APPLICATION NOTE

A Broadband 100 W Push Pull Amplifier for Band IV & V TV Transmitters based on the BLV861

AN98033

Philips
Semiconductors

PHILIPS

CONTENTS

1 INTRODUCTION

Intended for applications in TV transmitter output stages a broadband high power amplifier has been described with a single BLV861 transistor. The design objectives are given in Table 1. In the following sections a background information of the BLV861 will be given, followed by a description and tuning of the application circuit. A broadband small signal and large signal performance of the BLV861 will be described. Finally several tests results will be shown measured in channel 69 (855/860 MHz). Additional AM-AM and AM-PM (ICPM) characteristics are presented which is a commonly measured parameter in analog vs. digital television transmitters. Because of the increasing interest for combined amplification of sound and vision also two and three-tone performance has been presented.

Table 1 Design objectives of the BLV861 amplifier

	SYMBOL	VALUE	UNIT
Frequency band	BW	470 to 860	MHz
Output power @ 1 –dB compression *	P_{out}	>100	W
Power gain	G_P	>8.5	dB
Gain ripple	$G_{P\text{-ripple}}$	±0.5	dB
Efficiency	η	>55	%
Input Return loss	IRL	–3 to –8	dB
Conditions: V_{ce} = 28 V; P_{LOAD} = 100 W; I_{CQ} = 100 mA; T_{HS} = 25 °C			

2 TRANSISTOR DESCRIPTION

2.1 BLV861 Internal Configuration

The BLV861 is a 100 W transistor encapsulated in a SOT289 package. A simplified outline of this package is shown in Fig.1. The emitter is connected to the flange and the collector leads are internally shorted for DC because of the applied postmatching. Due to this configuration its not possible to measure both collector currents separately.

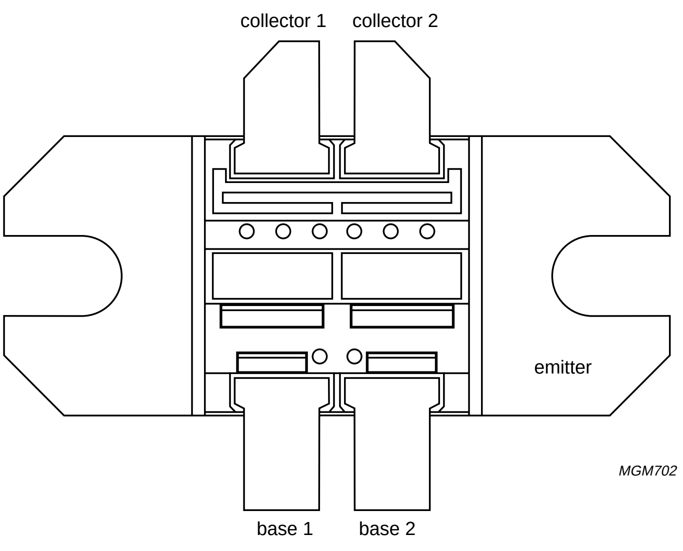


Fig.1 SOT289 package outline of the BLV861.

The active part of the BLV861 consists of two dies with a 6 µm emitter-pitch technology. It incorporates high value polysilicon emitter ballasting resistors for an optimum temperature profile in class-AB as well as in class-A operation (note 1). Combined with gold metallization it offers a high degree of reliability and ruggedness. The main transistor data is summarised in Table 2.

Table 2 Summary of main transistor data; note 1

MODE OF OPERATION	f [MHz]	V_{CE} [V]	P_L [W]	G_P [dB]	EFF. [%]	$G_{P\text{-}COMP.}$ [dB]	$R_{thj\text{-}hs}$ [K/W]
Class-AB	860	28	100	>8.5 dB	>55 %	<1dB	<1.0

Note

1. $P_{DISSIPATION} \leq 140$ W (DC) and $T_{junction,max} < 200$ °C.

2.2 BLV861 Internal Matching

The BLV861 is internally matched to increase the useable bandwidth and to elevate the device terminal impedance. Figure 2 shows the equivalent circuit of one section BLV861, with its matching circuitry. The input is pre-matched with two lowpass LC-sections to get low-Q transformation steps and high intermediate impedance level at the base terminals. The output is post-matched with a collector-to-collector shunt inductor which is designed to resonate with the transistor output capacitance at the low end of the band. This results in an increased broadband capability and increased impedance level at the transistor output.

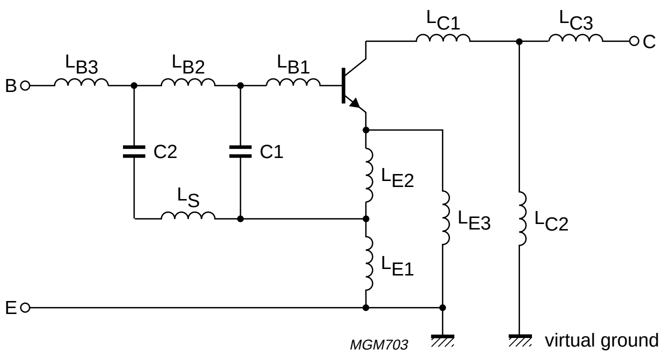


Fig.2 Internal circuit topology of one section BLV861.

2.3 Gain and Impedance Data

The gain and impedance data are listed in the Table 3 and curves are given in Figs 8 to 10. These data have been measured in a fixture tuned for maximum gain at rated output power for each frequency. The impedance data which is given has been measured from base-to-base and collector-to-collector terminals.

Table 3 Gain and impedance data (total device)

f MHz	G_P dB	η %	Z_{IN} (Ω)		Z_{LOAD} (Ω)	
			REAL{Z_{IN}}	IMAG{Z_{IN}}	REAL{Z_{LOAD}}	IMAG{Z_{LOAD}}
471	11.34	52.29	0.55	4.74	13.91	–10.13
519	10.97	52.99	1.23	5.17	13.40	–5.12
567	10.46	52.91	2.24	6.12	12.18	–3.71
615	10.12	53.54	3.26	6.82	10.10	–3.32
663	9.76	53.38	4.39	7.90	8.82	–3.65
711	9.99	54.28	5.44	8.42	6.94	–4.13
759	10.12	53.71	7.16	7.13	5.85	–4.45
807	9.96	54.03	8.04	4.14	5.28	–4.80
855	8.72	53.71	5.96	0.91	5.02	–5.81
Conditions: V_{CE} = 28 V; P_{LOAD} = 100 W; I_{CQ} = 100 mA; T_{HS} = 25 °C						

3 AMPLIFIER DESIGN

The total description of the amplifier is given in Figs 6 and 7 and Table 8. The amplifiers input and output matching networks contain mixed microstrip-lumped elements networks to transform the terminal impedance levels to approx. 25 Ω balanced. The remaining transformation to 50 Ω unbalanced is obtained by 1 : 2 balun transformers. The baluns B_1 and B_2 are 25 Ω semi-rigid coax cables with an electrical length of 45° at midband and a diameter of 1.8 mm, soldered over the whole length on top of microstrip lines. To keep the circuit in balance two stubs L_1 and L_8 with the same length have been added. For low frequency stability enhancement the input balun stubs are connected to the bias point by means of 1 Ω series resistors. Large capacitors (C_4 and C_{11}) are added at the biasing points to improve the amplifiers video response. The printed-circuit board laminate utilised is PTFE-glass with an ε_r = 2.55 and a thickness of 0.51 mm (20 mills). Specification of all components are given in Table 8.

3.1 Input Network

The input network is designed for high gain match and flat overall gain versus frequency. This is achieved by a three section lowpass filter with a series capacitor at 50 Ω input impedance level. Three variable capacitors are included for fine tuning of the gain. C_5 with an additional trimmer is utilised to tune the gain slope at low end of the frequency while C_7 is intended to tune the gain slope at 860 MHz. C_6 on the other hand is used to tune the gain ripple. See circuit diagram in Figs 6 and 7. The capacitor C_7 is placed close to the base of the BLV861 to maintain low Q transformation.

3.2 Output Network

The output network is designed for high output power and efficiency in full bandwidth. First two capacitors (C8 and C9) are placed close to each other. The physical distance between the capacitors is shown in Fig.7. RF dissipation in shunt capacitors, due to circulating currents, is a critical factor in the design of the output networks. The most critical component is the first shunt capacitor at the collector terminals. The current in this capacitor is at maximum level when operated at the upper end of the frequency band at max. power level. In practice this usually results in melting of the solder which on its turn degrades the power capability as experienced with ATC100B low Q capacitors. On the next page a comparison of ATC100B and ATC180R capacitors has been given. Calculations has been carried out in order to determine the heat development in this capacitors. The power transfer efficiency is given by:

$$\eta_{\text{power transfer}} = \left(1 - \frac{Q_L}{Q_U}\right)^2 \qquad (1)$$

Expressed in power losses we have:

$$P_{LOSS} = 10 \cdot \log\left(\frac{1}{\eta}\right)^2 \qquad (2)$$

To get an impression of the body temperature of a capacitor, which can be strongly influenced by its own unloaded Q, we first have to define heat intensity of a body. The temperature of this body is proportional to the heat intensity. Generally the heat intensity of a body is defined as Joule per unit volume per second:

$$Heat\_intensity = \frac{\text{Absorbed power}}{\text{Volume}} = \frac{Joule}{m^3} \cdot \frac{1}{s}\left[\frac{W}{m^3}\right] \qquad (3)$$

An example has been given in order to confirm the power capability of the ATC180R capacitors which has been used in BLV861 application circuit.

Table 4 Comparison of the electrical parameters of the ATC100B and ATC180R

TYPE OF CAPACITOR	ATC100B	ATC180R-1	ATC180R-2	UNIT
Value	13	10	2.7	pF
ESR	0.097	0.068	0.123	Ω
Unloaded Q (Q_U)	147	271	559	
Resonance frequency	1.79	3.33	5.73	GHz
Current	5.56	7.66	5.16	A
Dimensions	2.794 × 2.794 × 2.591	2.67 × 1.78 × 2.29	2.67 × 1.78 × 2.29	mm^3
Frequency of operation	860	860	860	MHz
Power to be transferred	100	100	100	W
Loaded Q (Q_L); note 1	3	3	3	

Note

1. Assumed high loaded Q is present at the upper end of the frequency (worst case).

Consider a single 13 pF ATC100B capacitor, see Table 4, then we get from [2]:

$$P_{LOSS} = 10 \cdot \log\left(\frac{1}{1 - \frac{2}{147}}\right)^2 = 0.0595 \text{ dB}, \qquad (4)$$

which means that 2.70% (2.7 W) of the through-put power is converted into heat.
The total heat intensity becomes:

$$Heat\_intensity = \frac{100\,[W] \cdot 0.027}{2.794\,[mm] \cdot 2.794\,[mm] \cdot 2.591\,[mm]} = 0.134\frac{W}{mm^3} \qquad (5)$$

In the same manner we can calculate the losses for the two paralleled ATC180R capacitors (10 pF//2.7 pF) which are used in the BLV861 output circuit. First we have to calculate the overall Q_U from the single component data as listed in table 4.

$$ESR_{TOT} = \frac{ESR_1 \cdot ESR_2}{ESR_1 + ESR_2} = 0.044\ \Omega \qquad (6)$$

$$C_{TOT} = C_1 + C_2 = 12.7 \text{ pF} \qquad (7)$$

$$Q_U = \frac{1}{2 \cdot \pi \cdot f \cdot ESR_{TOT} \cdot C_{TOT}} = 331 \qquad (8)$$

$$P_{LOSS} = 10 \cdot \log\left(\frac{1}{1 - \frac{2}{331}}\right)^2 = 0.0263\ \text{dB}\ , \tag{9}$$

which means that only 1.2% (1.2 W) of the through-put power is converted into heat.

The heat intensity is:

$$\text{Heat\_intensity} = \frac{100\,[\text{w}] \cdot 0.012}{2 \cdot 2.67\,[\text{mm}] \cdot 1.78\,[\text{mm}] \cdot 2.29\,[\text{mm}]} = 0.0554\frac{\text{W}}{\text{mm}^3} \tag{10}$$

As can be noticed, in case of two ATC180R capacitors the body temperature is more than factor 2 lower compared to an ATC100B capacitor. Taking into account the main parameters and power handling capability, it has been decided to utilise ATC180R as the first output matching capacitor. The capacitors need to be placed in full contact with the printed-circuit board in order to maintain better thermal resistance.

3.3 Bias Circuit

The class-AB bias circuit used is shown in Fig.3. This circuit has a very low power consumption allowing the use of low power SMD chip resistors. Two NPN transistors BD139 are used. T2 is chosen to operate in the reverse mode in order to have its lower collector to base diode voltage to track the base-emitter voltage of the BLV861. R3 mainly compensates for the difference between these two values. T2, T3 and BLV861 have been mounted on the same heatsink to have good temperature compensation. R4 is incorporated to improve video response and to protect T3 in case of short circuit in the BLV861 amplifier. Capacitor C15 bypass any RF leakage to T2. The bias circuit is fully integrated on the amplifier board, see Fig.7.

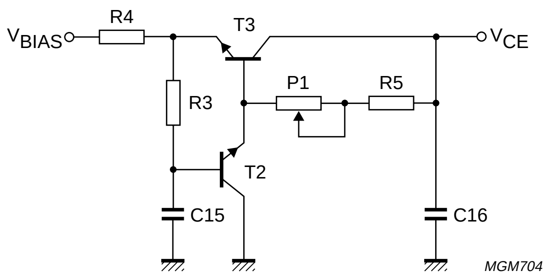


Fig.3 Class-AB bias circuit.

4 BROADBAND RF PERFORMANCE OF THE BLV861 AMPLIFIER

The amplifier has been tuned under class-A small-signal conditions and characterised under large signal class-AB conditions from 470 – 860 MHz. The conditions used shown in Table 5

Table 5 Conditions for class A and AB characterisation

	SMALL SIGNAL	LARGE SIGNAL
Class of operation	A	AB
Collector-emitter voltage	28 V	28 V
Quiescent current (I_{CQ})	1.0 A	0.1 A
Source/Load impedance	50 Ω	50 Ω
Heatsink temperature	25 °C	25 °C

4.1 Small Signal Response

Tuning high power amplifiers under small-signal class-A conditions to obtain optimum large signal performance was found to be a very suitable and save technique. The best small-signal response was determined experimentally.
The S_{11}, S_{22} and S_{21} response resulting in optimum large signal performance is given in Figs 11 to 14. The input is tuned for maximum gain and a flat response over the whole frequency band (470 – 860MHz). The output is tuned under both small signal and large signal to get an optimum power performance.

