ELIMINATION OF CROSS-MODULATION IN CATV AMPLIFIERS

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Introduction

In this paper we will show how increased channel loading and output capability affects the performance levels of future CATV amplifiers. The expected operating levels of the transistors used in CATV amplifiers tomorrow is shown to be much greater than in the past. Transistor non-linearities causing cross-modulation, second order and triple beat forms of distortion, are discussed in light of increased peak output capability. A new transistor design and its performance in a discrete CATV amplifier circuit is presented.

1. How expansion of channel loading affects CATV amplifier and transistor design.

A review of the history of cross-modulation specifications would appear in order at this point and is shown in Table I. Note that early systems were specified for 9 to 12 channel capacity. The trend since 1964 has been to add pilot, FM, mid and superband channels. The previously shown cross-mod specification necessitates not only higher output capability, so that more amplifiers can be cascaded, but also necessitates increased channel capability. Let's take a closer look at how channel loading affects not only amplifier design but the design of transistors used in those amplifiers. Applying the specifications given for the 1964 amplifier, the peak output power capability per channel is calculated as follows:

Single channel output levels into 75 ohms
+42 dBmV (=) 126 mV (=) 1.68 mA RMS
Average power = E x I = 211 μW
Peak power = $\sqrt{2}$ E x $\sqrt{2}$ I = 422 μW

When more than one channel is present, and, depending upon the phase relationship between the signals present at only one instant in time, the peak output power can exceed the calculated level for a single channel. Although the average power of the output signal is the sum of the average power of each signal present, the peak output power can be several times the average power. We will now see how peak envelope power levels of the 9 channel amplifier compare with today's and tomorrow's amplifier.

Table II gives the calculated output peak voltage, current and

Table I

History of cross-modulation specifications for an all-band transistor mainline amplifier				
Channel loading (flat response)	dBmV output capability for -57 dB XM			
	Year	1964	1968	1972
9		42		
12		41	48	56
20			46	
35				50

Table II

Peak CATV amplifier output levels for -57 dB XM			
Channel loading and level into 75 ohms	Volts (V)	Current (mA)	Power (mW)
9 ch @ +42 dBmV (1964)	1.6	21.4	34.
20 ch @ +46 dBmV (1968)	5.66	75.3	426
35 ch @ +50 dBmV (1972)	15.6	209	3300

power specifications for the amplifiers previously mentioned in Table I. The requirements for increased capacity and output level dramatically increased the peak levels of voltage, as shown in Fig. 1, and power in Fig. 2. It can be expected that the operating levels of the output semiconductors of newer amplifiers will have to provide these peak levels in addition to being linear. The nomograph in Fig. 3 allows us to take a quick look at what the very minimum operating points of a device would have to be. Besides offering second order cancellation, a push-pull output stage provides better efficiency for an additional 3 dB output capability. Assuming a push-pull output stage was needed to meet the 35 channel operating level of +50 dBmV, we next choose the collector load line impedance we expect to operate on. Since the curve was initially constructed for 12 channel operation, the output level contours must be corrected by approximately 5 dB to represent peak device requirements for 35 channel operation. The output contour is thus determined to be (+50 -3 +5) +52 dBmV. On the nomograph the 150 Ω collector load intersects the +52 dBmV locus at V_{CE} = 9.5V and I_C = 65 mA.

In addition to these specifications the end item user expects improved noise figure and efficiency so that he may be able to cascade more amplifiers and lower his cost per mile. Cost and reliability are always important in consumer service and these factors find their way back to the equipment and component manufacturer as tougher specifications.

As a semiconductor manufacturer TRW has been providing discrete devices and hybrid function modules specifically to the CATV industry for several years. The realization of the increased need for output capability, which we have just shown, caused TRW to accelerate its intensive R&D effort on the development of transistor devices capable of meeting the future industry needs while continually upgrading present products. To do this we started our investigation at the beginning; the design of a high power, high frequency, linear amplifier.

Once an amplifier is designed it is constructed with mostly passive components. Resistors, capacitors and inductors used in CATV amplifiers are normally linear devices and do not directly contribute to the resulting distortion levels. The most generally applied technique for lowering distortion is the application of negative feedback. Since we know the feedback will reduce distortion, stabilize operating point, and increase frequency response, we expect that some form of feedback will be used. If a single stage, for example, exhibits 1% distortion and we can apply 10 dB of feedback, the distortion and gain is reduced by 10 dB. It's obvious that it would be better to have 0.1% distortion to start with and reduce that initial level by 10 dB with the feedback. The approach taken by R&D at TRW has been to provide the semiconductor division with the design of an ultra-linear transistor with increased dynamic range. This device design has set new standards of performance in the reduction of distortion in the form of cross-modulation, second order and triple beat. In order to develop this highly linear microwave transistor, an in-depth investigation into all of the non-linearities of a bipolar transistor was initiated.

2. Transistor non-linearities

2.1 Relationship of output capability and dynamic load line

The minimum transistor bias requirements for the above mentioned +50 dBmV output capability were presented in Fig. 3. Next we transfer the bias point (A) to a set of over-idealized transistor transfer characteristics and construct a dynamic load line, slope equal to $-1/RL$, with the expected peak voltage swings, Fig. 4a. Having completed this operation it becomes apparent that when the signal swings positive, increasing voltage, it will transverse to a V_{CE} voltage of zero, provided the device had no limiting resistance. However transistors, like tubes, have their limitations. A slightly more representative set of device curves with a load line is shown in Fig. 4b. Note that for the same bias point, when the signal swings positive, it will transverse into a very nonlinear current region and when decreasing, the signal will be limited by the saturation resistance of the device. These nonlinear regions, cutoff and saturation, cause severe clipping of the signal and can be avoided by increasing the bias current and voltage to the point C where the expected voltage and current swing is well within the capability of the device. All quality CATV amplifiers are designed with bias points well within the device characteristics and are analyzed over the expected temperature range to insure bias point stability.

2.2 Temperature variation

The temperature of the active portion of the device is given by,

$$T = (P_{bias} + P_{signal})\theta_{jc} + T_s$$

and

$$dT = (dP_b + dP_s)\theta_{jc}.$$

We have chosen the bias point and load line to follow a constant power contour so that $dP_b = 0$. The impact of synchronous multi-channel operation is seen in that dP_{signal} is the power in a single channel times the number of channels. The relative change in temperature is then:

$$dT/T = -NdP_s\theta_{jc}/(V_C I_C \theta_{jc} + T_s),$$

since P_{signal} is initially zero and dP_b is made zero by the choice of load line. The change is negative since power is removed via the output signal. The change in dB relative to the operating temperature is plotted versus output level for 9, 12, 20, and 35 channels in Fig. 5. The impact of this change on various parameters can be significant at low cross-modulation levels. For example, Fig. 6 shows the temperature variation in resistivity of typical collector and base material.

2.3 Collector non-linearity

Nearly all CATV amplifier circuits are designed to have minimum capacitance (junction, metal, stray) projected across the collector circuit in order to obtain the wide bandwidths required. However there is another reason to minimize the output circuit capacity. Where a high frequency signal is present, the collector will see a reactive load. Owing to the angle between the sinusoidal current and voltage in reactive loads, the collector current will not have the same value for increasing collector voltage that it has for decreasing voltages. The load line is a form of an ellipse, varying from a straight line for a pure resistive collector load to a circle for loads of pure reactance. Fig. 7 shows that for extreme swings the elliptical load line can swing into nonlinear regions of the device. Microwave power transistors exhibit two fixed output capacitances plus junction and stray capacitances which degrade second order, cross-modulation, and triple beat performance.

We have discussed the importance of minimizing fixed capacitance to reduce the elliptical load line. The forms of fixed capacitance significant in bipolar transistors are metal over oxide (MOS) and header or case capacitance. Great care has been taken by semiconductor manufacturers in the design of their devices to reduce the area of metallization over silicon oxide and make the silicon oxide as thick as possible to reduce the basic chip capacitance. Since a poor package can often have much more stray capacitance than the chip exhibits, new stripline packages with low parasitic lead inductance, case capacitance, and excellent thermal characteristics were developed.

The properties of any semiconductor junction are a function of the voltage across it and the current through it. The collector-base junction capacitance is inversely proportional to the collector emitter voltage and barely independent of the collector current at the low level class "A" bias for linear CATV. A large V_{CE} bias is desired to minimize C_{ob} as well as dC_{ob}/dV_{CE}, the differential of C_{ob} with respect to V_{CE}.

The collector capacitance is the result of the parallel plate analogue of the widening collector-base depletion region. This widening gives rise to a non-linearity in the collector resistance, r_c, which is a resistor with length equal to the epitaxial thickness minus the junction depth and the depletion width. To minimize the resistance, the epitaxial layer should be barely thick enough to contain the depletion width plus a little cushion for second breakdown resistance. Again a large V_{CE} is desired to reduce r_c and to minimize dr_c/dV_{CE} since the depletion width variation is least at high V_{CE}.

The magnitude of C_{ob} could be reduced by using a higher collector resistivity; but that would be at the expense of the collector resistance and the allowable current density, which hasn't been a good tradeoff. Actually, device ruggedness requires thicker, higher resistivity epitaxial layers at the expense of performance.

As noted earlier, when power is taken from the device the temperature drops and the result is a decrease in r_c. This will be reflected as a decrease in $V_{CE(SAT)}$ and an increase in f_T, both of which cause a gain expansion and second order distortion.

2.4 Emitter-base non-linearities

There are two identified non-linearities associated with the emitter base junction, a shift in the contact potential, V_{BE}, and the multifaceted effects of temperature. The microwave transistors used in CATV amplifiers have the injection shift due to crowding from the bottom of the emitter to some point up the side, where the concentration and hence V_{BE} is higher. This shift in injection and V_{BE} is effectively suppressed at typical bias currents by the use of a shallow diffused emitter with a second contact. This forces injection at the bottom of the emitter away from the edge. It also avoids a premature roll-off of β and f_T with current as a result of a wider base when injection shifts up the side. The simple diode equation becomes:

$$I_e = I_o(T)\exp(qV/kT - \ln(N_{EBj}/Ni(T)^2))$$

and one realizes that saturation current, the nominal 26 millivolts, and the intrinsic concentration are all functions of temperature.

First order contribution of the base-emitter non-linearity to second and third order distortion has been observed and analyzed in several papers.[1,2] By observing the familiar ideal diode current-voltage relationship, Fig. 8, we would expect the distortion due to operation at about point A for a small excursion in base-emitter voltage to be very small as the curve at this point appears very straight. But, we are talking about distortion on the order of 100 dB down and that is approximately 0.001% non-linearity. Note the nonlinear exponential characteristic of current with respect to voltage, in the diode equation.

From the equations of Mallinckrodt[1] we have generated a family of curves representing the degree of second and third order distortion as a function of both emitter current and device current gain, Fig. 9, 10.

As can be seen from Fig. 10 it appears that the third order distortion term disappears at unique values of current gain and emitter current. (Mallinckrodt also shows this sharp null is dependent upon source resistance.) Two important factors obtained from Figs. 9 and 10 are the reduction in second and third order distortion due to low current gain and high emitter current. These conclusions are verified from our knowledge of the ideal diode current-voltage relationship shown in Fig. 8. Higher current gain implies lower base current drive (point B) operating in the more nonlinear base-emitter region. Likewise higher current requiring more base drive forces the device to operate in the more linear region of the base emitter junction characteristic.

2.5 Power gain non-linearity

Cross-modulation and other distortions are the result of a non-linear power transfer curve. The factor relating the output power to the input power is the power gain of the transistor amplifier, which is given by,

$$P_g = f_T/8\pi f^2 r_b' C_c$$

The f_T is related to the low frequency beta by,

$$f_T = \beta_c f/(1 + jf/f_T).$$

It can be seen from the equation that not only does beta roll off with frequency, but there is also a phase shift. Upon expanding the power transfer curve into a power series, many of the coefficients normally considered real will be imaginary leading to additional null possibilities. The obvious design goal is a high f_T to reduce the beta roll-off and phase shift. For an f_T of 4 GHz the beta roll-off is pushed out past the superband and the phase shift is only a few degrees. The power gain is then:

$$P_g = \beta_o/8\pi f(1 + jf/f_T) r_b' C_c.$$

The negative temperature shift when signal power is removed reduces β_o resulting in a compression and cross-modulation. The effect of frequency will result in compression in the high band unless appropriate feedback is employed.

