMP2	
BAL1	1 0		
BAL1	2 0		
TRAFO	3 1 2 0		
DEFIP	3	IMP	

FREQ

SWEEP	50 550 25

OUT

IMP RE(Z1)		
IMP IM(Z1)		
IMP DB(S11)	GR1	
IMP S11 SC2		
!IMP VSWR1		
BAL3 RE(Z1)		!TO DETERMIN E
BAL3 IM(Z1)		!ZLOAD
IMP2 DB(S12)		

GRID

RANGE	50 550 50
GR1	–30 0 5

TERM

IMP2	BAL1 REFIMP	!50 Ω PORT1

OPT

RANGE	50 550
!IMP	VSWR1 < 1.6
IMP	MODEL REFIMP

Note

1. Slp file BLF5480U does contain data given in “Appendix C”.

9 APPENDIX B

Input BLF548 application (300 W/28 V/500 MHz); input in 1P file fit between 100 – 500 MHz with 1 : 4 transformer (Zc = 10 Ω) to determine Gp and inputVSWR ZiN and Gp data derived from calculations, which represent the performance of the device after applying the output NEtwork, given in “Appendix A” to IT.

Table 5

DIM			
FREQ	MHz		
RES	OH		
IND	NH		
CAP	PF		
LNG	MM		
VAR			
C1\57.83880	!C = 27		
L2\64.28568	!L = 25		
L3 = 55	!L = 25		
R11 = 0.000207	!4.1		
C11 = 241.8059	!65		
L12 = 0.000048	!0.41		
CKT			
RES	1 0	R = 50	
DEF1P	1	REFIMP	!50 Ω LOAD
S1PA	2 0	BLF54812	
DEF1P	2	BAL1	!BLF548’s Zin
RES	1 2	R^R11	
SLC	2 0	L^L12; C^C11	
DEF1P	1	BAL3	
GAIN	1 2	A = 22.65; S = –4.79; F = 150	!Gp DATA DERIVED FROM
GAIN	2 3	A = 0; S = 0.5; F = 350	!Table 3
GAIN	3 4	A = 0; S = 1.5; F = 400	!13 dB GAINCORRECTION SINCE
DEF2P	1 3	GPAF	!2 SECTIONS ARE INVOLVED
SLC	1 0	L = 0.5; C\0.016169 !2	
COAX	1 2 0 3	DI = 0.91; DO = 2.98	L^L3; ER = 2.03; TAND = 0.0002; RHO = 1
MSUB		ER = 2.2; H = 0.79; T = 0.035; RHO = 0.72; RHG = 1	
SLC	2 4	L = 0.5 C^cl	
SLC	3 5	L = 0.5 C^cl	
COAX	4 6 7 8	DI = 1.15; DO = 1.45; L^L2; ER = 2.03; TAND = 0.0002	RHO = 1
COAX	5 8 7 6	DI = 1.15; DO = 1.45; L^L2; ER = 2.03; TAND = 0.0002	RHO = 1
UNIT	7 0		
SLC	6 8	L = 0.5; C\0.103064 IC – 2.1	
MCLIN	6 8 9 10	W = 8; S = 2.5; L\3.056280	
SLC	9 10	L = 0.5; C\0.923788; !c – 3.5	
DEF3P	1 9 10	TRAFO	INPUT NETWORK
BALI	2 0		
TRAFO	3 5 4 0		
GPAF	4 2		
GPAF	5 1		
DEF2P	1 3	IMP2	!TO DETERMINE S11, S12
FREQ			
SWEEP	100 500 25		
OUT			
IMP2 TE(Z2)			
IMP2 IM(Z2)			
IMP2 DB(S22)	GR1	!S11 AT 50 Ω PORT	
IMP2 DB(S12)	GR1	!CALCULATED POWERGAIN	
IMP2 VSWR2			
GPAF DB(S21)		!BLF548’s GP	
GRID			
RANGE	100 500 25		
GRI	–20 20 5		
TERM			
IMP2 BAL1 REFIMP			
OPT			
RANGE	150 550		
IMP2 DB(S12) > 11			
IMP2 DB(S12) < 12.4			

Note

1. S1p file BLF548I2 does contain data given in Table 4. Format as shown in “Appendix C”.

10 APPENDIX C

Fit dummyloadmodel to calculated BLF548 Zoutput !FIlename: E:\users\blf548ou.s1p
!reference: calculated data derived from transmod program;!Zload converted to Zoutput

DIM

FREQ	GHZ
RES	OH
IND	NH
CAP	PF
LNG	MM

CKT

MSUB	ER = 6.5	H = 1.02; T = 0.035 ; RHO = 1; RGH = 0		
IND	1 0	L = 0.168		!inductance to ground
RES	1 2	R = 5.28 0		
SLC	1 2	L = 0.3	C = 40	
SLC	1 2	L = 0.3	C\65.8	
CAP	2 0	C = 1		
IND	2 3	L = 0.165		!bonding wires at drainside
MLIN	3 4	W = 10.8	L = 1.6	!metallization of header
DEF1p	4	MOD		!1-port of dummyload
S1PA	1 0	BLF5480U.s1p		!calculated z-parameters of
DEF1p	1	MES		!BLF548 outputimpedances

FREQ

SWEEP	0.1	0.5	.050

OUT

MOD	RE(Z1)
MOD	IM(Z1)
MES	RE(Z1)
MES	IM(Z1)

OPT

MOD MES
MODEL

Table 6

# GHz Z	RI	R 1			
0.050	5.26	–0.86	! freq	R1	X1
0.075	5.04	–1.22			
0.100	4.73	–1.51			
0.125	4.39	–1.73			
0.150	4.04	–1.87			
0.175	3.70	–1.94			
0.200	3.36	–1.95			
0.225	3.05	–1.92			
0.250	2.76	–1.85			
0.275	2.50	–1.75			
0.300	2.27	–1.63			
0.325	2.06	–1.50			
0.350	1.87	–1.35			
0.375	1.71	–1.21			
0.400	1.56	–1.05			
0.425	1.43	–0.90			
0.450	1.31	–0.74			
0.475	1.21	–0.59			
0.500	1.11	–0.43			
0.525	1.03	–0.28			
0.550	0.95	–0.13			

Note

1. Zload is derived from data given in Fig.3.

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Belarus: Hotel Minsk Business Center, Bld. 3, r. 1211, Volodarski Str. 6,
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France: 51 Rue Carnot, BP317, 92156 SURESNES Cedex,
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Germany: Hammerbrookstraße 69, D-20097 HAMBURG,
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Italy: PHILIPS SEMICONDUCTORS, Piazza IV Novembre 3,
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