4.2 Large Signal Response

After the small-signal class-A tuning the amplifier was biased into class-AB operation. Gain, collector efficiency, input return loss and compression was determined versus frequency at a power level of 100 W (CW). The data are summarised in Figs 15 and 16 and Table 9. The power gain compression and collector efficiency are strongly sensitive to the location of capacitors C_8 and C_9, which have to be optimized experimentally. Shifting this capacitors from their initial location to the left will result in an improved power gain compression and a poor efficiency, while shifting to right will improve the efficiency. The average gain power level is about 9.0 dB with a ripple of less than ± 0.3 dB. Broadband collector efficiency is fluctuating around 56% and shows a dip at midband (663 MHz, i.e. channel 45). Power gain compression in the band of interest is below 0.8 dB. Highest compression of 0.79 dB occurs at 860 MHz which is referenced to 40 W output power level (CW). The broadband input return loss varies from –3.5 dB at the lower end to less than –10 dB at the upper end of the frequency range.

4.3 Amplifier Overdrive Capability Test

An 3 dB input overdrive test has been performed in order to force the amplifier beyond its saturation power and to check its overdrive capability. P_{OUT} vs. P_{IN} measurements have been done from zero to >3 dB above its nominal drive level at 860 MHz. The amplifier has proven to withstand a drive level of above 25 W many time for several minutes without degradation of the device. The power level associated with this level was 135 W (CW). Figs 17 and 18 presents the recorded data.

5 NON-LINEAR DISTORTIONS

Amplitude dependent waveform distortions are often referred to as non-linear distortions. This classification includes distortions which are dependent on average picture level (APL) changes and/or instantaneous signal level changes. Generally, amplifiers are linear over only a limited range, they may tend to compress or clip large signals. Non-linear distortions may also manifest themselves as crosstalk and intermodulation effects. The first three distortions measured and discussed in this section are:

- Intermodulation:
  - Two tone intermodulation, if sound and vision are amplified separately
  - Three tone intermodulation, in case of combined amplification.
- Incidental carrier phase modulation.

5.1 Intermodulation

Because of the increasing interest for combined carrier operation, the linear performance of the amplifier for two-tone and three-tone operation have been determined. Two tone and three tone IMD-measurement have been performed as defined in Fig.4. For two tone performance two carriers have been chosen which represents the vision and sideband carrier. Three tone measurement is done with an additional carrier which represents the sound carrier. The different tone systems used are listed in Table 6.

Table 6 Survey of used tone system for intermodulation measurements

CHANNEL 69	SYSTEM A	SYSTEM B	SYSTEM C
f_{visin} = 855.25 MHz **$f_{sideband}$ = 859.68 MHz** **f_{sound} = 860.75 MHz**		**dB**	
Vision amplitude	–8	–5	–3
Sideband amplitude	–16	–17	–20
Sound amplitude	–10	–10	–10

Two tone IMD-performance is depicted as a function of the output peak-sync power ($P_{O,SYNC}$) in Figs 19 to 21. Figure 22 shows three tone IMD performance of all three systems, shown in Table 6, measured in channel 69. As can be noticed $P_{O,SYNC}$ of each system is different. System A has a much higher output sync power related to system B and C, at the same average output power level. In all cases $P_{O,SYNC}$, is assumed to be at a certain reference level which is 0 dB. Based on this assumption conversion formulas are given to calculate different power levels regarding all systems, see “Appendix A”.

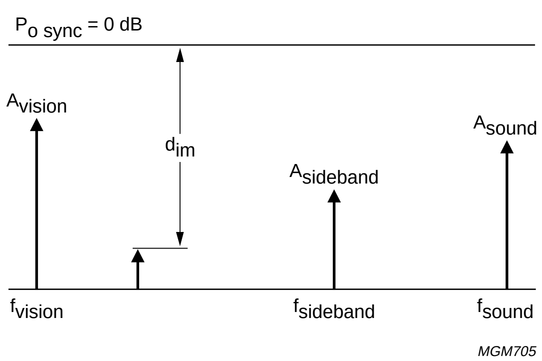


Fig.4 Definition of IMD measurement.

Finally a full band intermodulation performance has been given which is measured according to system A. As can be noticed a better linearity can be obtained around channel 45, see Table 7. A 3D graph which represents IMD = ($P_{O,SYNC}$, frequency channel) is given in Fig.23.

Table 7 Intermodulation vs. output power for 9 TV channels in Band IV and V (referred to $P_{O,SYNC}$ level)

SYSTEM A (–8/ –16/ –10)		TV CHANNELS								
$P_{O,AVG}$	$P_{O,SYNC}$	21	27	33	39	45	51	57	63	69
W		dB								
0.1	0.35	–42.3	–39.5	–38.6	–39.7	–41.4	–39.2	–37.8	–38.2	–37.8
1	3.53	–41.9	–39.8	–39.3	–39.5	–40.2	–38.9	–36.8	–37.8	–36.8
10	35.26	–49.0	–48.2	–47.9	–47.4	–46.5	–45.6	–41.5	–44.2	–41.5
20	70.52	–48.0	–50.6	–51.3	–49.6	–50.1	–51.0	–46.4	–50.5	–46.4
30	105.78	–44.6	–47.7	–48.7	–46.6	–55.8	–52.2	–50.6	–56.3	–50.6
40	141.04	–39.7	–43.3	–43.7	–42.1	–55.3	–45.9	–43.3	–45.3	–43.3
50	176.30	–35.4	–39.0	–39.2	–37.6	–45.0	–39.5	–37.6	–38.9	–37.6
60	211.56	–32.5	–35.3	–35.4	–34.0	–39.5	–35.3	–33.6	–34.4	–33.6

5.2 Incidental Carrier Phase Modulation

Incidental carrier phase modulation (ICPM) is a commonly measured parameter in analog television transmitters. This type of distortion is also commonly referred to as AM to PM distortion. The phase shift through an amplifier has the tendency to vary with output power. The capacitance of a reversed biased diode then varies with bias voltage. In an amplifier the trick is to avoid phase shift variations with output power level. Measurements have been carried out in order to determine the phase distortion of the amplifier using a network analyser. ICPM and also AM to AM distortion vs. input drive power is plotted in Figs 25 and 26 under several bias conditions.

The total setup for power sweep is reflected on Fig.24. The sweep range of the network analyser was set from –5 to +20 dBm corresponding with 0.05 to 15.6 W input drive power. Slight gain expansion at low output powers is obvious due to turn-on effects.

The phase is very linear up until the point where compression emerges. Important points for observation are the compression and phase deviation at 12.25 W drive power shown by marker 3 (valid for I_{CQ} = 100 mA). The phase shift is about ≈6.2° at 12.25 W input drive power (which corresponds to 100 W output load power) and the gain compression is around 1 dB referred to marker 2 (Figs 25 and 26).

6 TV CHARACTERISATION

Finally the amplifier is characterised with a PAL Composite Video Signal (CVS) (without soundcarrier) according CCIR standard G. The TV test setup used, is depicted in Fig.27. The following measurements have been performed under TV conditions:

- Differential gain
- Differential phase
- Transient sync compression vs. output peak sync power level
- Peak output power @ 1 dB compression.

TV measurements including differential gain and differential phase have been also characterised at V_{CE} = 32 V and I_{CQ} = 100 mA in order to attain higher output peak sync power.

6.1 Differential Gain

Differential gain is present if chrominance gain is dependent on luminance level. These amplitude errors are a result of the systems inability to uniformly process the high-frequency chrominance signal at all luminance levels. Differential gain is expressed in percentage of the chrominance gain at blanking level. The input video waveform used for differential gain evaluation is a modulated staircase with 10% rest carrier as given in Figs 28 and 29. Figures 33 to 40 reflects differential gain and differential phase in channel 69.

6.2 Differential Phase

Differential phase is present if a signals chrominance phase is affected by luminance level. This phase distortion is a result of a systems inability to uniformly process the high-frequency chrominance information at all luminance levels. The amount of differential phase distortion is expressed in degrees. See Figs 33 to 40.

6.3 Sync Compression vs. Peak-Sync Power

One effect produced by non-linearity above the blanking level is compression of the sync pulse. This effect is compensated in transmitters by making the sync pulses correspondingly greater before amplification. The degree of this so called sync-stretching required, depends on the sync compression due to the non-linearity in the amplifier. Evaluation of the sync compression is done using a input video waveform at black level, see Figs 28 and 5. The sync power is calculated by from the measured average output power and the sync-to-bar ratio after demodulation.
The sync-to-bar ratio is measured with the video waveform on line 18 containing a 100% white-bar. With this available ratio the sync amplitude can be calculated referenced to a 1 V sync-to-bar top level. The sync content is then normalised to a 1.11 V RF amplitude. An undistorted signal corresponds to 27% sync content. The sync power can then also be determined from the obtained sync level. The formula and definitions used for this calculation are given in formula 11 to 13 and in Fig.5. The output sync pulse content versus $P_{O,SYNC}$ power is presented in Fig.30.

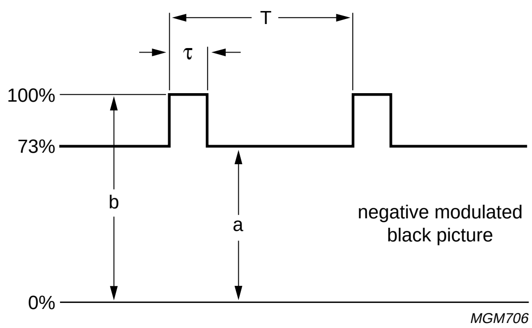


Fig.5 Composite video signal with black level for determining peak-sync power.

$$P_{RMS} = \frac{{U_{RMS}}^2}{R} = \frac{\left(\sqrt{\frac{1}{T}\int_o^{\tau} b^2 \cdot dt + \frac{1}{T}\int_{\tau}^{T} a^2 \cdot dt}\right)^2}{R} = \frac{\frac{\tau}{T} \cdot b^2 + \left(1 - \frac{\tau}{T}\right) \cdot a^2}{R} \quad (11)$$

$$P_{SYNC} = \frac{b^2}{R} \quad (12)$$

From [11] and [12] we have:

$$k = \frac{P_{SYNC}}{P_{RMS}} = \frac{1}{\frac{\tau}{T} + \left(1 - \frac{\tau}{T}\right) \cdot \left(\frac{a}{b}\right)^2} \text{(black picture)} \quad (13)$$

In case of no sync compression or expansion (a = 73% and b = 100%), then k = 0.567. In Fig.31 $P_{O,SYNC}$ versus $P_{IN,SYNC} = P_{IN,RMS}/k$ is depicted. In practice the allowable sync compression is bound to a maximum since sync-stretching is limited.

6.4 Output Sync Power Capability

Figure 32 shows gain versus $P_{O,SYNC}$ power for channel 69. The input video signal is at black level. The 1 dB compression point at I_{CQ} = 100 mA is above 120 W $P_{O,SYNC}$. At V_{CE} = 32 V on the other hand, 1 dB compression is above 150 W peak sync power.

7 CONCLUSIONS

A complete TV transmitter amplifier has been designed and characterised based on the BLV861, capable of operating in full band IV and V with flat gain and high output power in class-AB. BLV861 is able to generate 100 W CW power and a power gain compression below 1 dB in band IV and V. Overall gain of the amplifier is >8.5 dB and an efficiency of ± 55%. TV-measurements have been carried out showing a 1 dB compression point above 120 W $P_{O,SYNC}$ at V_{CE} = 28 V and 150 W at V_{CE} = 32 V.

- Amplifier shows an agreed linearity performance in class AB operation both under two tone and three tone conditions
- Biasing the amplifier at a V_{CE} = 32 V results in a higher output peak sync power and a better linearity response.

8 REFERENCES

Ref.1: Rohde & Schwarz Sound and Broadcasting:
"Rigs and Recipes how to measure and monitor...".

Ref.2: Philips Semiconductors Nijmegen, Prod. group Transistors and Diodes
BLV862 Application note: AN98014.

Ref.3: American Technical Ceramics:
The RF capacitor handbook, June 1970 / first edition.

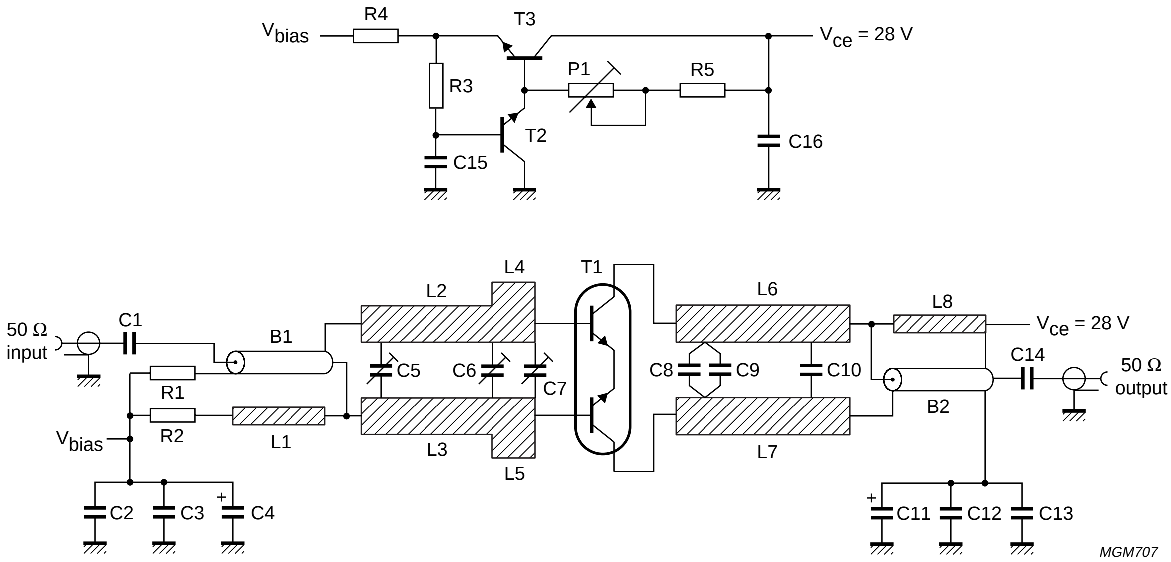


Fig.6 BLV861 amplifier circuit.