Of particular interest are the effects of I_C and V_C on the $r_b' C_c$ product. With increasing V_C, r_b' increases via base narrowing, giving rise to gain compression and C_c decreases giving rise to gain expansion. With increasing current r_b' can decrease, with base widening, giving rise to gain expansion. Increasing I_C can also decrease C_c for further gain expansion. If the magnitudes of the distortions are nearly equal and opposite, nulls can result from increasing V_C and be retraced as the result of increasing current. Shallow diffused processing with higher base concentrations reduces dr_b'/dV_{CE} and dr_b'/dI_C since there is less depletion of the collector-base junction into the base region. Again high V_{CE} lowers C_c and dC_c/dV_C.

2.6 S-parameter contour evaluation

The distortion parameters, previously described, characterize the device in terms of measured quantities, such as, h_{FE}, C_{ob}, etc., at lower than operating frequencies.

Muller indicated that a good measure of output capability, i.e., linearity, could be obtained by taking the forward S-parameters $|S_{21}|^2$ of a device over the expected current and voltage operating range and frequency. He showed that the more non-linear the $|S_{21}|^2$ parameter, forward transducer gain, the more output distortion would be present. Carrying this concept further, we can project a load line on the forward transfer characteristics $|S_{21}|^2$ to obtain a relative feel for its non-linear contribution. As an example Fig. 11 shows the $|S_{21}|^2$ parameter for a low-noise small-signal MATV device. The load line required for 35 channel output level intersects several constant contours indicating the relative distortion contributed by this parameter. The next figure represents the constant $|S_{21}|^2$ contours for a medium power microwave transistor normally used in a class C mode. Although its non-linear contribution is less than the previous device it is still too large to be of use in low cross-modulation amplifiers. In the analysis used at TRW to develop an ultra-linear RF transistor we have gone several steps further. These steps are the analysis of the behavior of all four S-parameters at band edges and over the expected operating range. The reason this somewhat elaborate analysis is taken, is to determine which of the previously mentioned non-linearities is the most dominant one at a particular operating point.

The $|S_{21}|^2$ contour of the CATV device shown in Fig. 12 exhibits much greater linearity over the load line than either of the two previously shown devices. The analysis now proceeds to the investigation of the effects of the other S-parameters. The S_{11} and S_{22} contours, shown in Fig. 13, represent the amount of signal reflected by the device at certain voltage and current bias points. Projecting a typical load line over these input and output characteristic curves and visualizing several instantaneous operating points we obtain a feel for the amount of distortion contributed due to the non-linear input and output behavior.

S-parameter characteristics allow us to measure the combined effects of collector-saturation, cutoff, current gain non-linearity, conduction modulation, etc., at the operating frequencies. By further analysis of these plots, using the ideas presented in the work of Linvill and Gibbons[3], we have been able to assign the dominant distortion contributing factors in the device. What we have learned through all of this analysis has been applied to the design of our almost totally cross-modulation free SLAM, a Super Linear And Microwave transistor.

3. The SLAM transistor

The SLAM transistor is a matrix pattern, which has an emitter like an expanded tic-tac-toe diagram with base contacts in the openings of the matrix or grid. The emitter perimeter to base area ratio is 7 for good microwave performance and operation over the CATV band. The photolithography of 0.1 mil lines and 0.07 mil space has produced good yields. As stated earlier the emitter is contacted through a final oxide layer rather than being contacted through the emitter diffusion opening as other microwave transistors are. This produces more linear operation by forcing injection uniformly across the bottom of the emitter.

Shallow diffusions with high base and emitter sheet resistances are employed to achieve better base transport. The steeper diffusion front of the shallow base diffusion also serves to reduce the collector-base depletion back into the base region. The highest collector concentration consistent with device ruggedness toward second breakdown and energy surge, further suppresses collector-base depletion back into the base region. The higher emitter sheet resistance, in conjunction with the contacted emitter pattern provides significant ballasting for balanced operation of this microwave transistor at CATV frequencies The metallizing on the device is currently a long-life nichrome-aluminum with more than adequate cross-sectional area and a glass overcoating to avoid electromigration and other reliability problems. A tungsten-gold metallizing system, for even longer operating lifetime is nearing completion in R&D.

4. Distortion performance of the SLAM in discrete CATV amplifier circuits.

The proof of a successful transistor chip design is left in the hands of the circuit designer. That circuit designer usually has little or no knowledge of what has gone into the development of the device. Unique in the industry is the approach taken at TRW where device and circuit engineers work together toward a mutual understanding of the important distortion factors and their reduction.

To evaluate the performance of SLAM devices it was first necessary to characterize it by S-parameters as shown earlier. The parameter values were then combined with a circuit design computer program to optimize the feedback for gain flatness and input/output impedance match to 1000 MHz. Because the device exhibits considerable current gain phase shift over frequency small external matching elements are added to further improve flatness.

Initial cross-modulation performance of the design is shown in Fig. 14 thru 16. Device XM-28 had a 12 channel flat cross-modulation performance better than -65 dB on both channels 2 and 13. The figures indicate that the optimum voltage and current ranges for maximum performance are very broad and the device appears well behaved.

Mentioned previously were the possible effects of temperature and thermal modulation in producing cross-modulation. By lowering the junction temperature and forcing the thermal cross-modulation upward, there appears in cross-modulation/current curves an improvement in cross-modulation for 16 volt V_{CE} and bad behavior at 17 and 18 volts. This is due to the fact that some of thermal components of cross-modulation exhibit a phase term in their coefficient. The various phases change as a function of collector voltage, current, output level, temperature and frequency. It should not be suprising by now to imagine one instant in time an upward modulation caused by low-band signals and a downward modulation caused by superband signals resulting in an apparent cross-modulation level of -100 dB. This could also be accomplished by changing the form contribution of thermal modulation to null out base-emitter non-linearity. This was done as shown in Fig. 16 where the additive cross-modulation components at a poor bias point were modified by changing the duty cycle and repetition rate of the modulation. This reduced the thermal modulation effect of the junctions and resulted in cross-modulation level of -90 dB, Fig. 17.

It's apparent that the complex cross-modulation levels measured here do not reflect the cross-modulation produced only by a real third order non-linearity term. A truer indication of the magnitude of the third order contribution is obtained by measuring the triple beat performance of the device. The results of this test, Fig. 18, indicate that the triple beat characteristic was smooth, no nulling behavior, and that the best triple beat was close to the best cross-modulation level. We would expect that the triple beat level would be slightly more than 20 dB below the measured cross-modulation as shown by Simons[4]. This is if the cross-modulation is dominantly third order and phase independent.

It has been brought out by some that cross-modulation due to thermal modulation is not as important as triple beat performance. They argue that in actual usage the device will never experience synchronous modulation and therefore the thermal modulation contribution will be zero. We caution purchasers of equipment and devices against adopting this view. Besides the major networks many independents are using atomic clocks to derive their timing standards. This means that the chance of true synchronous modulation is ever increasing and that the worst case NCTA cross-modulation test for output capability is still a very valid one.

Although we've limited most of our discussion to cross-modulation effects, a device with only good cross-modulation is useless. Along with reducing cross-modulation we were able to design and control the device parameters to reduce second order distortion at a particular bias point. At some bias points the base emitter second order non-linearity will cancel the collector-base second order contribution giving a null or minimum in the distortion level as shown in Fig. 19. We are also able to guarantee this null point when the device is used properly over the complete operating band of frequencies. Of course as shown by the curves the relative width of the null for a given second order level is dependent upon both voltage and current bias points. Care must be taken by circuit designers to insure that the bias circuits are capable of providing the degree of bias stability required over temperature according to the level of performance desired.

A single stage transformerless amplifier design shown in Fig. 21 performed as follows:

Frequency response:	5 MHz to 1000 MHz
Gain:	7 dB
Gain flatness:	±0.5 dB
Cross-mod @ +50 dBmV:	-85 dB Ch 2 and Ch 13
2nd order @ +50 dBmV:	-76 dB Ch 2
	-72 dB Ch G
	-74 dB Ch 13
V_{CC} = +24V, I_C = 100 mA	
Return loss:	>16 dB 50-250 MHz

Two stages were combined to evaluate the additive effects on performance and to obtain a minimum gain of 15 dB over a somewhat smaller bandwidth. Because of the stray circuit elements the flatness and input match are not indicative of what can be obtained when this circuit is transformed to hybrid form where the parasitics are designed into the matching and gain flatness networks.

Two stage performance:

Frequency response:	5-1000 MHz
Gain:	18 dB
Gain flatness:	±0.5 dB
Cross-mod ± +50 dBmV:	-82 dB
2nd order @ +50 dBmV:	-66 all three channels
V_{CC} = +24V, I_C = 196 mA	
Return loss:	>18 dB 50-250 MHz
	>16 dB 250-500 MHz
	>10 dB 500-1000 MHz

Next a four stage circuit was fabricated and exhibited the following characteristics:

Frequency response:	5-300 MHz
Gain:	30 dB ±0.5 dB
Cross-mod @ +50 dBmV:	-72 dB Ch 2
	-78 dB Ch 13
2nd order @ +50 dBmV:	-60 dB Ch 2
	-64 dB Ch G
	-64 dB Ch 13
N.F.:	9.4 dB Ch 13
V_{CC} = 24V, I_C total = 200 mA	

Return loss:	>18 dB	50-250 MHz
	>16 dB	250-500 MHz

5. Conclusions

This work serves to demonstrate that distortion (cross-modulation, second order, triple beat, etc.) in CATV amplifiers is caused by many mechanisms, all of which must be suppressed simultaneously to achieve the desired low levels of distortion. If one form of distortion is severe,suppression of all the others does not show in the result. If two forms remain significant and one is rising while the other is falling, as a function of voltage, current, frequency or temperature, nulls appear in the results. At TRW, we have identified (1) power loading as a cause of temperature variation, (2) signal swing around the dc bias as a variation in the instantaneous characteristic of the transistor, and (3) bandwidth as a frequency variation in the instantaneous characteristic of the transistor as the primary distortion producing mechanisms. We then designed the SLAM transistor to suppress the effects of temperature, V_C, and I_C on the transistor characteristic. Finally the circuit, to match the transistor, was designed to provide a flat response over the CATV band, to reduce the remaining distortion, and to compensate the phase over the CATV band via feedback. The transistor is packaged in a low parasitic stripline stud to preserve the thermal design.