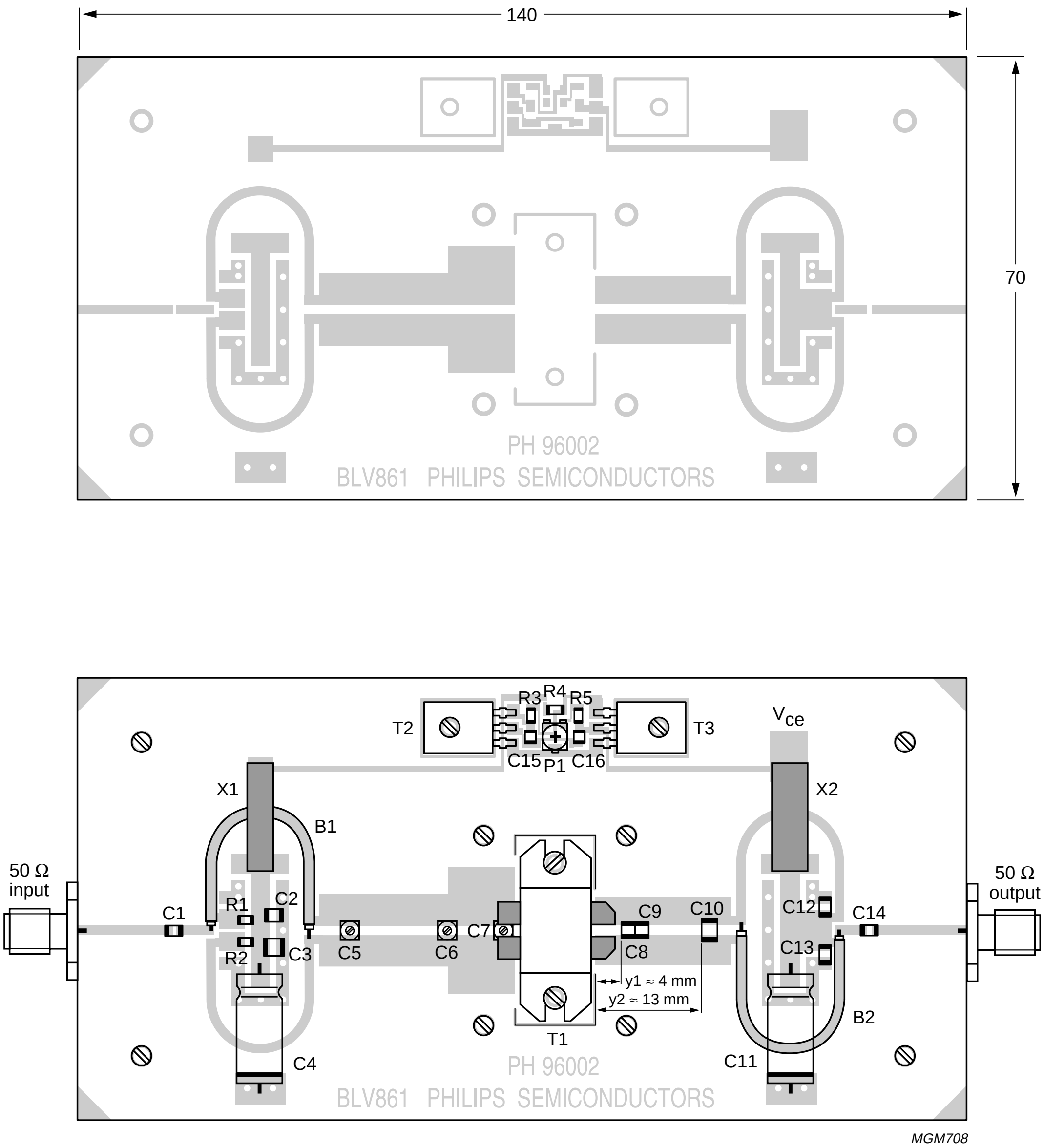


Fig.7 BLV861 Amplifier Circuit Board and layout.

Table 8 List of components

COMPONENT	DESCRIPTION	VALUE	DIMENSIONS
C1	multilayer ceramic chip capacitor; note 1	15 pF	
C2 and C12	multilayer ceramic chip capacitor	15 nF	2222 590 16629
C3 and C13		100 nF	2222 581 16641
C4 and C11	solid aluminium capacitor	100 µF/40 V	2222 031 37101
C5	multilayer ceramic chip capacitor; note 2 + Tekelec trimmer	2.2 pF	
C6		10 pF	
C7		15 pF	
C8	multilayer ceramic chip capacitor; note 3	2.7 pF	
C9		10 pF	
C10	multilayer ceramic chip capacitor; note 2	3 pF	
C14	multilayer ceramic chip capacitor; note 1	30 pF	
C15		100 pF	
C16	multilayer ceramic chip capacitor	15 nF	
R1 and R2	SMD resistor	1 Ω	805
R3		47 Ω	
R4		1 Ω	
R5		1 k2 Ω	
P1	potentiometer	5 kΩ	
T1	NPN push-pull RF-transistor	BLV861	9340 542 40112
T2 and T3	NPN transistor	BD139	9330 912 20112
B1	semi rigid coax balun UT70-25	Z = 25 ±1.5 Ω	47.0 mm
B2			

Notes

1. American Technical Ceramics type 100A or capacitor of same quality.
2. American Technical Ceramics type 100B or capacitor of same quality.
3. American Technical Ceramics type 180R or capacitor of same quality.
4. The striplines are on a double copper-clad printed-circuit board: PTFE-glass material (TLX8) from Taconic (epsilon of 2.55).

Narrowband Gain and Impedance Data.

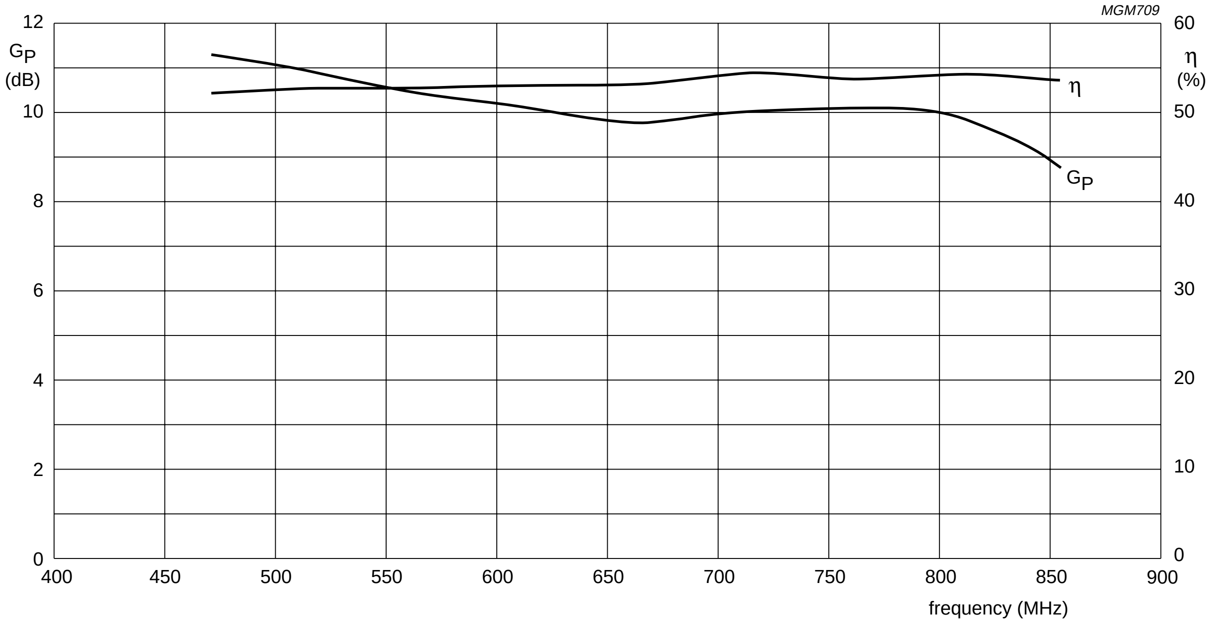


Fig.8 Maximum power gain and Collector efficiency.

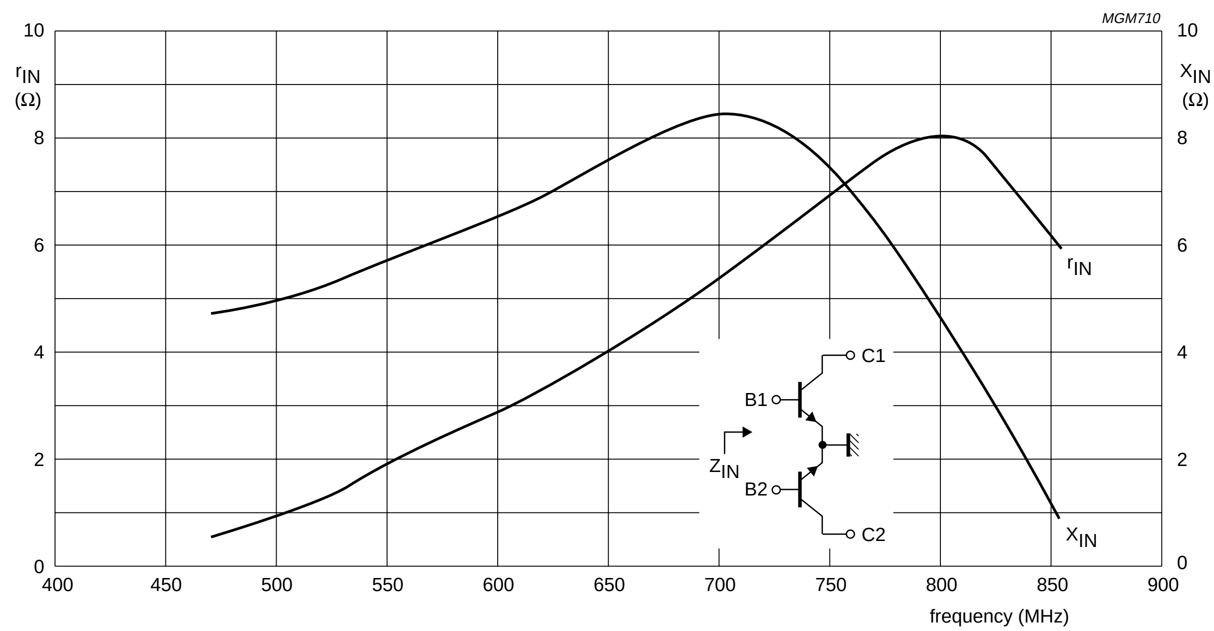


Fig.9 Input Impedance.

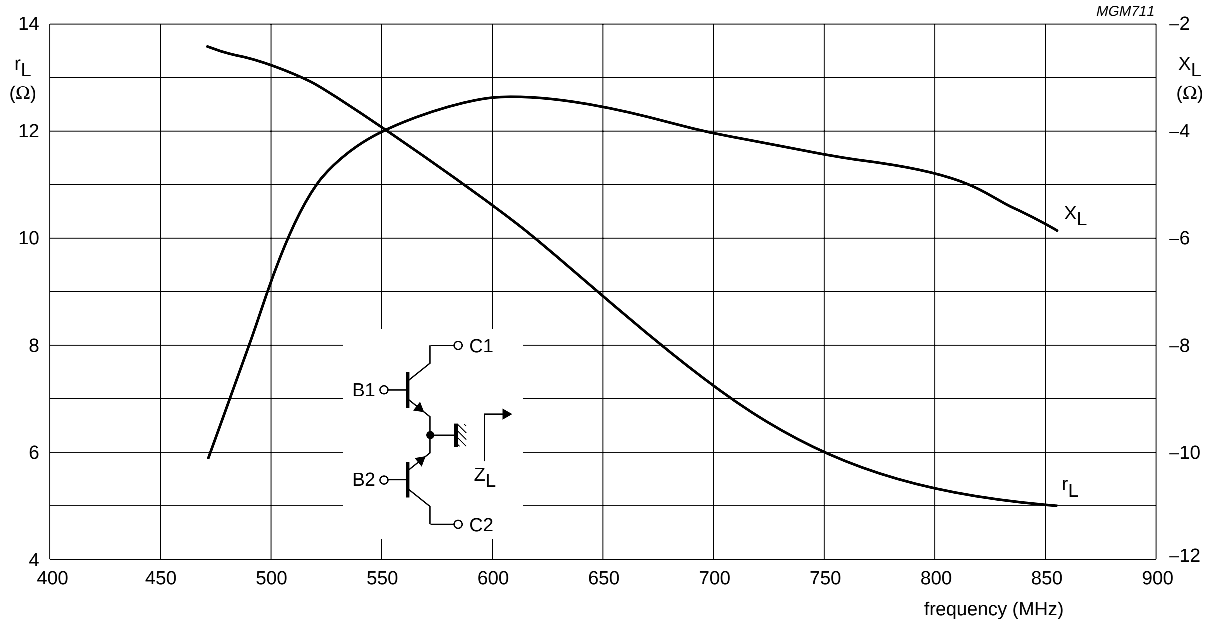
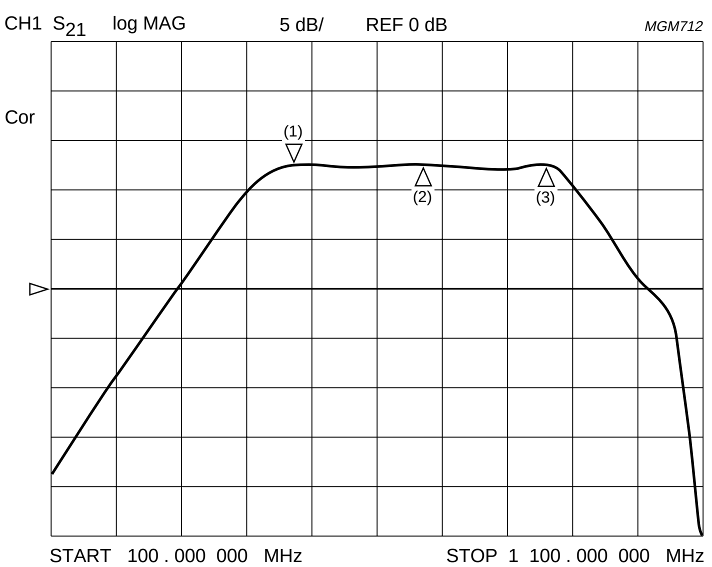
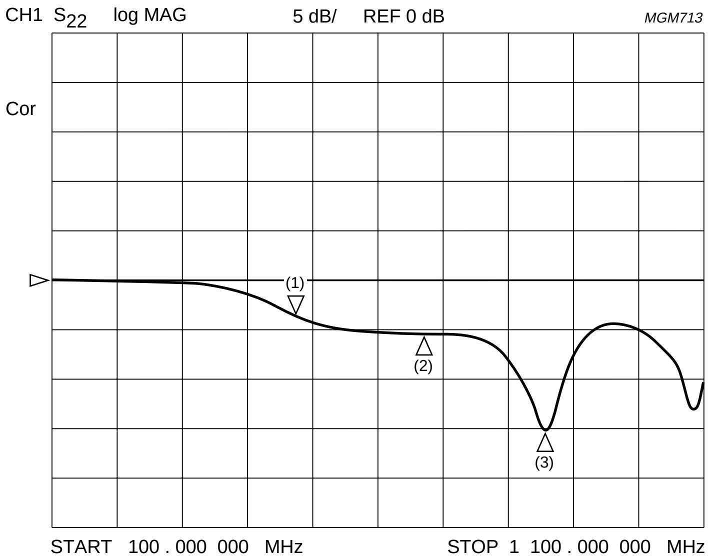


Fig.10 Output Impedance.



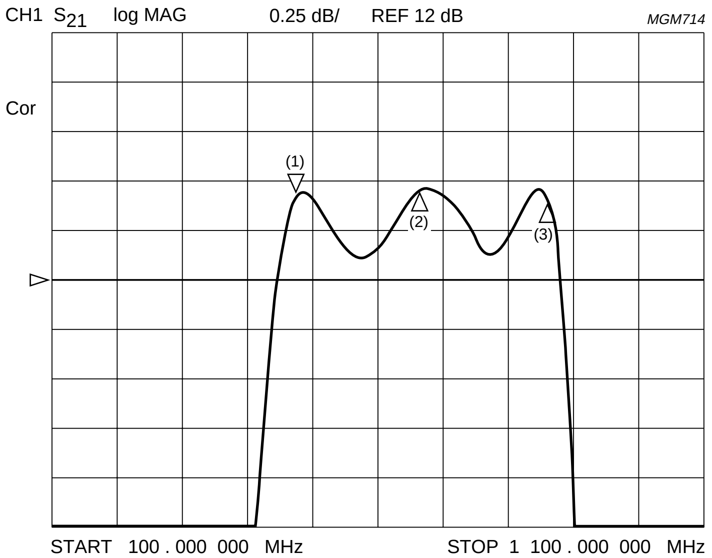
(1) 470 MHz: 12.38 dB.
(2) 665 MHz: 12.45 dB.
(3) 860 MHz: 12.38 dB.

Fig.11 Broadband Small Signal Respons S_{21}.



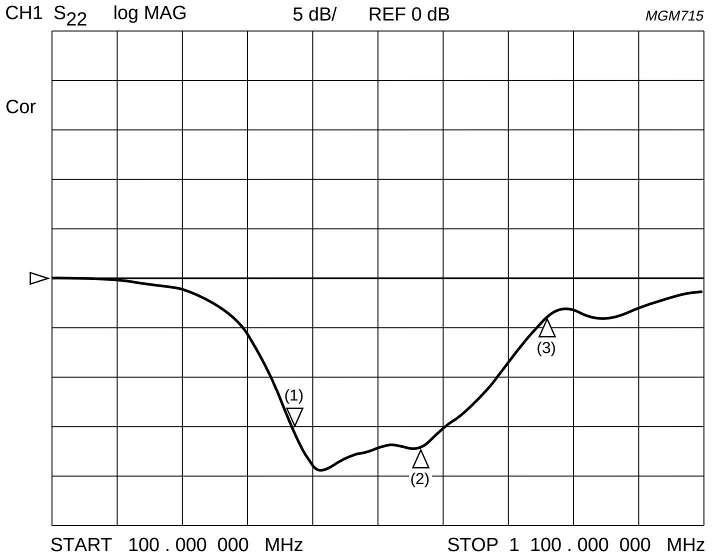
(1) 470 MHz: –3.50 dB.
(2) 665 MHz: –5.56 dB.
(3) 860 MHz: –15.09 dB.

Fig.12 Broadband Small Signal Respons S_{11}.



(1) 470 MHz: 12.38 dB.
(2) 665 MHz: 12.45 dB.
(3) 860 MHz: 12.38 dB.

Fig.13 Broadband Small Signal Respons S_{21}-ripple.



(1) 470 MHz: –15.34 dB.
(2) 665 MHz: –16.96 dB.
(3) 860 MHz: –4.14 dB.

Fig.14 Broadband Small Signal Respons S_{22}.

Table 9 Broadband Large Signal Performance

FREQUENCY MHz	Gp dB	ΔGp dB	IRL dB	η_c %
471	9.27	–0.67	–3.49	57.70
519	9.37	–0.47	–4.25	54.11
567	9.33	–0.46	–5.08	56.24
615	8.98	–0.74	–5.64	57.51
663	8.86	–0.67	–5.81	50.02
711	9.22	–0.55	–5.52	57.60
759	8.94	–0.59	–5.60	58.55
807	8.76	–0.69	–7.14	57.05
855	9.10	–0.79	–10.12	56.42

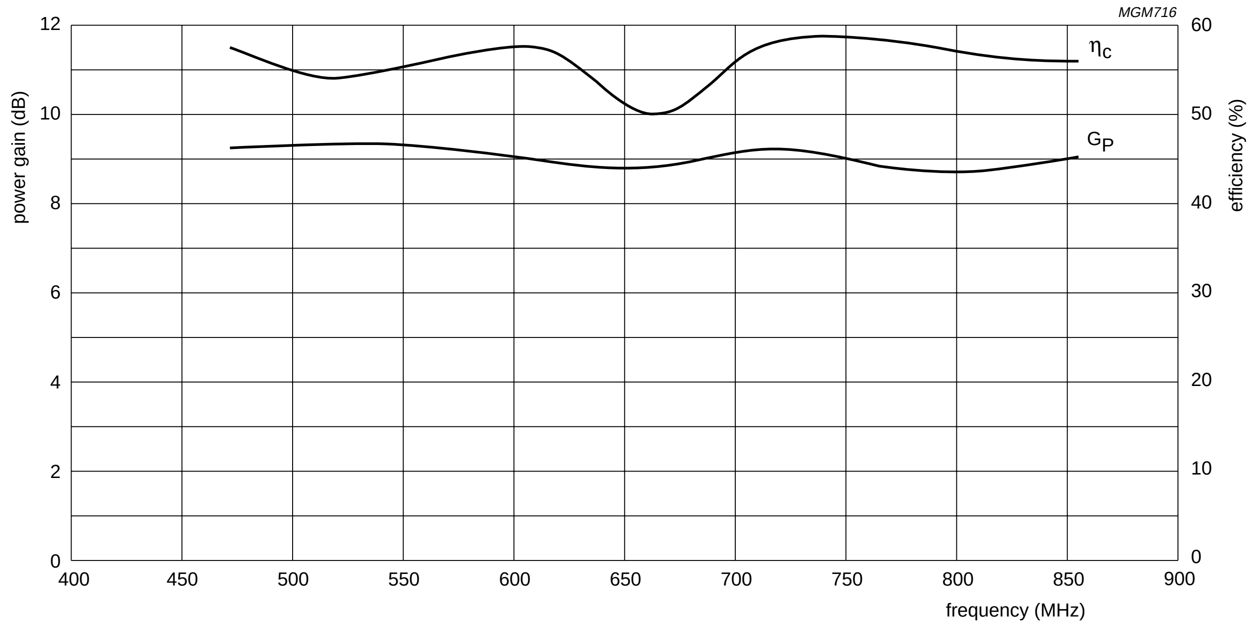


Fig.15 Broadband power gain and collector efficiency.

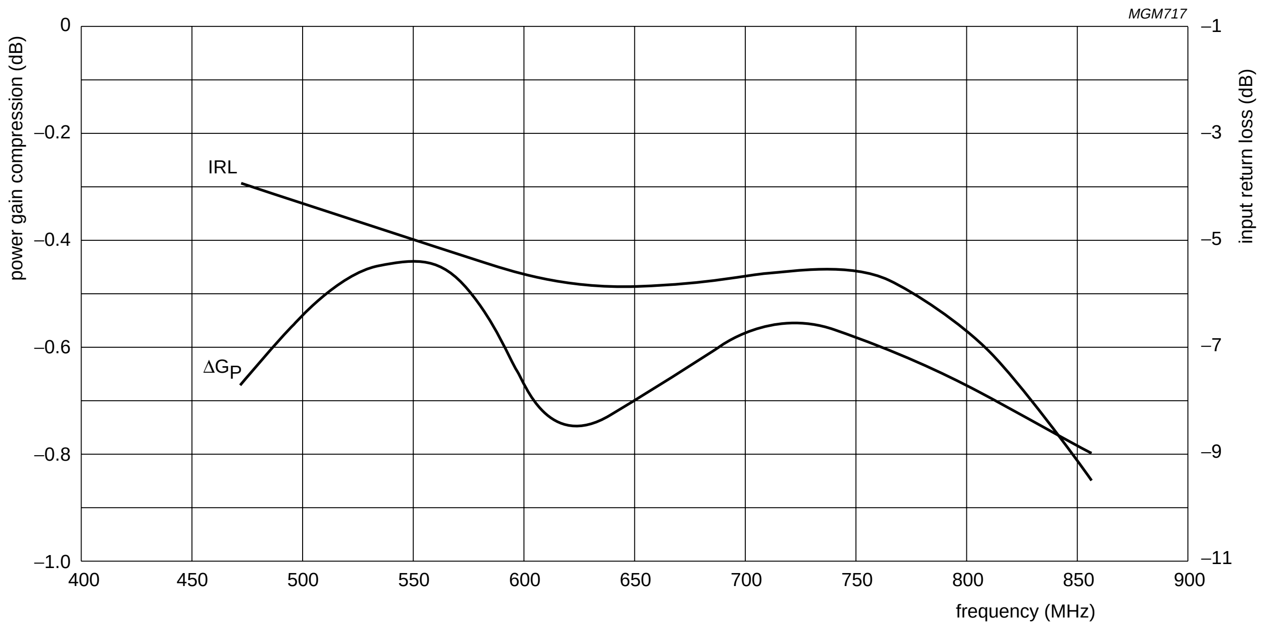


Fig.16 Input return loss and power gain compression vs. frequency.

Amplifier Overdrive Capability Test @ 860 MHz.

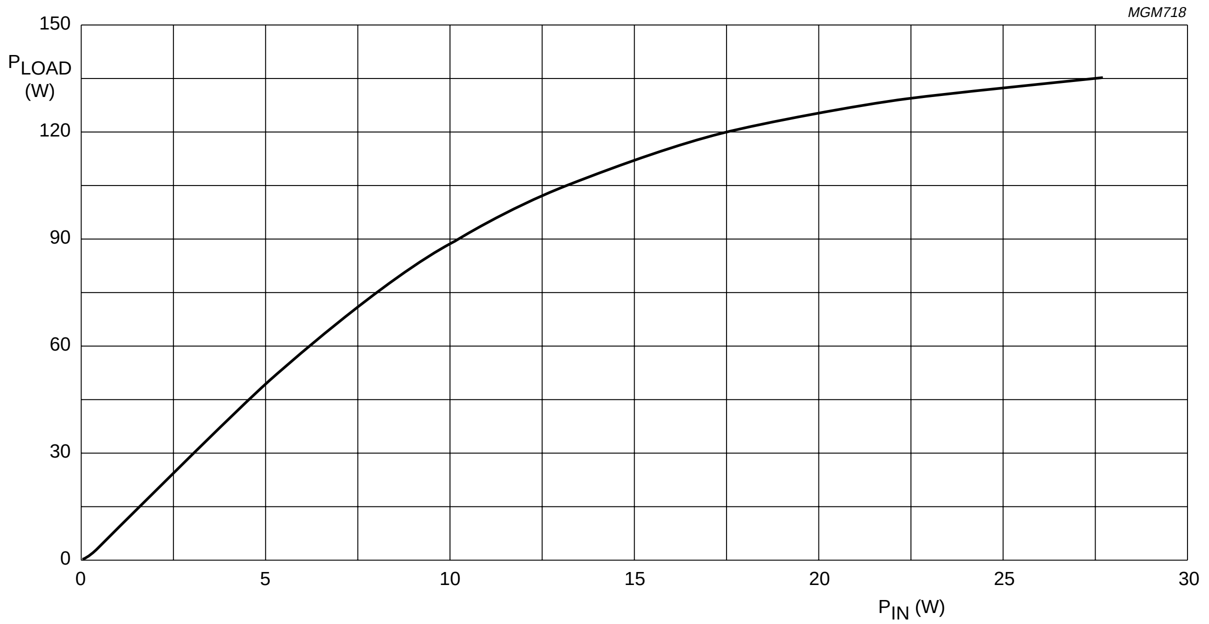


Fig.17 Load power vs. input drive power.