References:

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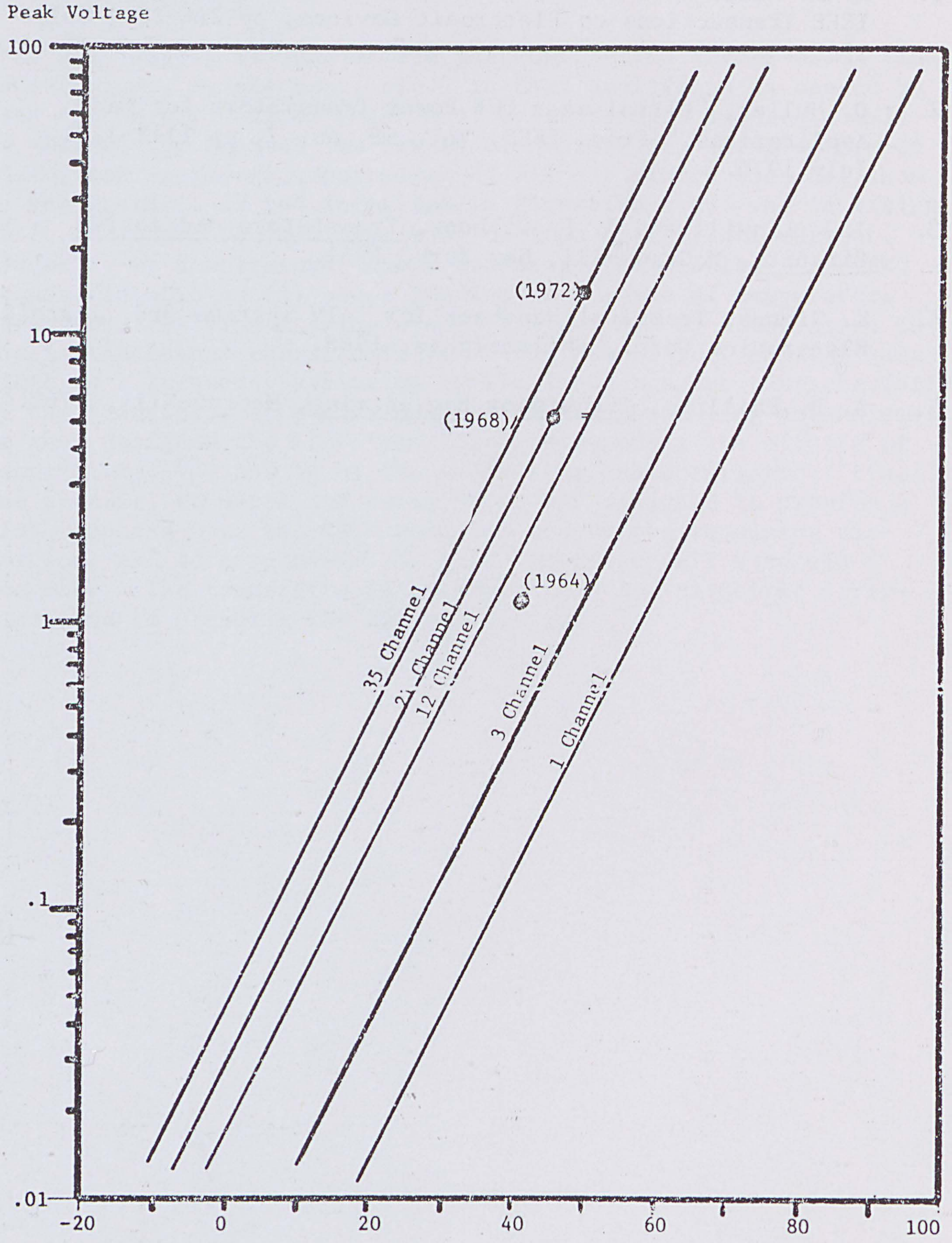


Fig. 1. Peak RF voltage across 75 ohms as a function of output level and channel loading.

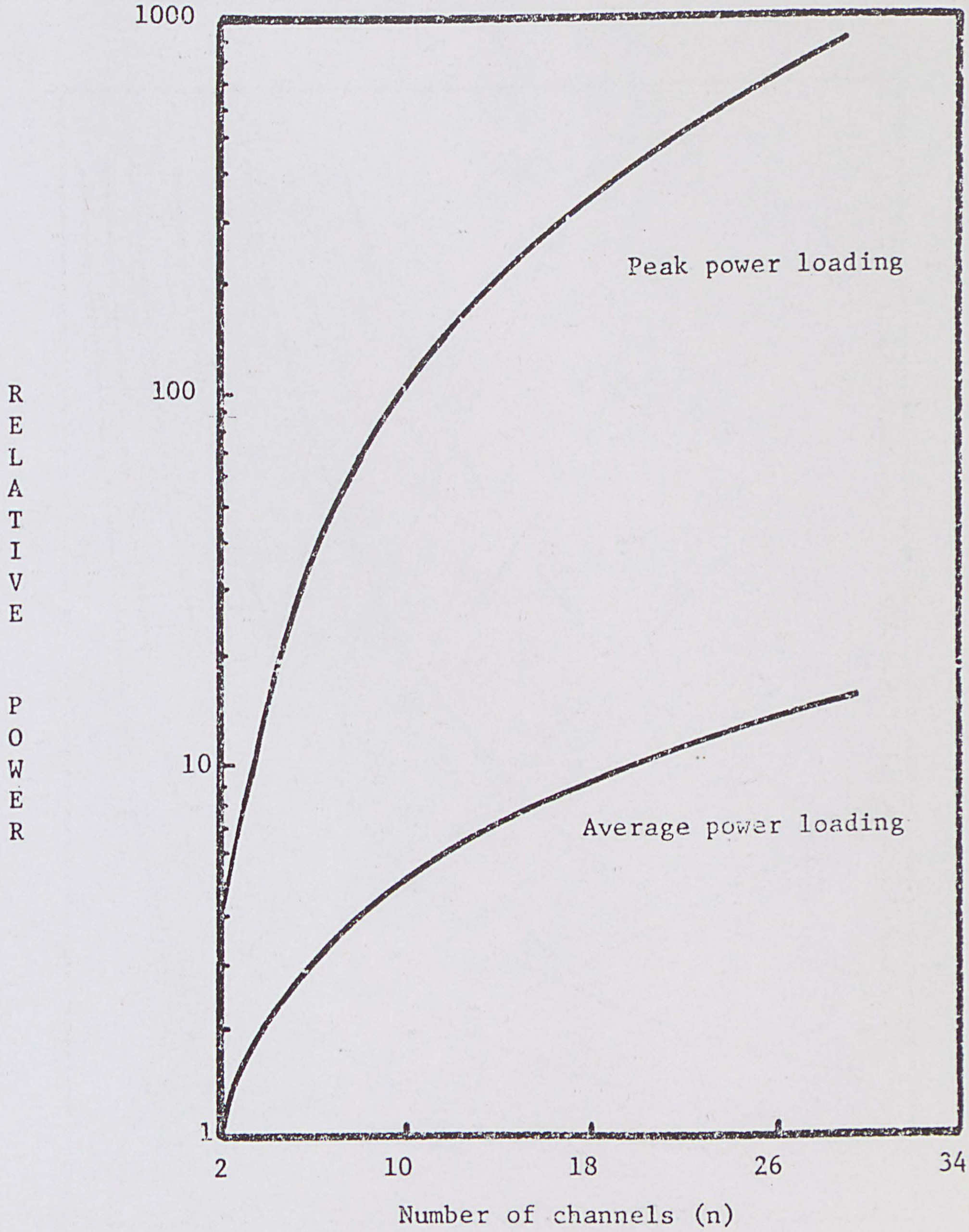


Fig. 2. Normalized peak and average power loading as a function of number of channels present.

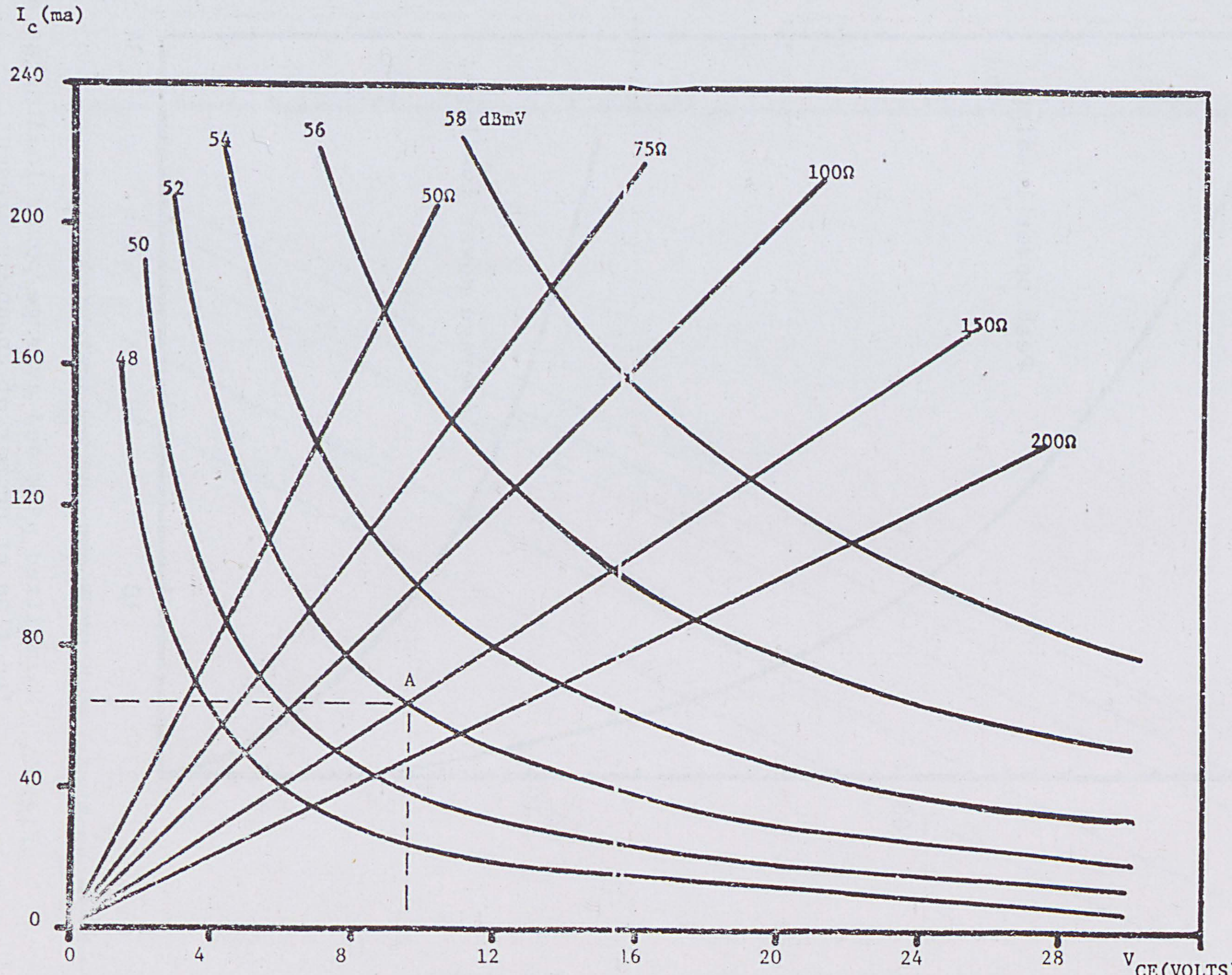
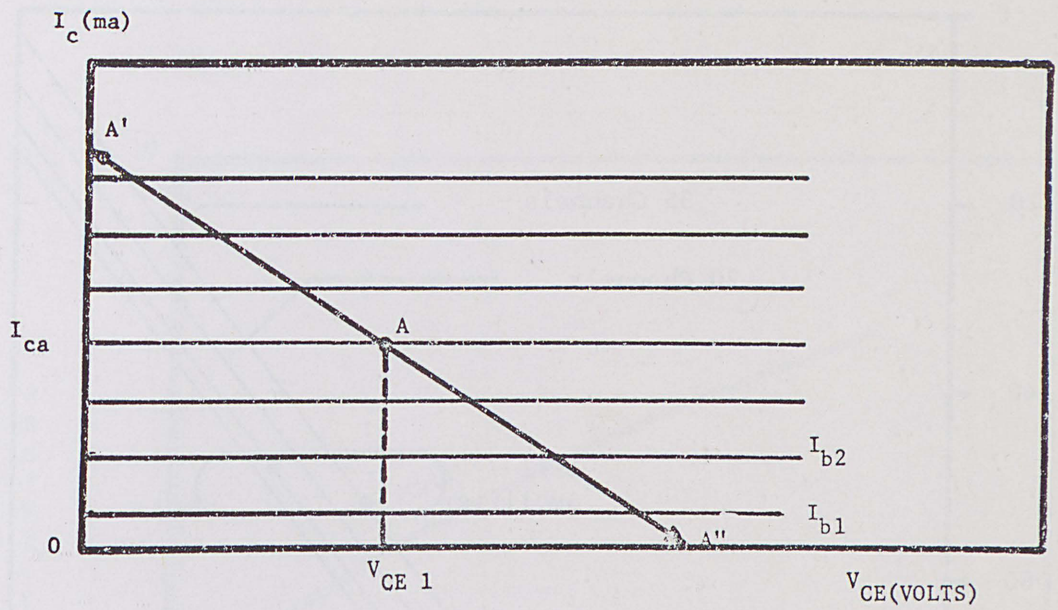
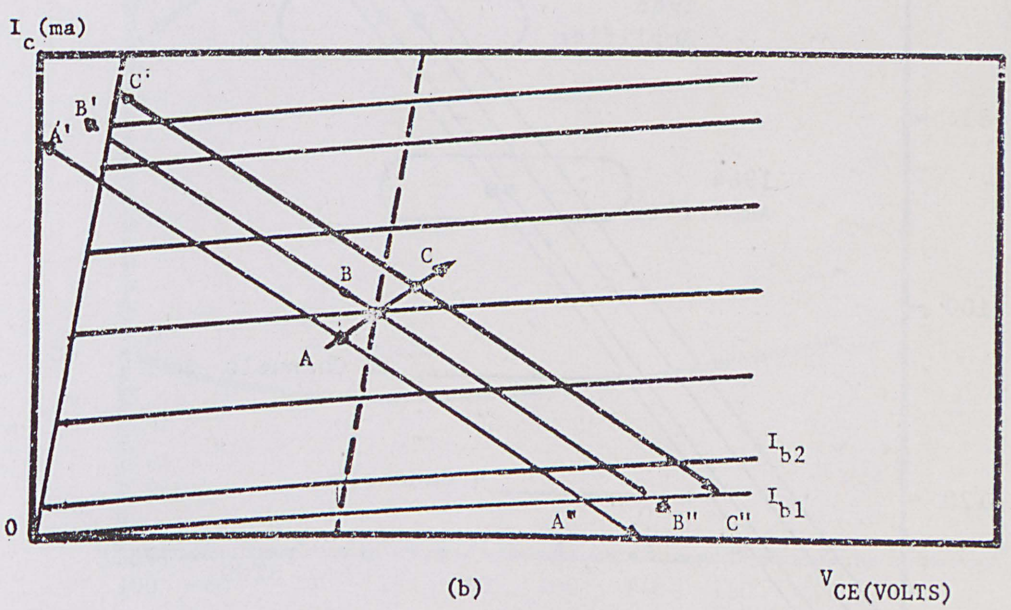


Fig. 3. Nomograph to determine minimum bias point for a given collector load impedance and 12 channel flat operating level.



(a)



(b)

Fig. 4. (a) Over idealized device curves with 150Ω loadline for + 50 dBmV, 12 channel operating.
(b) Idealized curves showing required bias point shift to prevent waveform clipping.

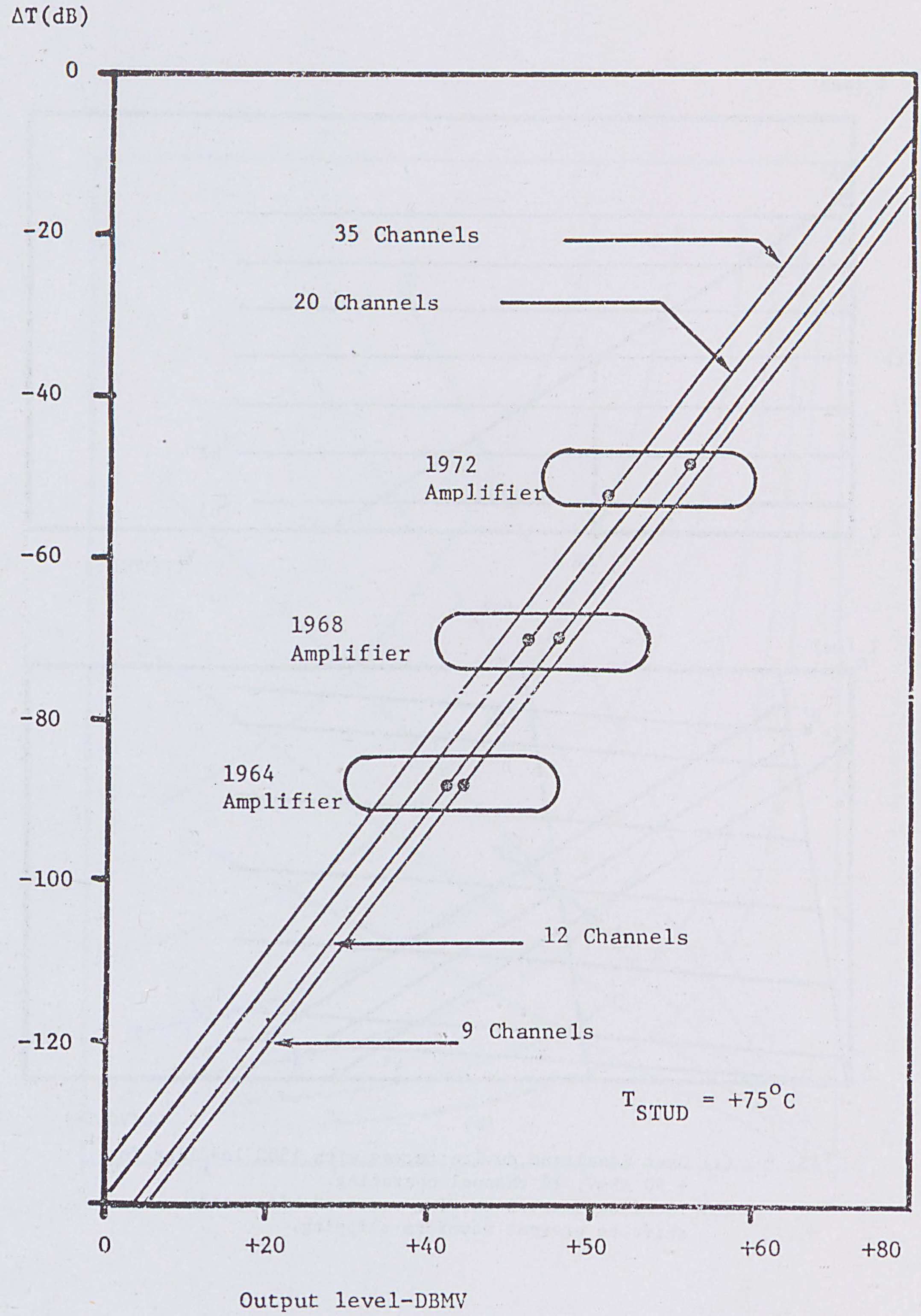


Fig. 5. Thermal modulation of device temperature

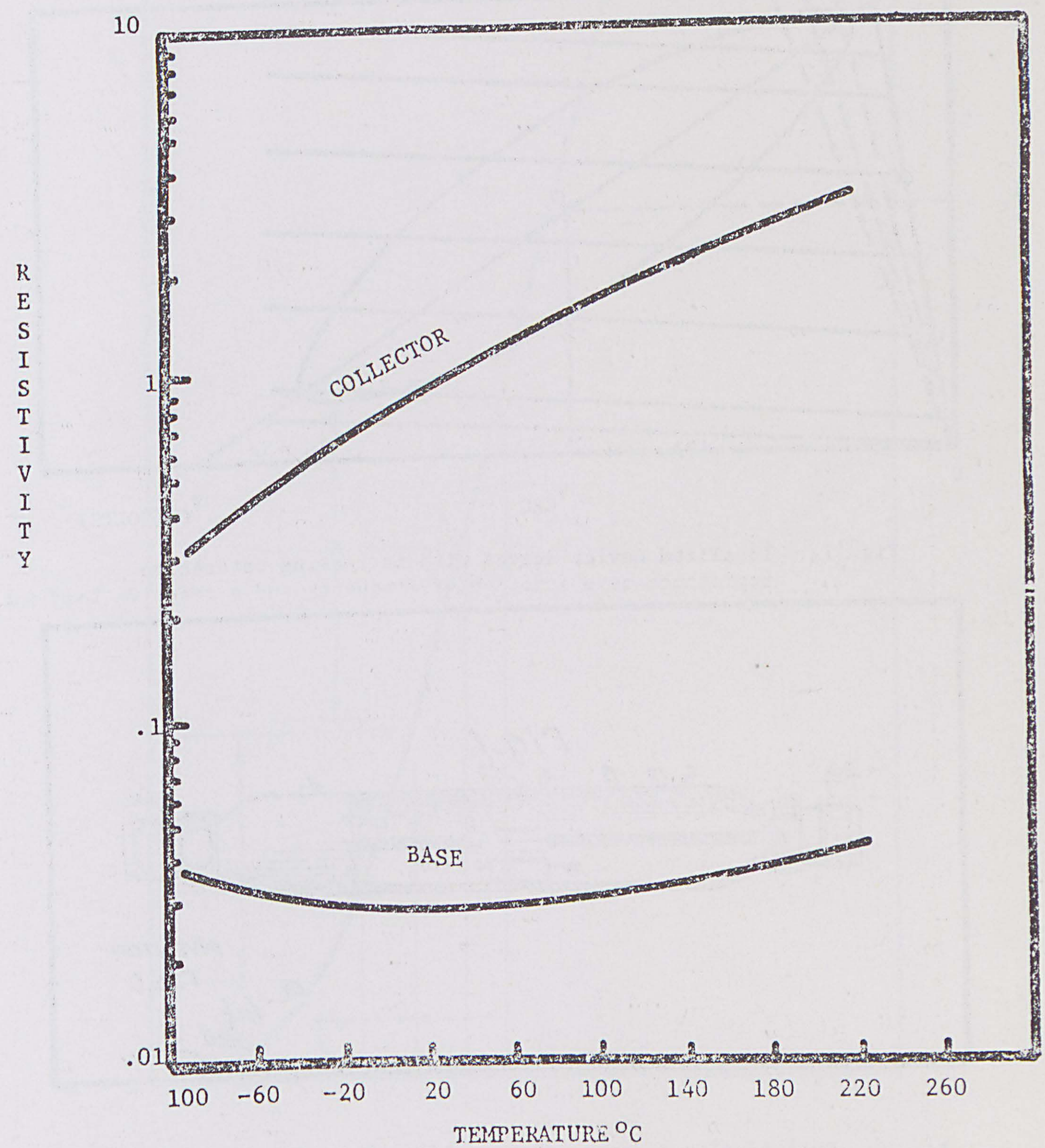


Fig. 6. Resistivity versus temperature for collector and base material.

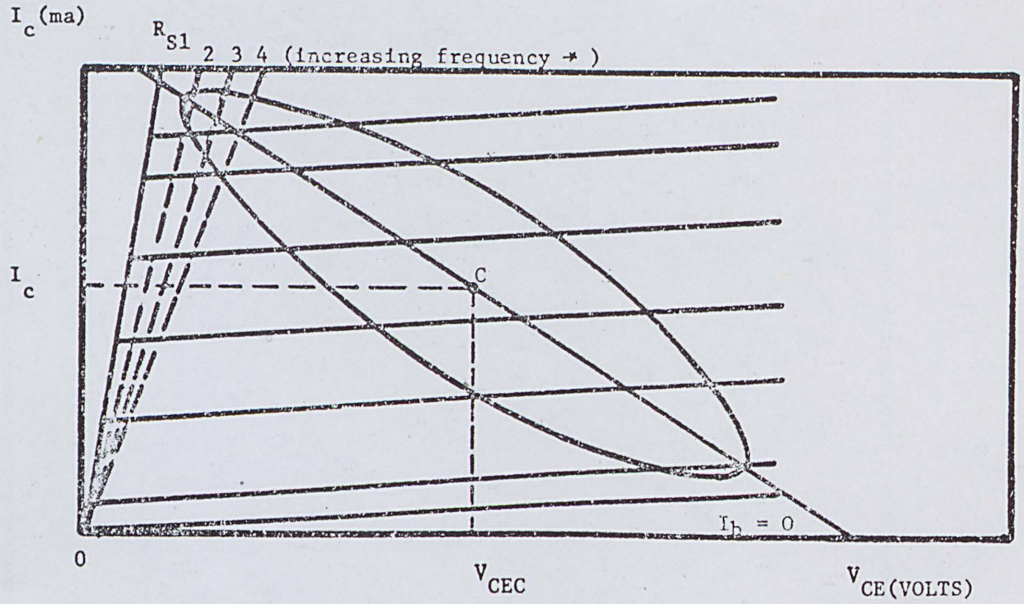


Fig. 7a. Idealized device curves with increasing saturation resistance as a function of frequency and a reactive loadline.