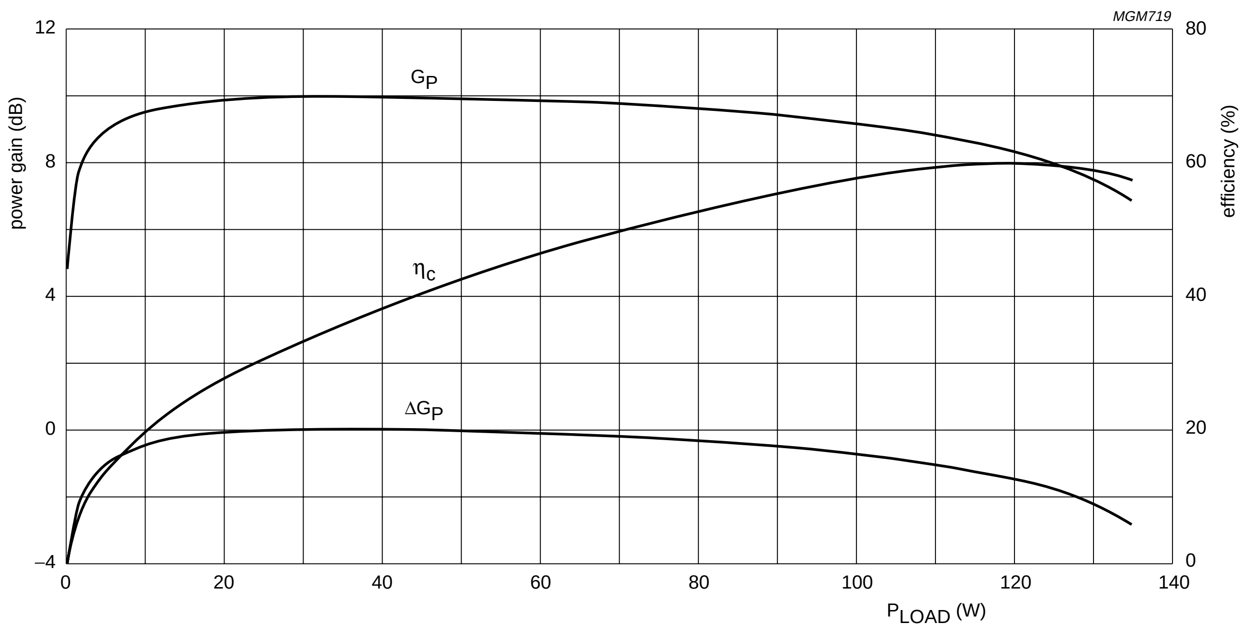


Fig.18 Power gain, power gain compression and collector efficiency vs. load power.

Two Tone Intermodulation Performance.

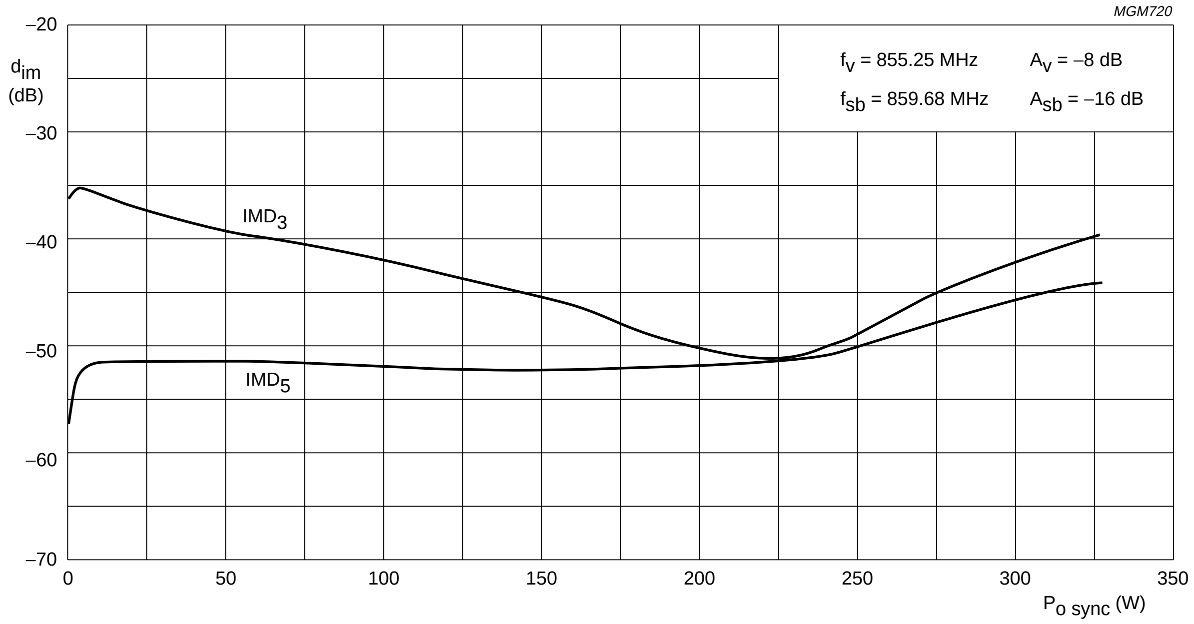


Fig.19 Two tone IMD according to System A.

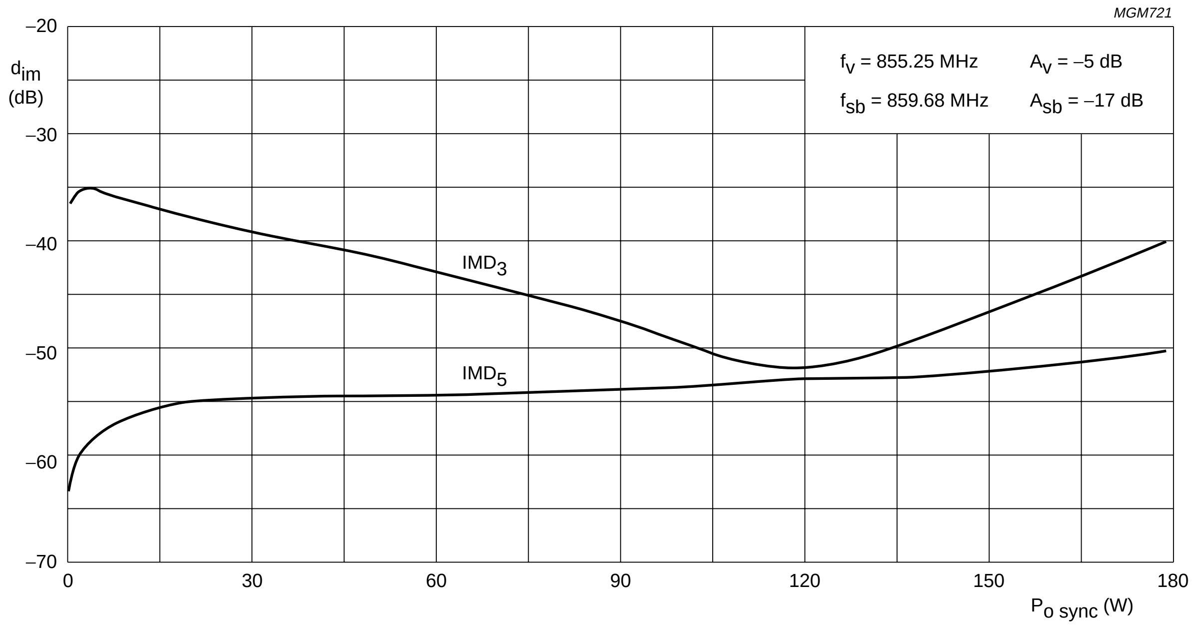


Fig.20 Two tone IMD according to System B.

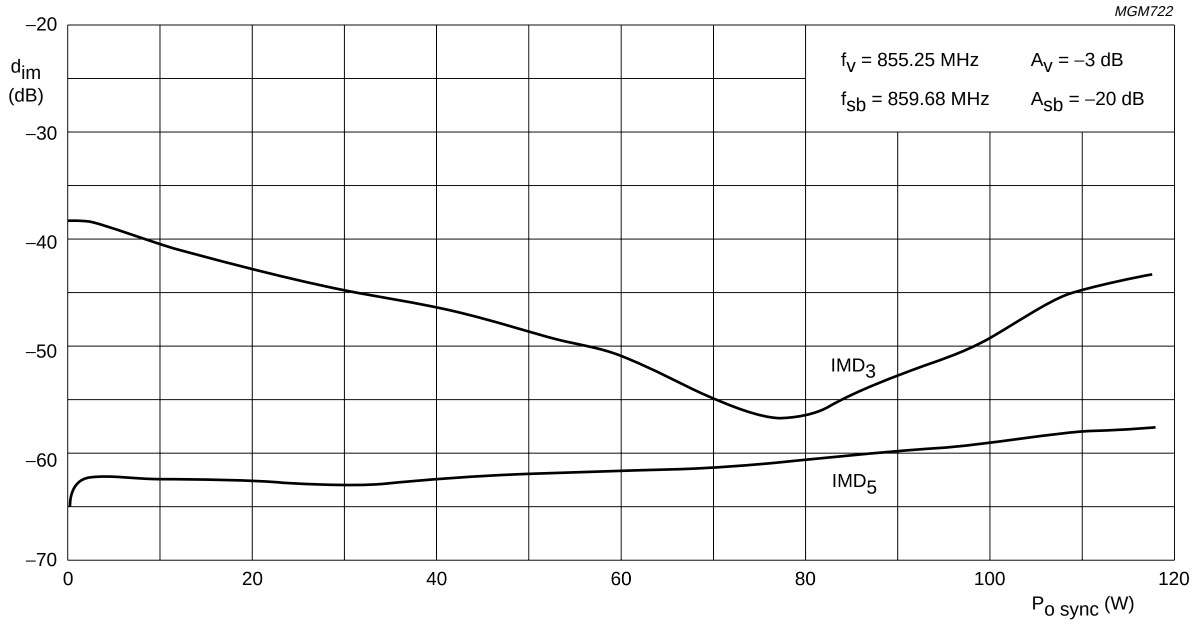


Fig.21 Two tone IMD according to System C.

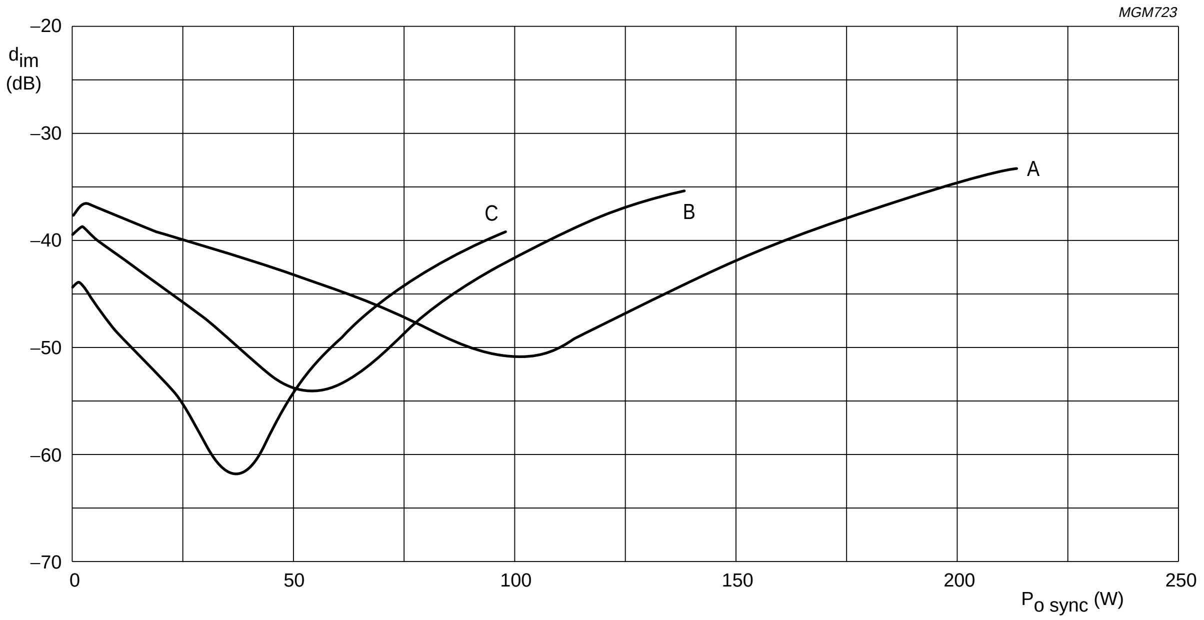


Fig.22 Tree Tone IMD at Channel 69 according to system A, B and C.

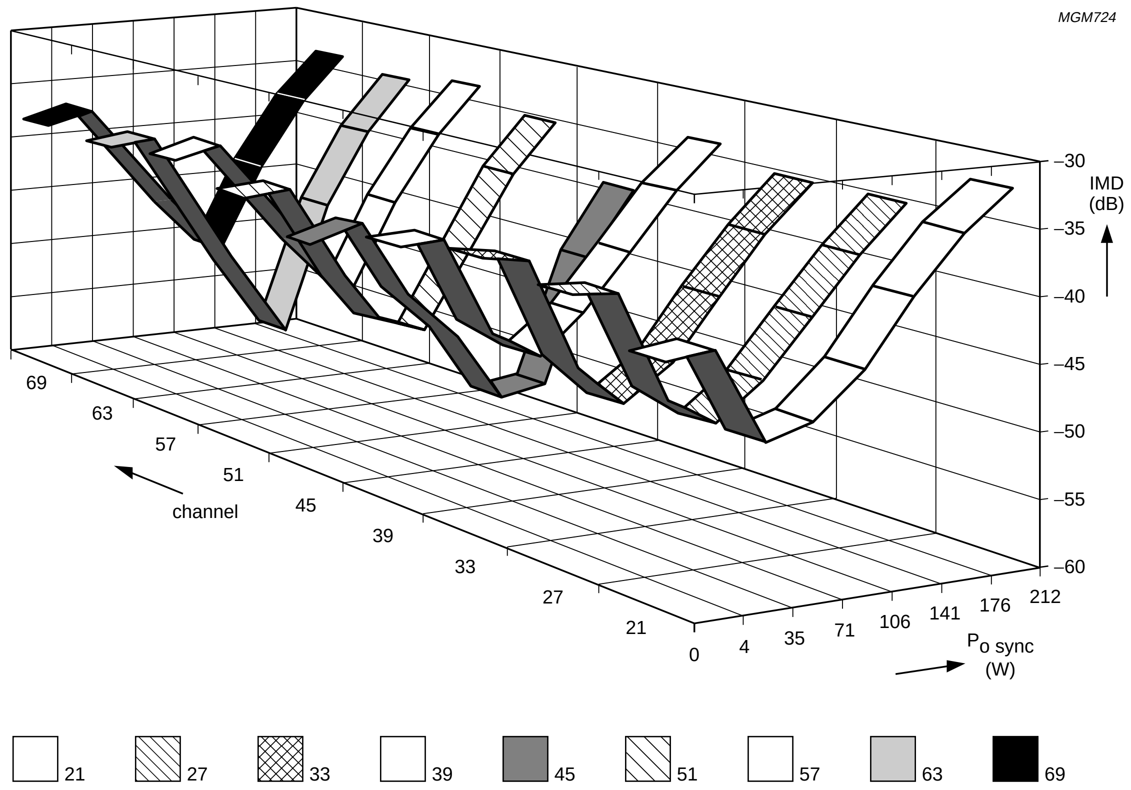


Fig.23 IMD vs. $P_{o,sync}$ level in band IV and V according to system A.