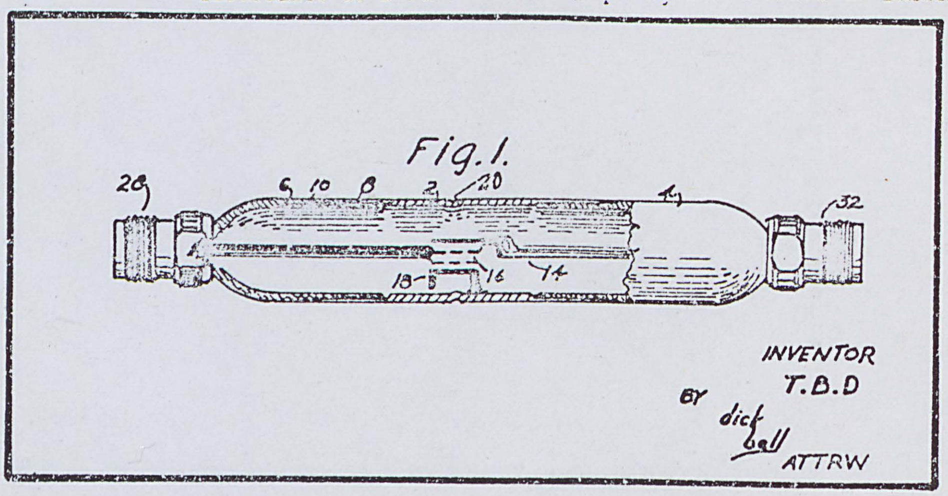


Fig. 7b. Over-idealized device that could produce the curves in Fig. 4a.

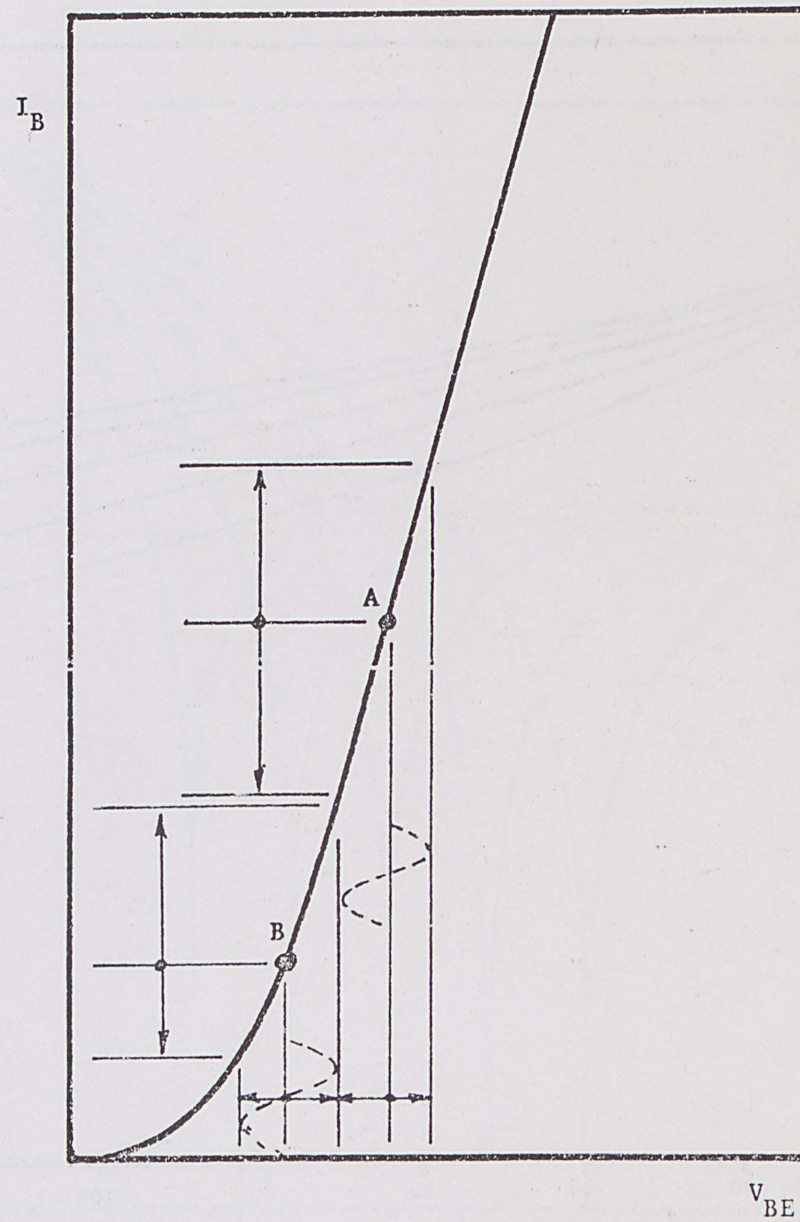


Fig. 8. Base-emitter diode characteristic.

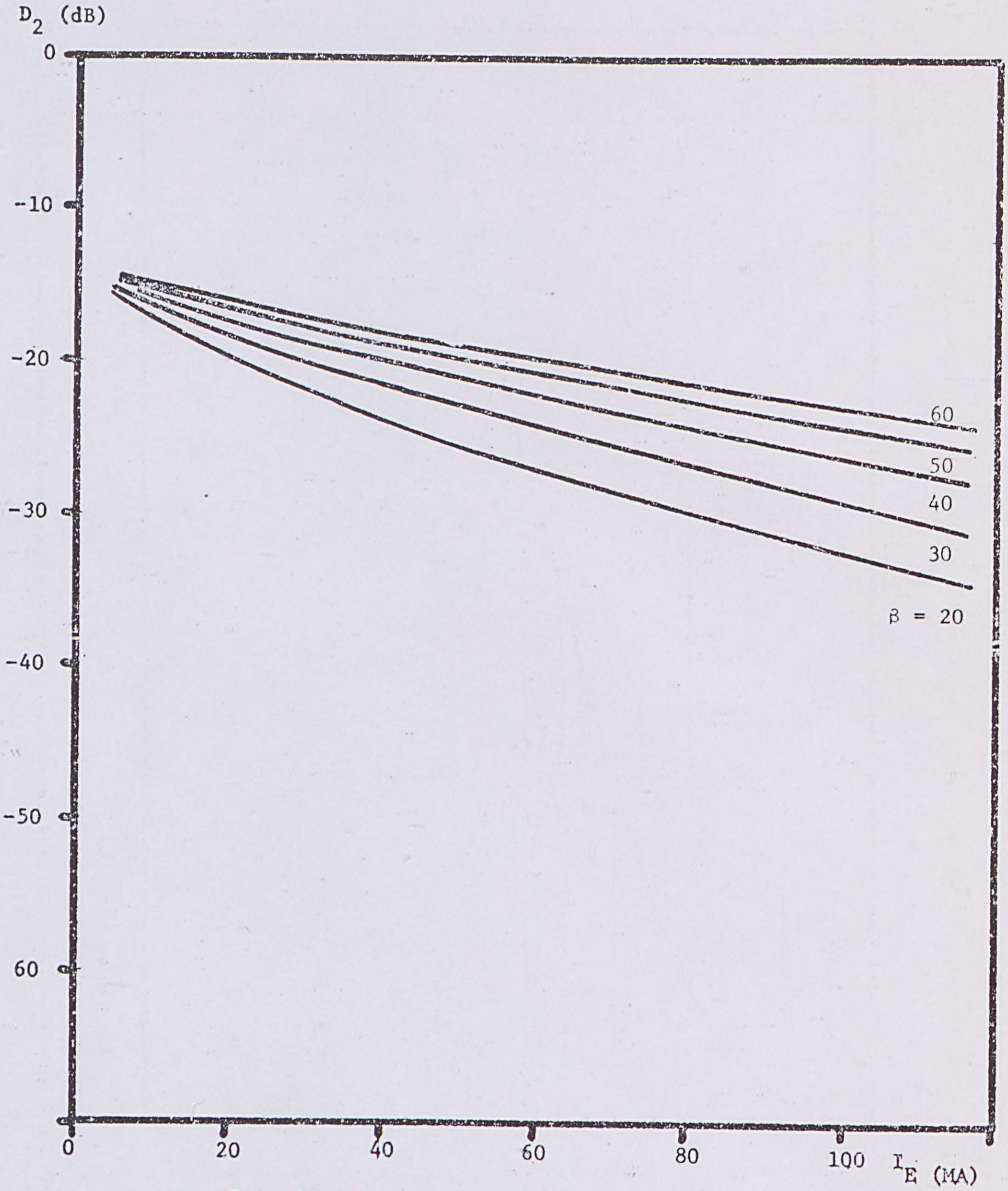


Fig. 9. Second order base-emitter distortion characteristics for Rs = 10Ω

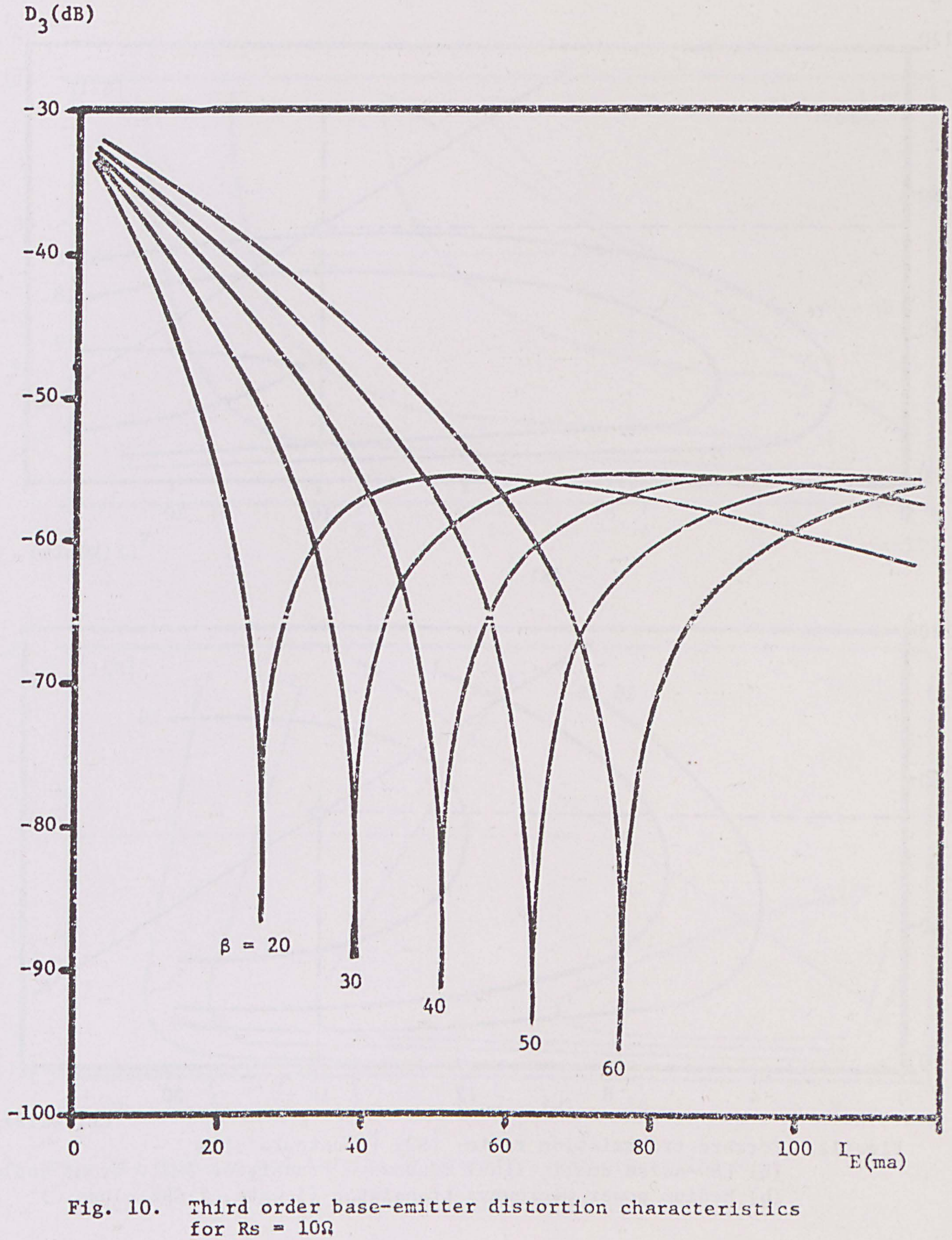


Fig. 10. Third order base-emitter distortion characteristics for Rs = 10Ω

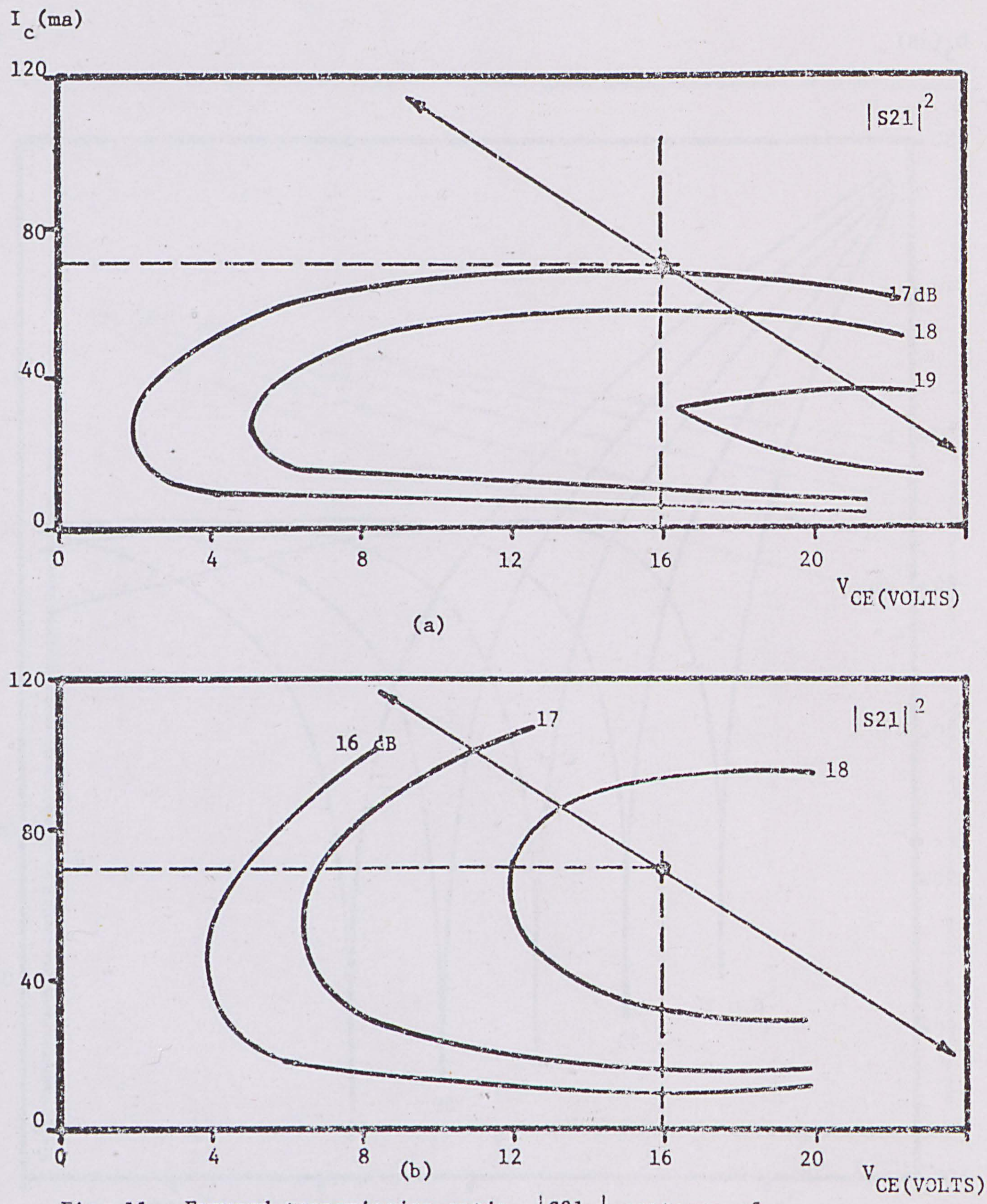


Fig. 11. Forward transmission ratio $|S21|$ contours of a
(a) Low-noise small signal microwave transistor (MATV front end)
(b) Medium power microwave transistor (1 watt, 2 GHz class C)

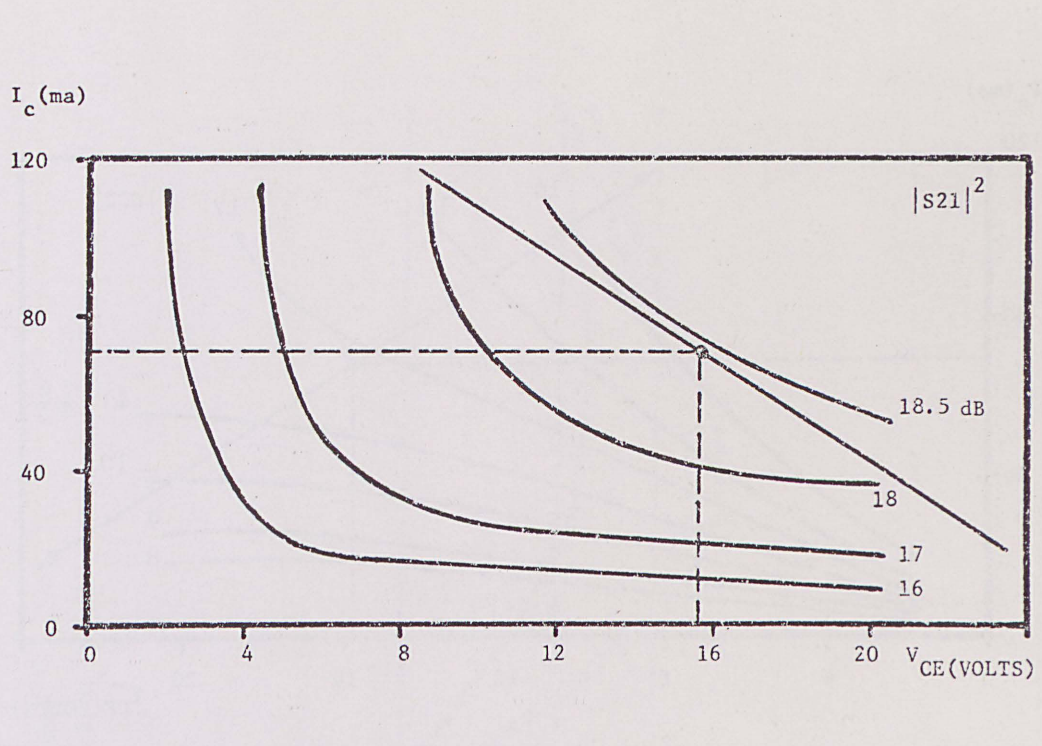
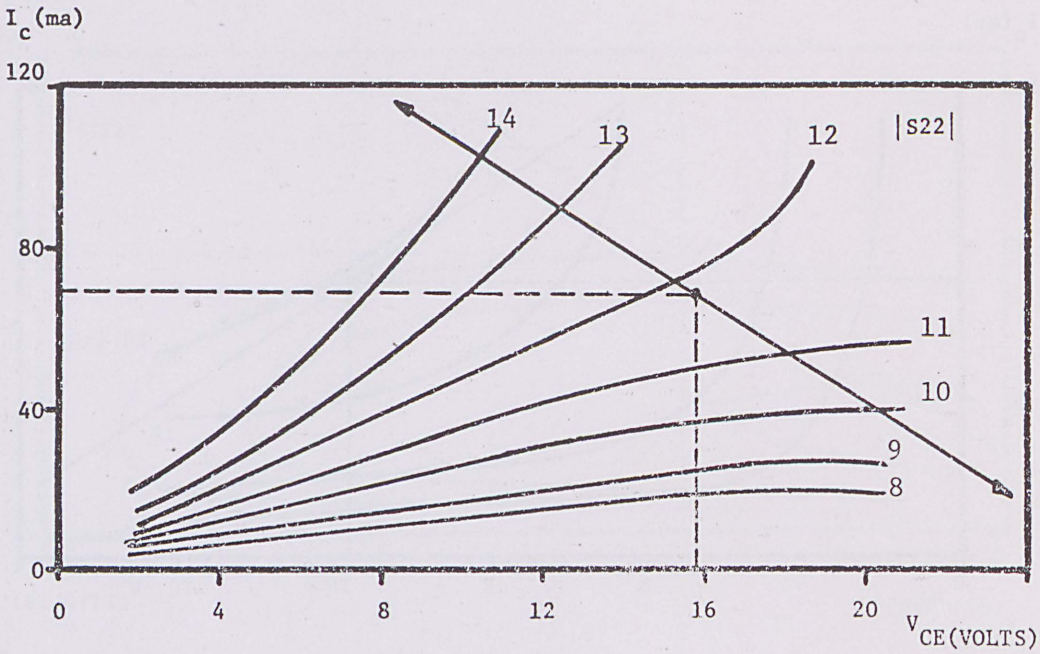