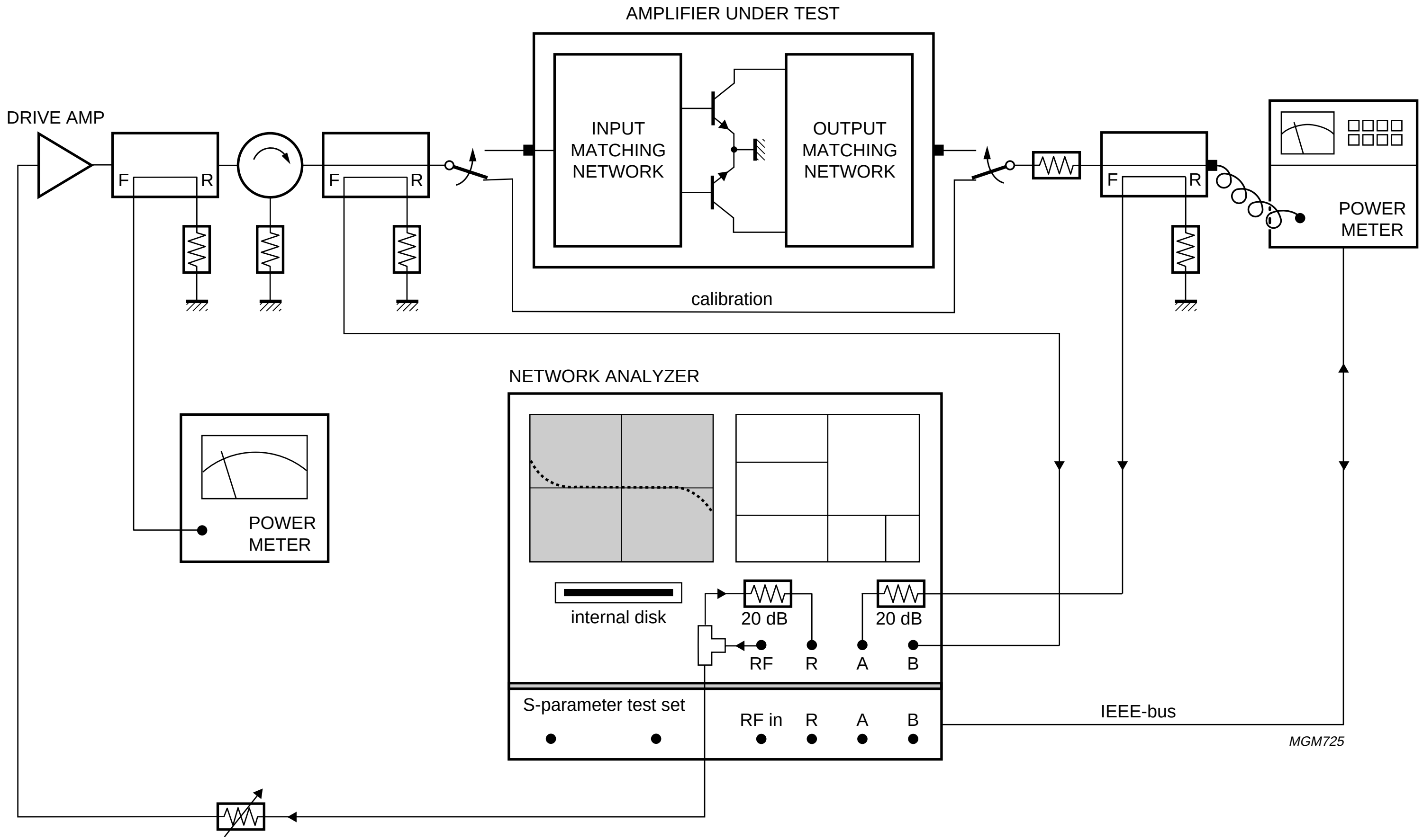
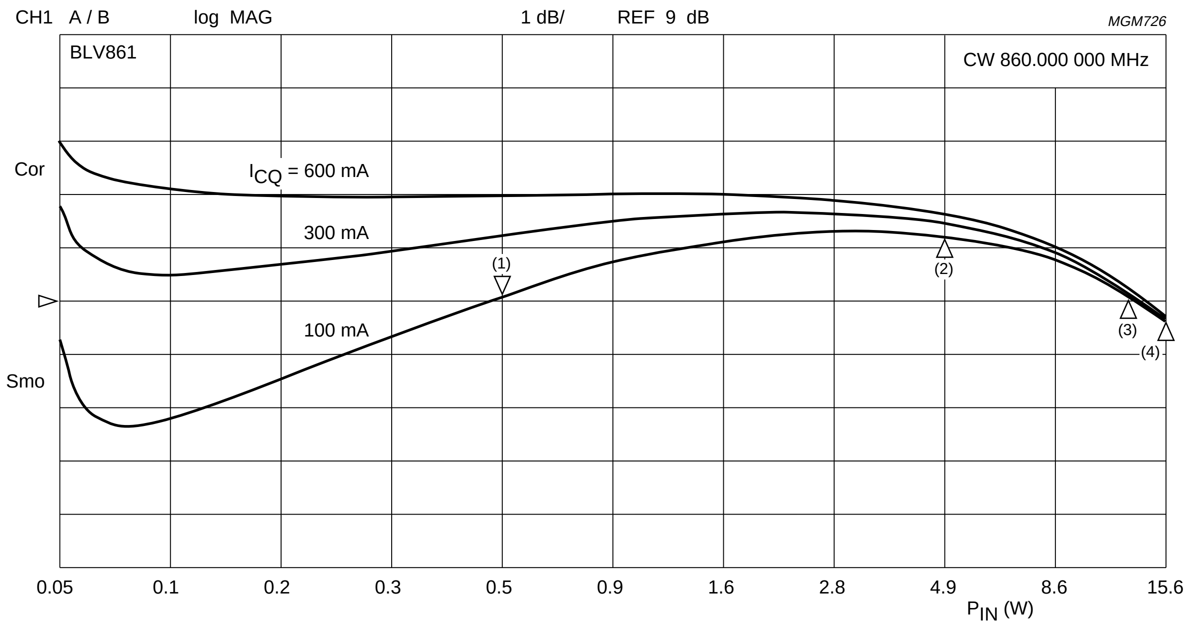
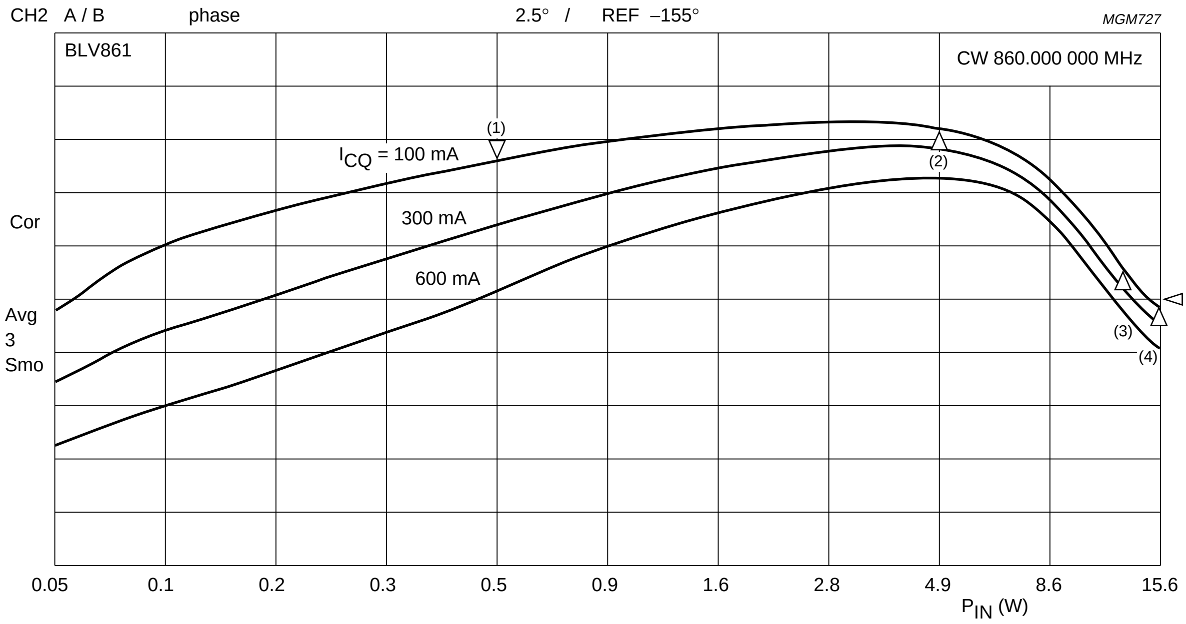


Fig.24 Setup for Power Sweep Measurement.



(1) I_{CQ} = 100 mA.
(2) I_{CQ} = 300 mA.
(3) I_{CQ} = 600 mA.

Fig.25 Amplifier Gain Distortion vs. Input Drive Power.



(1) I_{CQ} = 100 mA.
(2) I_{CQ} = 300 mA.
(3) I_{CQ} = 600 mA.

Fig.26 Amplifier Phase Distortion vs. Input Drive Power.

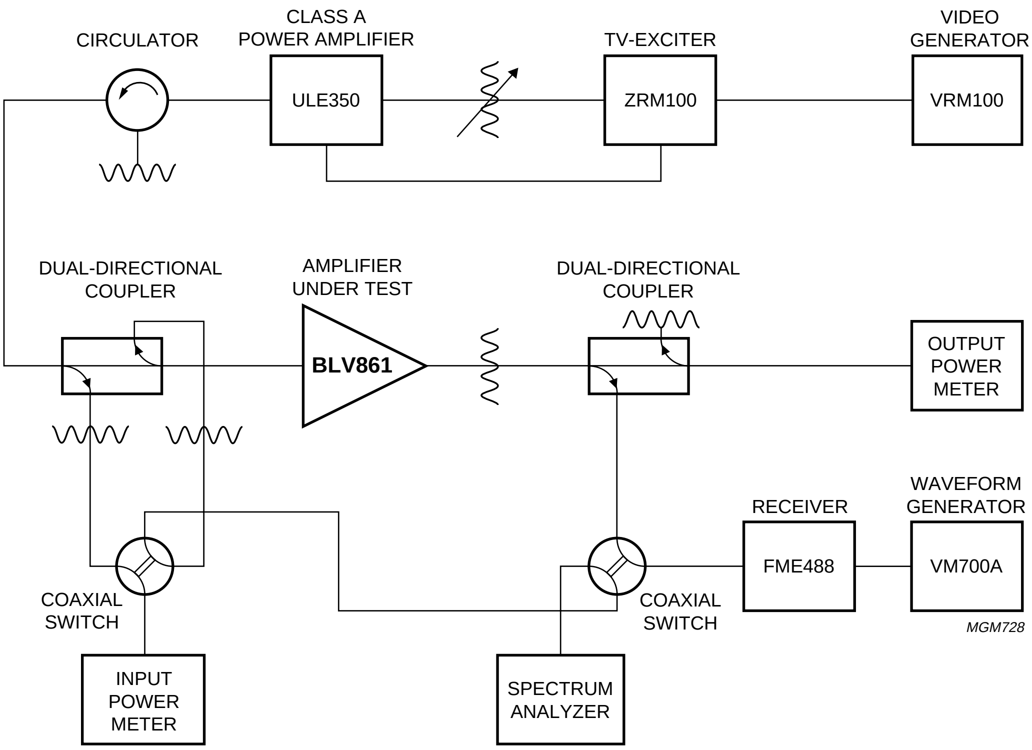


Fig.27 TV Measurement Setup.

Sync Pulse Compression vs. $P_{O,SYNC}$ @ Channel 69.

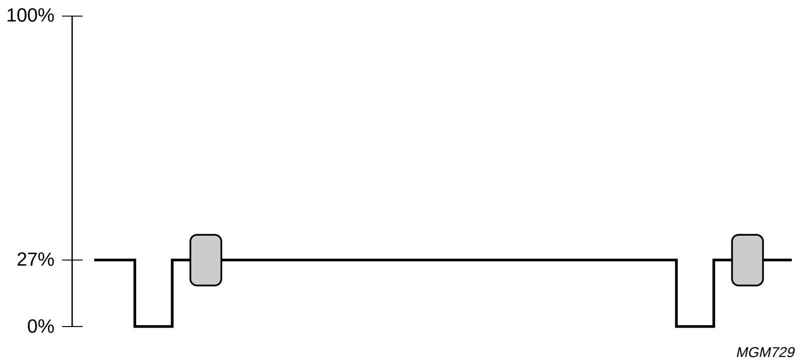


Fig.28 Black picture level (Line 18).

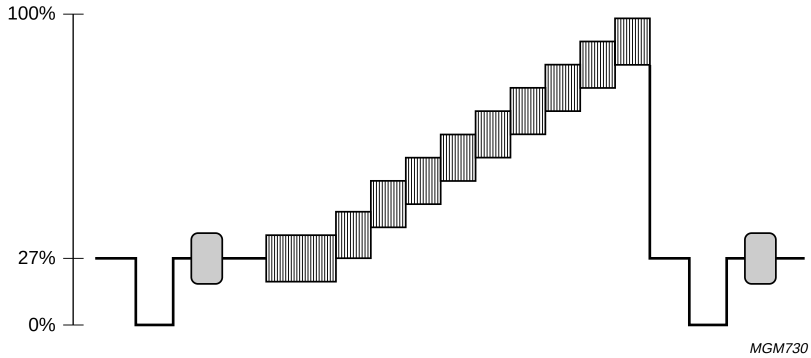


Fig.29 Compisite Video Signal, Modulated 10-steps staircase (Line 8).