Fig. 12. Forward and reverse S-parameter contours at 250 MHz



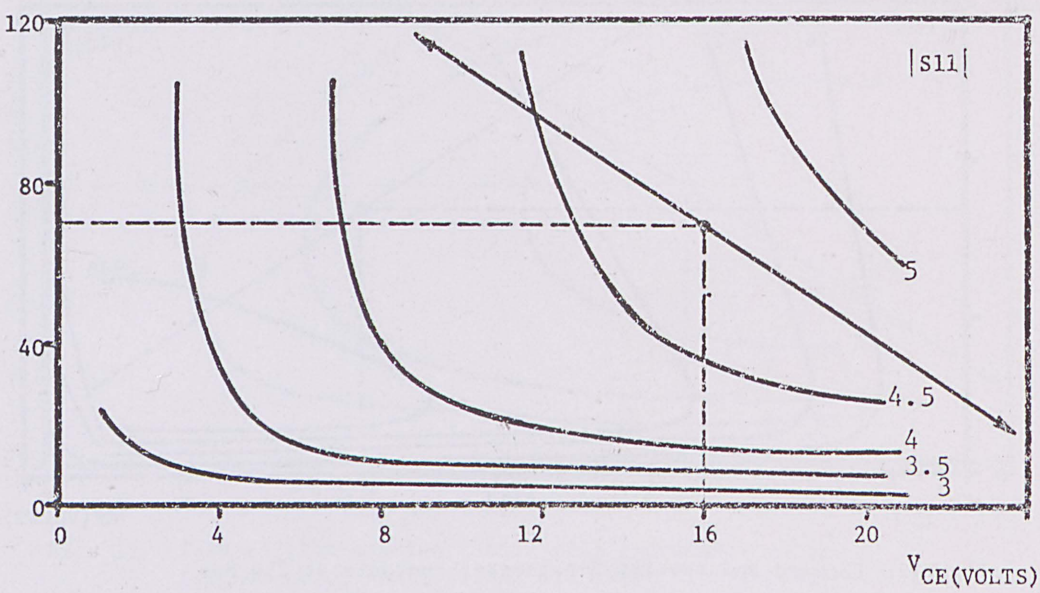
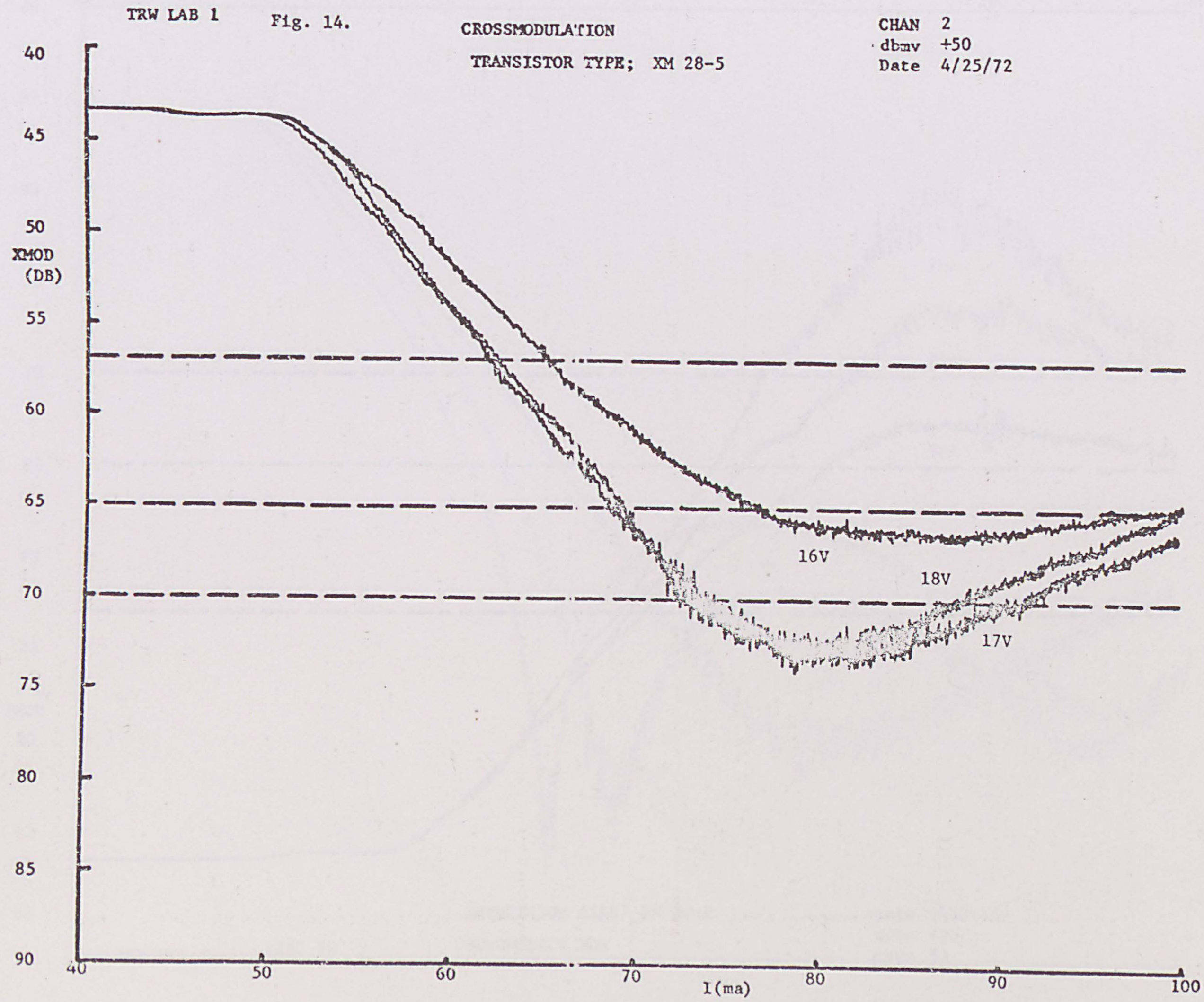
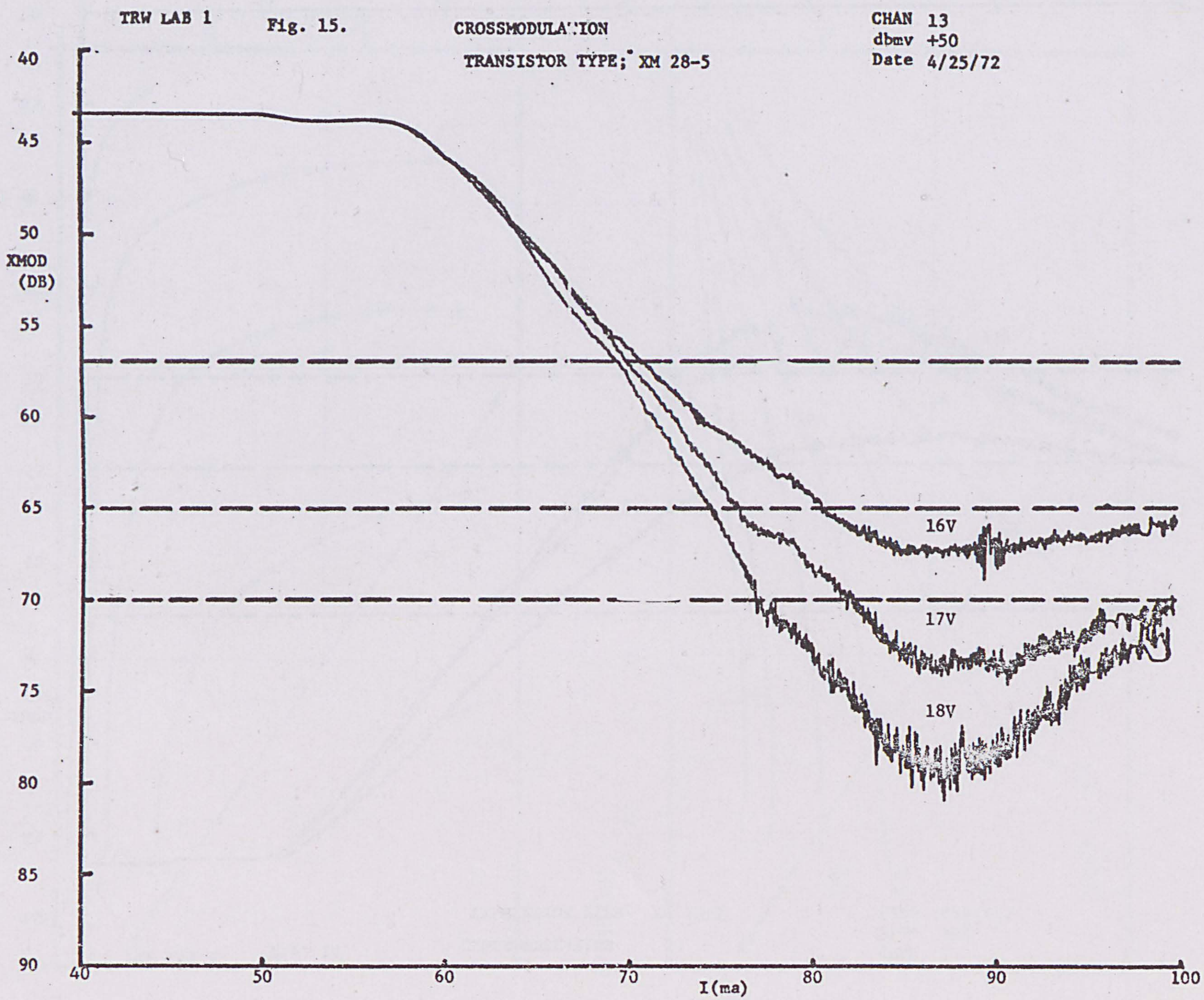
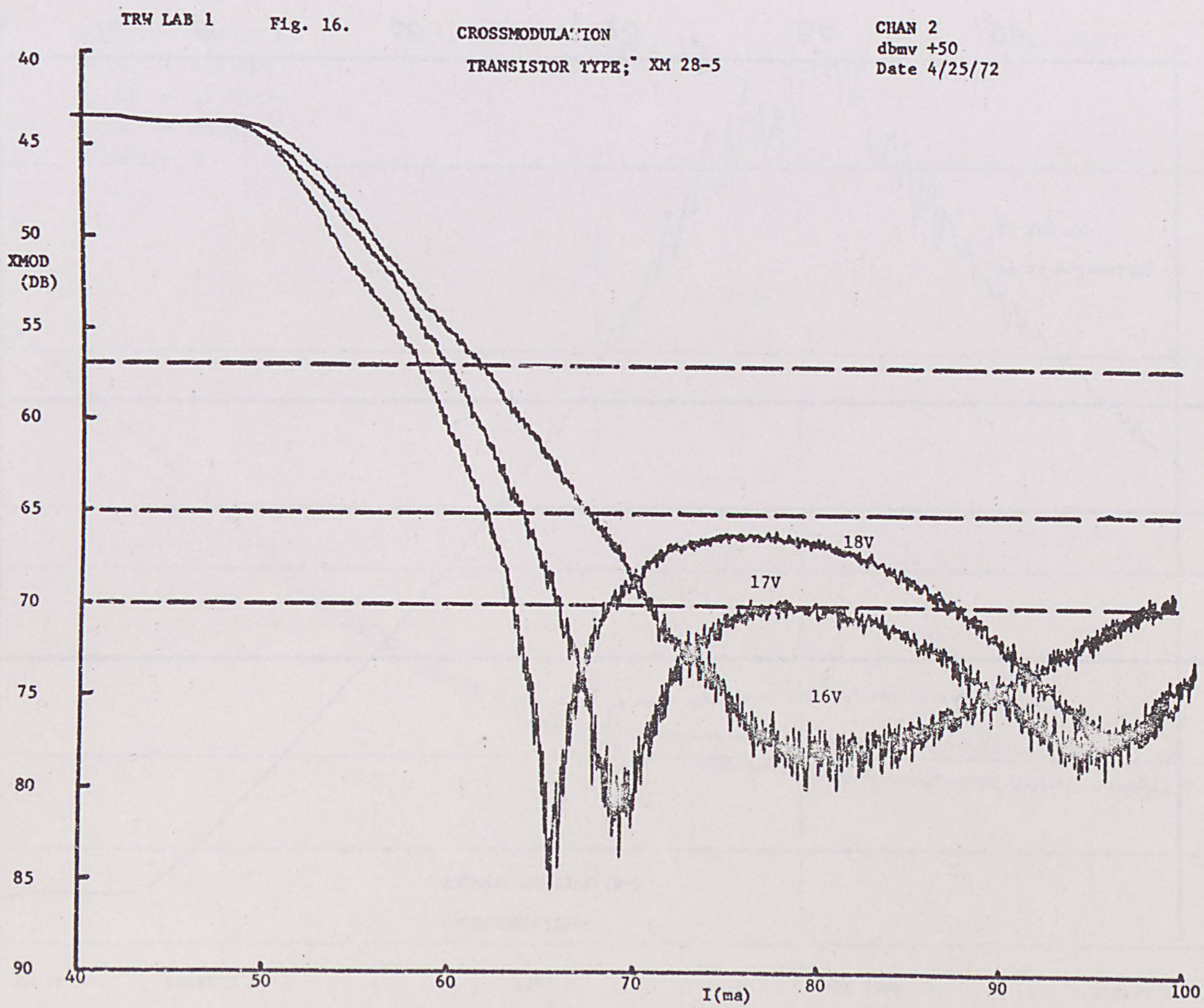