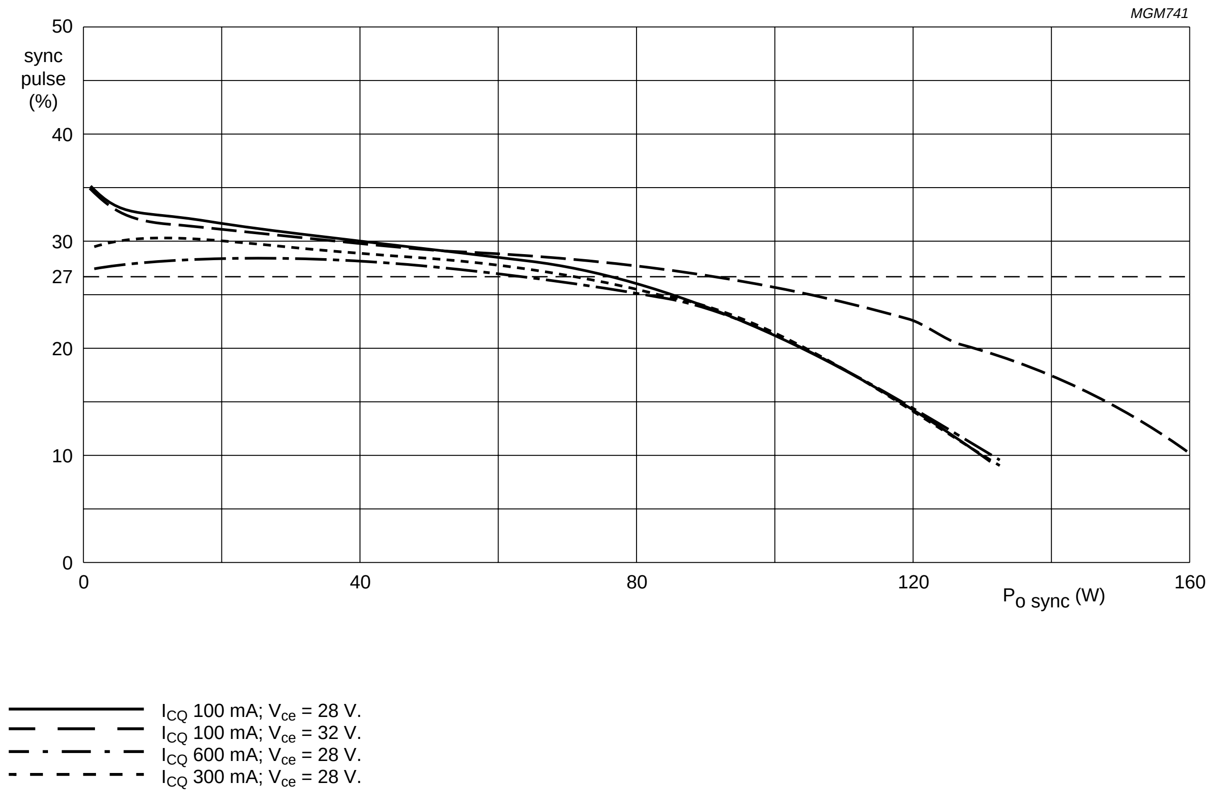


Fig.30 Sync Pulse Compression versus $P_{O,SYNC}$.

Output Sync Power Capability @ Channel 69.

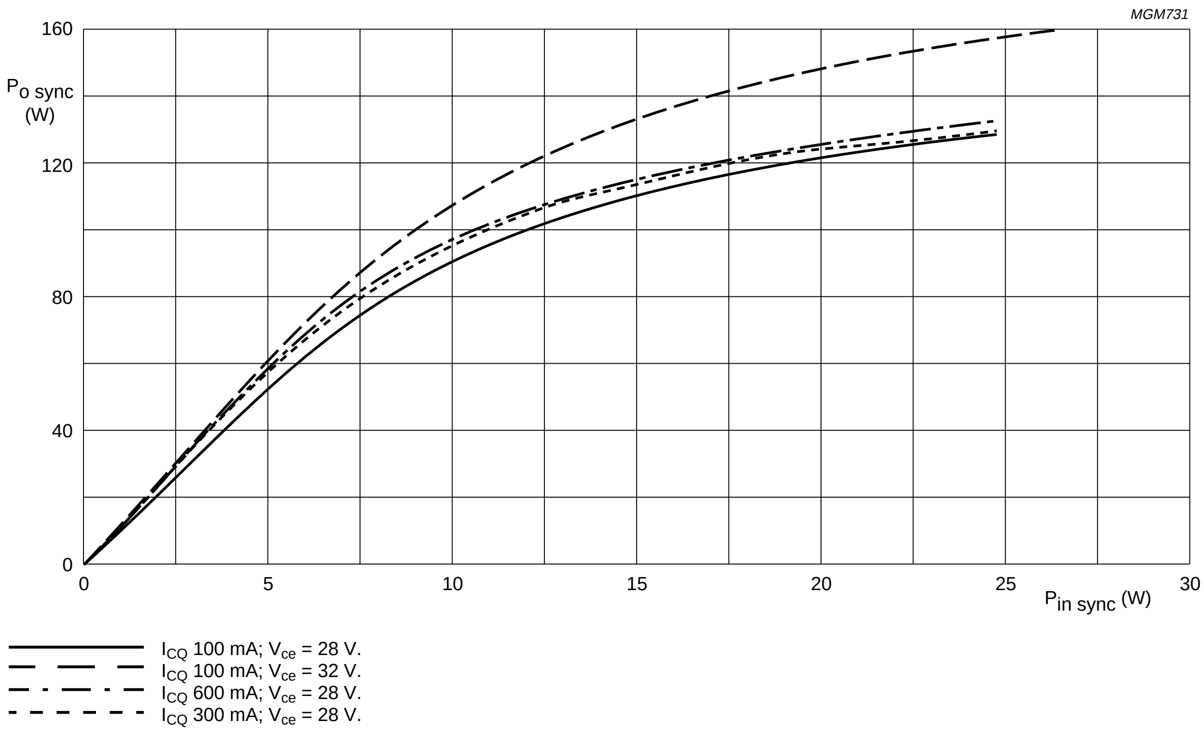


Fig.31 $P_{O,SYNC}$ versus $P_{IN,SYNC}$ @ Channel 69.

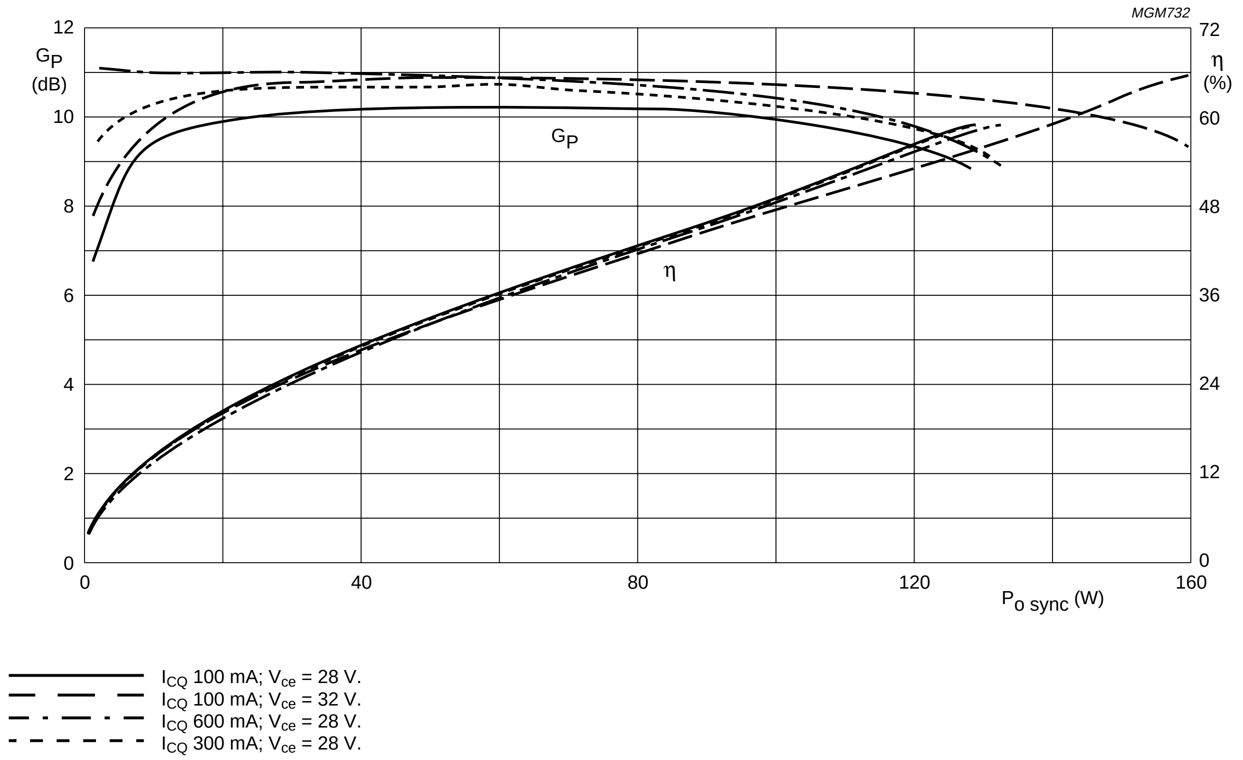
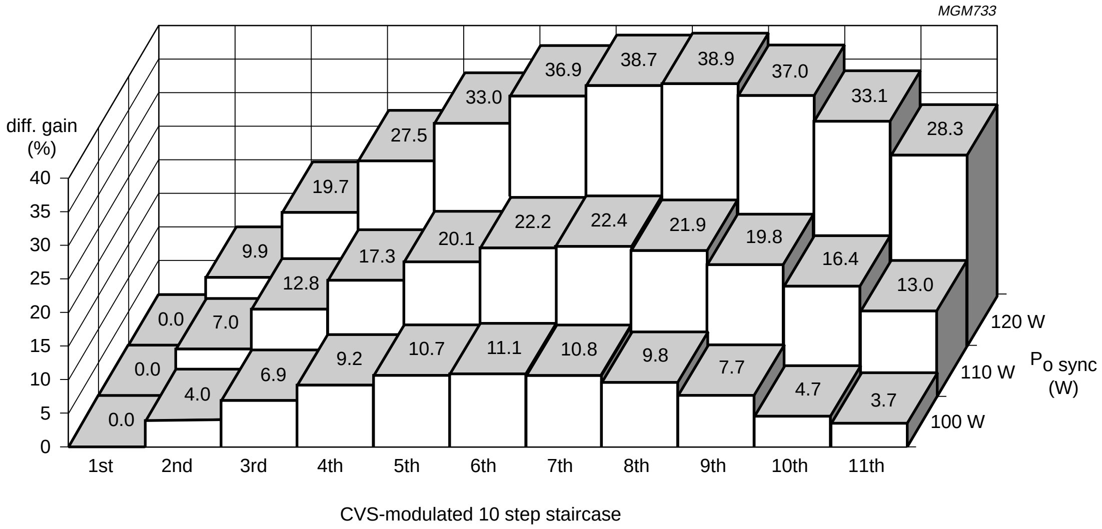


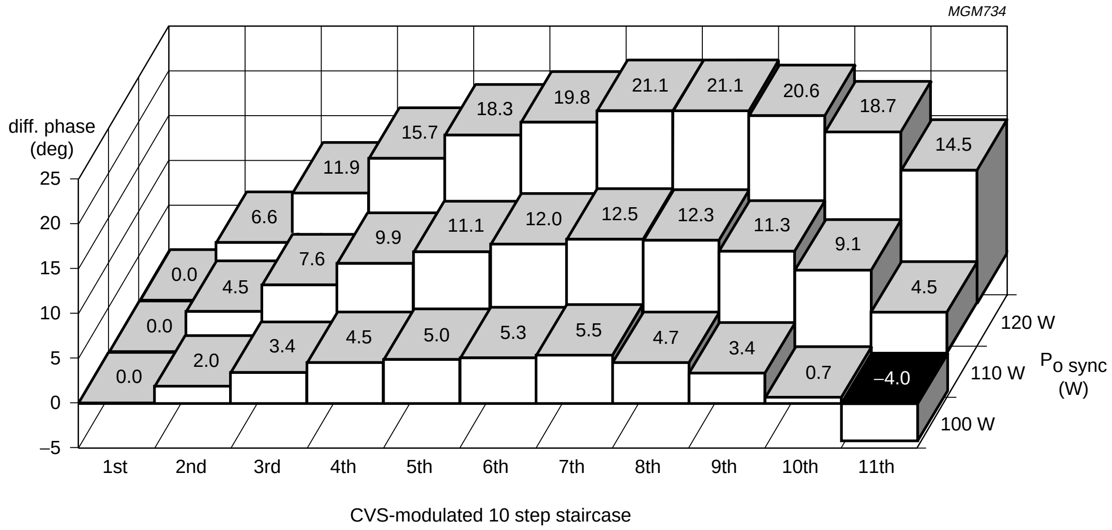
Fig.32 Power gain and collector efficiency versus $P_{O,SYNC}$ @ Channel 69.

Differential Gain and Differential Phase.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 100 mA.
T_{HS} = 25 °C.
Channel = 69.

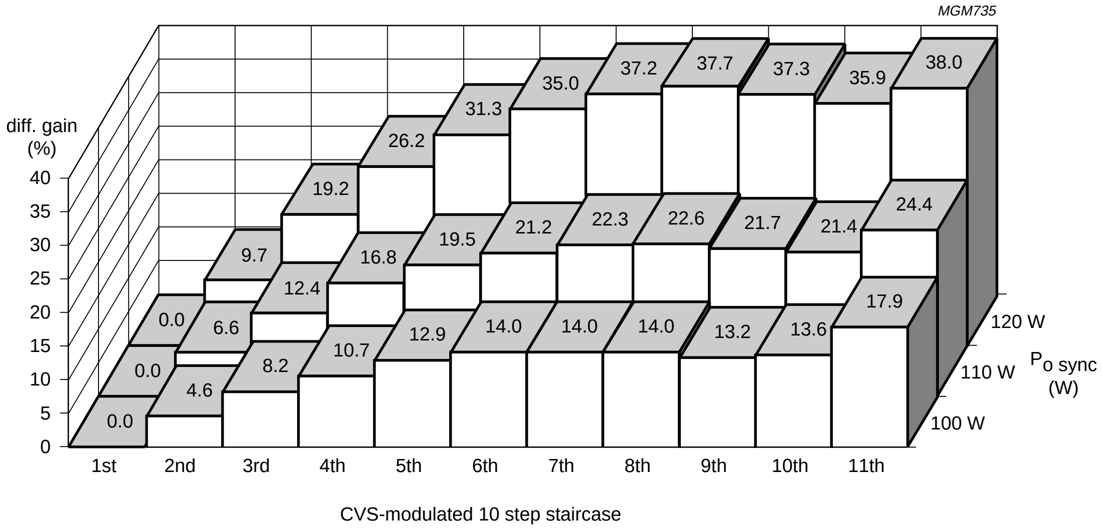
Fig.33 Differential Gain vs. Output Peak Sync Power.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 100 mA.
T_{HS} = 25 °C.
Channel = 69.

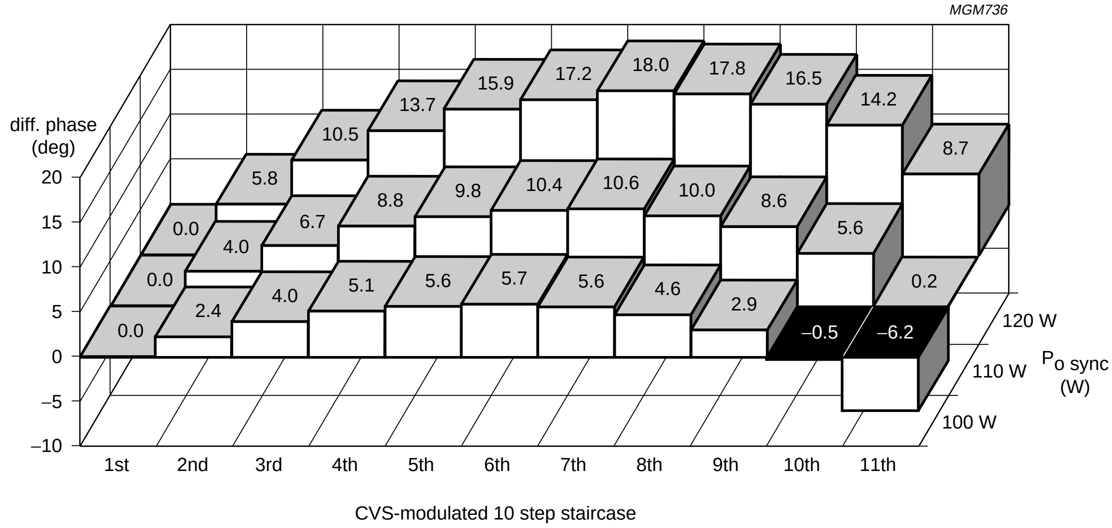
Fig.34 Differential Phase vs. Output Peak Sync Power.

Differential Gain and Differential Phase.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 300 mA.
T_{HS} = 25 °C.
Channel = 69.

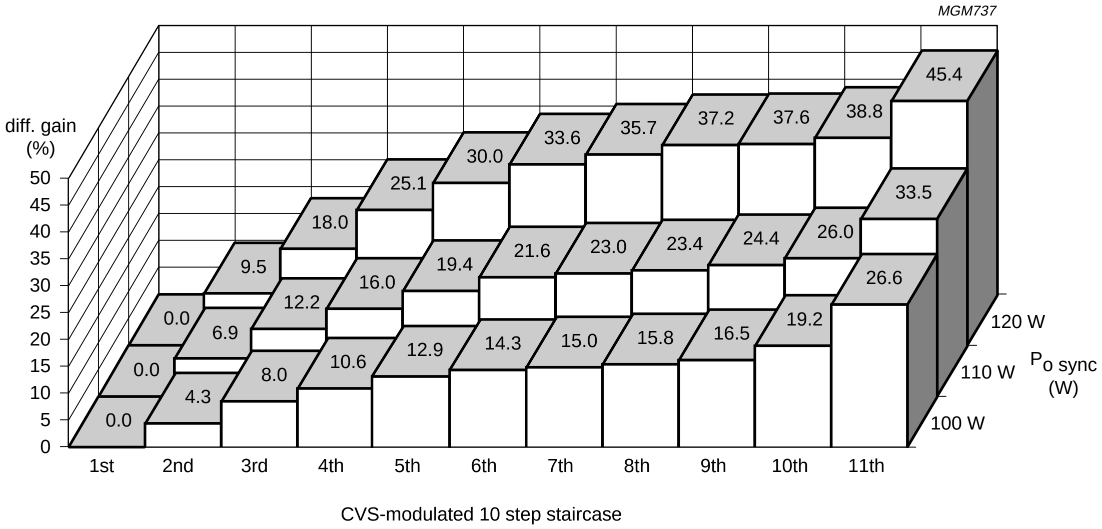
Fig.35 Differential Gain vs. Output Peak Sync Power.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 300 mA.
T_{HS} = 25 °C.
Channel = 69.

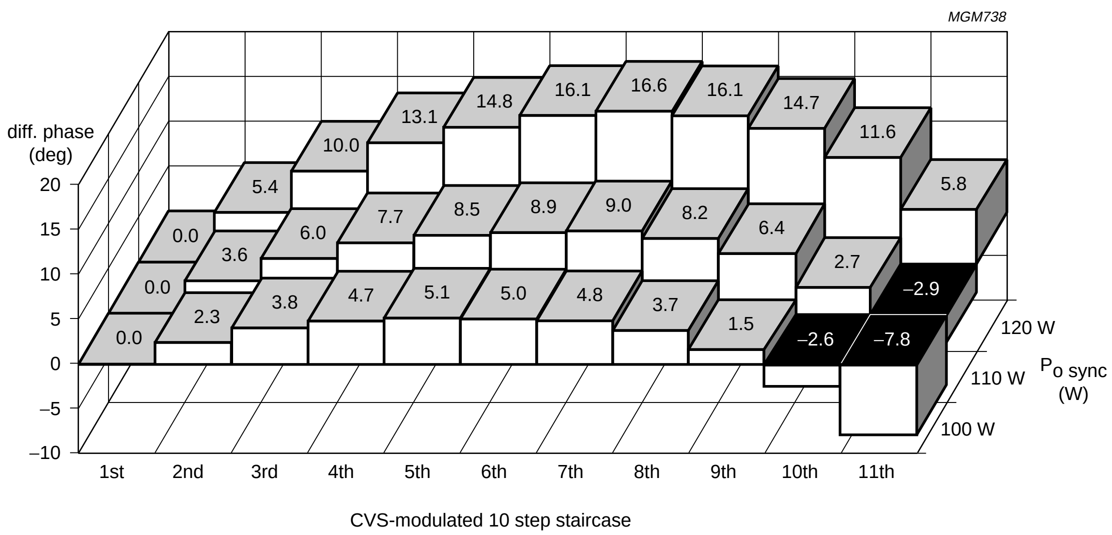
Fig.36 Differential Phase vs. Output Peak Sync Power.

Differential Gain and Differential Phase.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 600 mA.
T_{HS} = 25 °C.
Channel = 69.

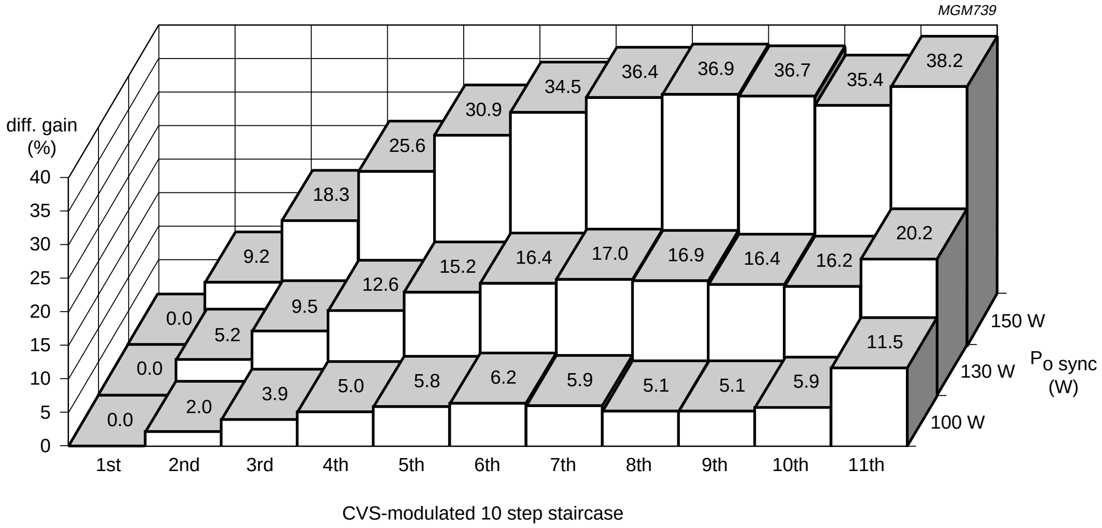
Fig.37 Differential Gain vs. Output Peak Sync Power.



Conditions:
V_{CE} = 28 V.
I_{CQ} = 600 mA.
T_{HS} = 25 °C.
Channel = 69.

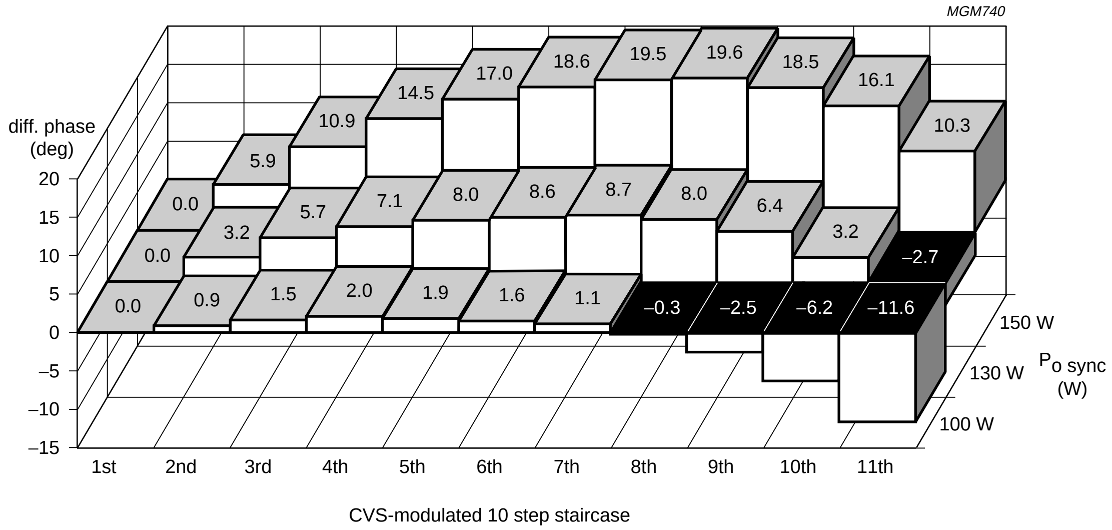
Fig.38 Differential Phase vs. Output Peak Sync Power.

Differential Gain and Differential Phase.



Conditions:
V_{CE} = 32 V.
I_{CQ} = 100 mA.
T_{HS} = 25 °C.
Channel = 69.

Fig.39 Differential Gain vs. Output Peak Sync Power.



Conditions:
V_{CE} = 32 V.
I_{CQ} = 100 mA.
T_{HS} = 25 °C.
Channel = 69.

Fig.40 Differential Phase vs. Output Peak Sync Power.

9 APPENDIX A

Tree Tone and Two Tone Power Levels

Relative power levels of a tree tone system:

$$V_{SYNC} = 1 \qquad V_{VISION} = 10^{\frac{Avision}{20}} \qquad V_{SIDEBAND} = 10^{\frac{Asideband}{20}} \qquad V_{SOUND} = 10^{\frac{Asound}{20}}$$

$$P_{SYNC} = \frac{(V_{SYNC})^2}{R}$$

$$P_{PEAK} = \frac{(V_{VISION} + V_{SIDEBAND} + V_{SOUND})^2}{R}$$

$$P_{AVG} = \frac{(V_{VISION})^2 + (V_{SIDEBAND})^2 + (V_{SOUND})^2}{R}$$

Table 10

SYSTEM	VISION	SIDEBAND	SOUND	$\frac{P_{SYNC}}{P_{AVG}}$	$\frac{P_{PEAK}}{P_{AVG}}$	$\frac{P_{SYNC}}{P_{PEAK}}$
A	0.398	0.158	0.316	3.526	2.686	1.313
B	0.562	0.141	0.316	2.293	2.384	0.962
C	0.708	0.100	0.316	1.636	2.068	0.791

Relative power levels of a two tone system:

$$V_{SYNC} = 1 \qquad V_{VISION} = 10^{\frac{Avision}{20}} \qquad V_{SIDEBAND} = 10^{\frac{Asideband}{20}}$$

$$P_{SYNC} = \frac{(V_{SYNC})^2}{R}$$

$$P_{PEAK} = \frac{(V_{VISION} + V_{SIDEBAND})^2}{R}$$

$$P_{AVG} = \frac{(V_{VISION})^2 + (V_{SIDEBAND})^2}{R}$$

Table 11

SYSTEM	VISION	SIDEBAND	$\frac{P_{SYNC}}{P_{AVG}}$	$\frac{P_{PEAK}}{P_{AVG}}$	$\frac{P_{SYNC}}{P_{PEAK}}$
A	0.398	0.158	5.446	1.687	3.228
B	0.562	0.141	2.975	1.473	2.020
C	0.708	0.100	1.956	1.277	1.532

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

Printed in The Netherlands

Date of release: 1998 Mar 23

Let's make things better.

Philips Semiconductors

PHILIPS