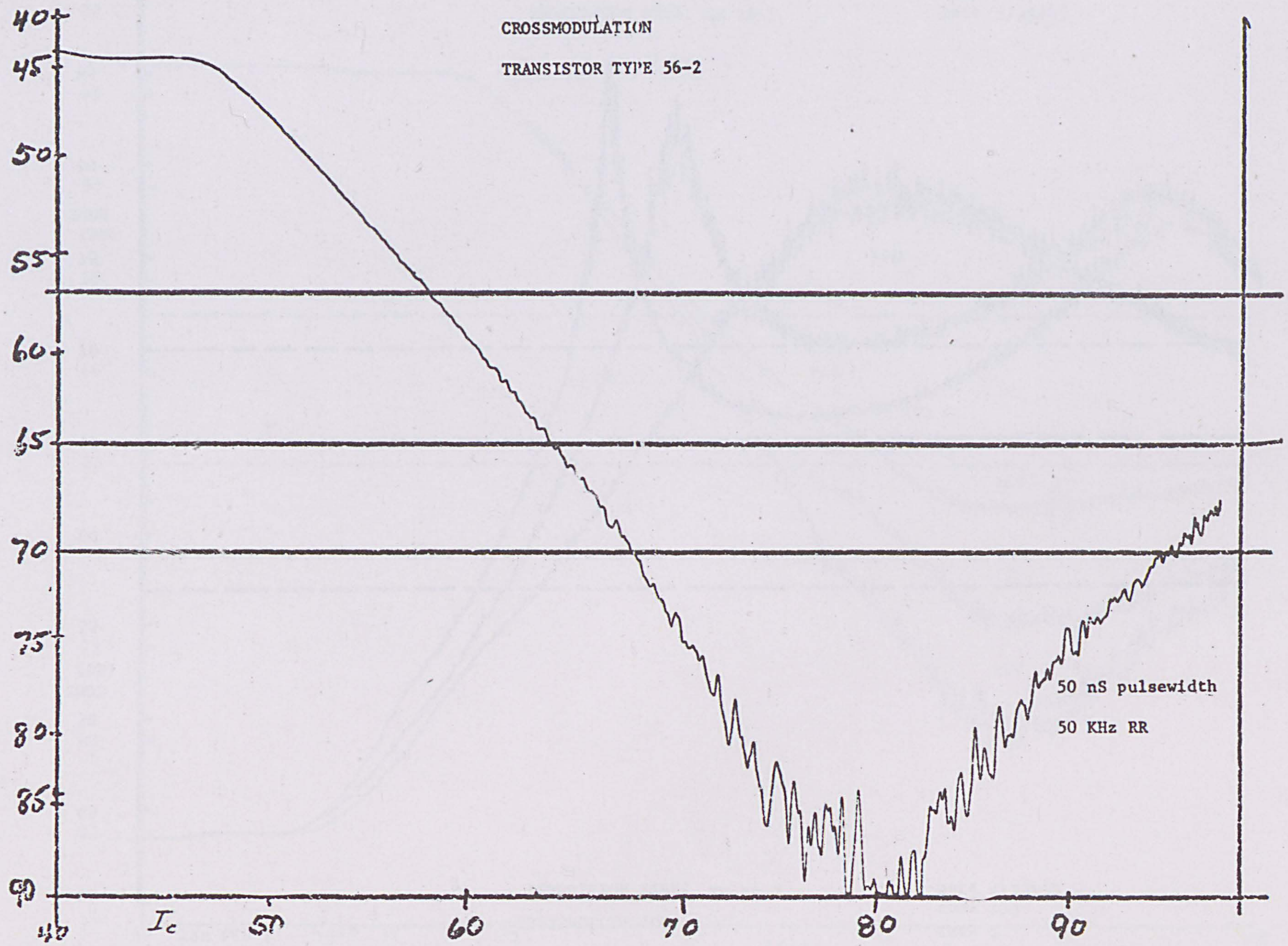
Fig. 13. Output and input S-parameters contours at 250 MHz.

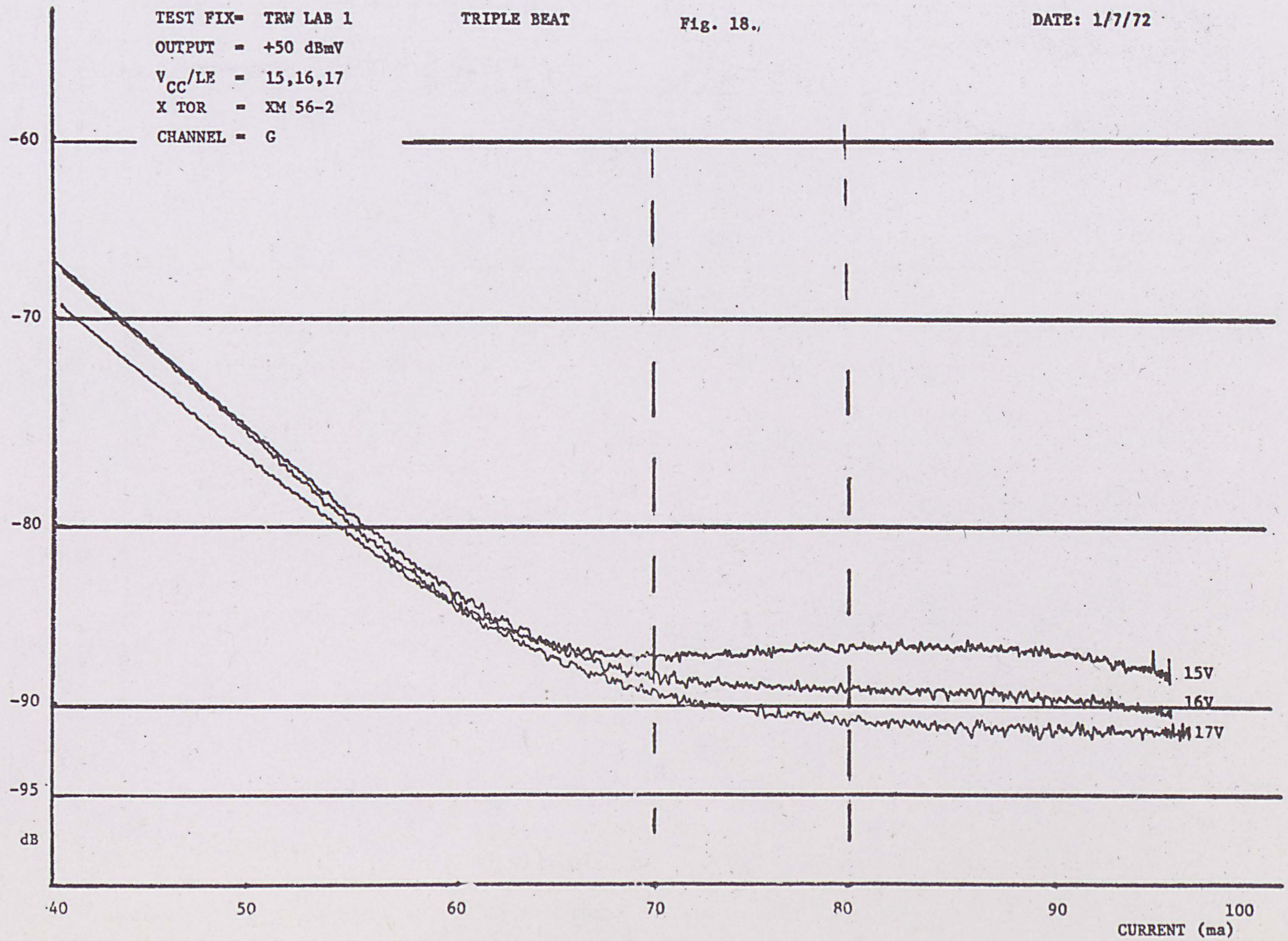

TRW LAB 1
Fig. 14.
CROSSMODULATION
TRANSISTOR TYPE; XM 28-5
CHAN 2
dbmv +50
Date 4/25/72
XMOD
(DB)
40
45
50
55
60
65
70
75
80
85
90
40
50
60
70
80
90
100
I(ma)
16V
18V
17V


TRW LAB 1
Fig. 15.
CROSSMODULATION
TRANSISTOR TYPE; XM 28-5
CHAN 13
dbmv +50
Date 4/25/72
XMOD (DB)
40
45
50
55
60
65
70
75
80
85
90
16V
17V
18V
40
50
60
70
80
90
100
I(ma)


TRW LAB 1
Fig. 16.
CROSSMODULATION
TRANSISTOR TYPE; XM 28-5
CHAN 2
dbmv +50
Date 4/25/72
XMOD (DB)
40
45
50
55
60
65
70
75
80
85
90
40
50
60
70
80
90
100
I(ma)
18V
17V
16V

XM 56-2 CHANNEL 13 Fig. 17. +50 dBmV

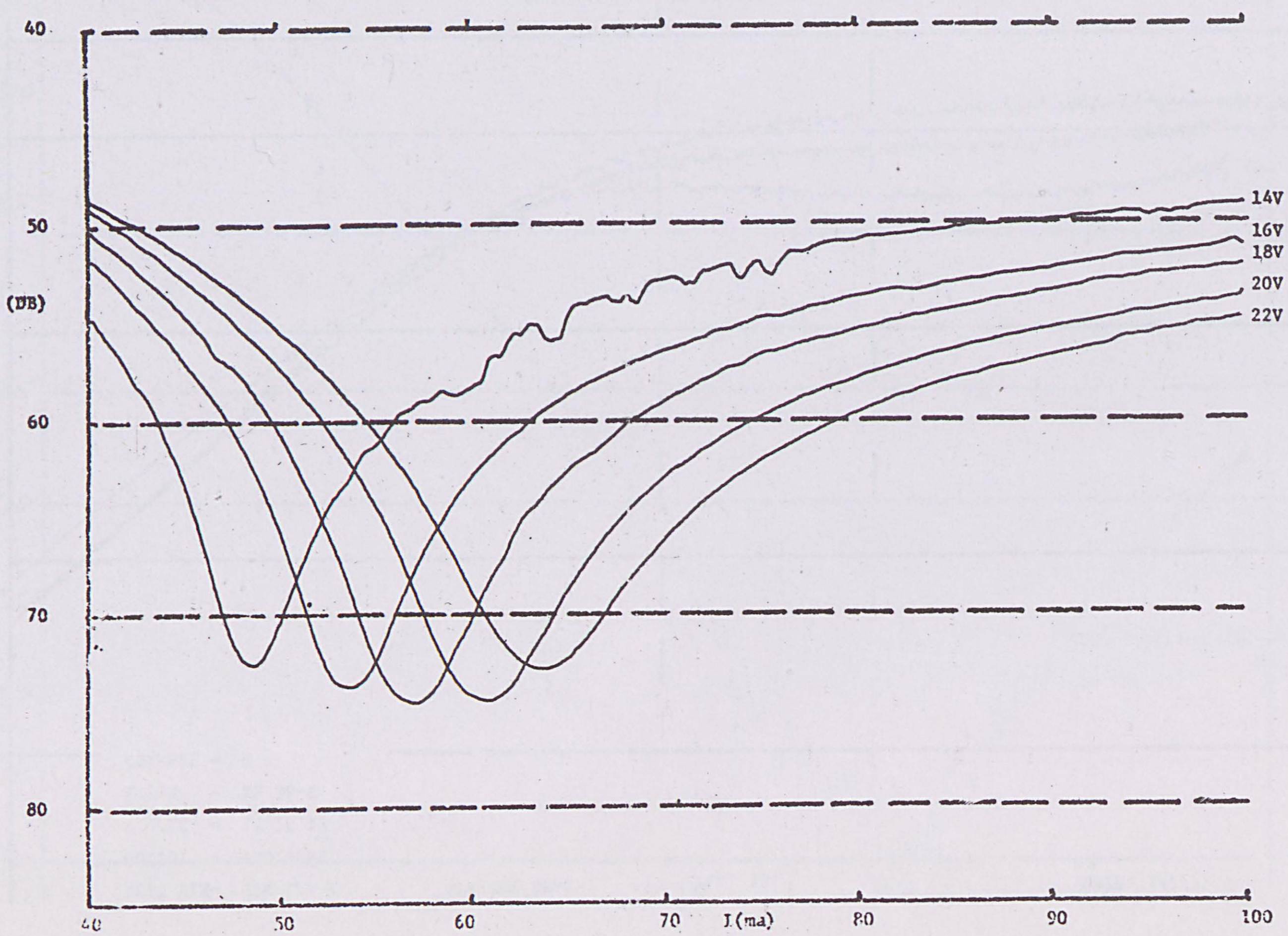



TEST FIX= TRW LAB 1
OUTPUT = +50 dBmV
V_{CC}/LE = 15,16,17
X TOR = XM 56-2
CHANNEL = G
TRIPLE BEAT
Fig. 18.
DATE: 1/7/72
-60
-70
-80
-90
-95
dB
40
50
60
70
80
90
100
CURRENT (ma)
15V
16V
17V

TRW LAB 1 Fig. 19. SECOND ORDER

TRANSISTOR TYPE ; CMP 56-2

CHANNEL 13
dbmv(out) 50
DATE 4-25-72



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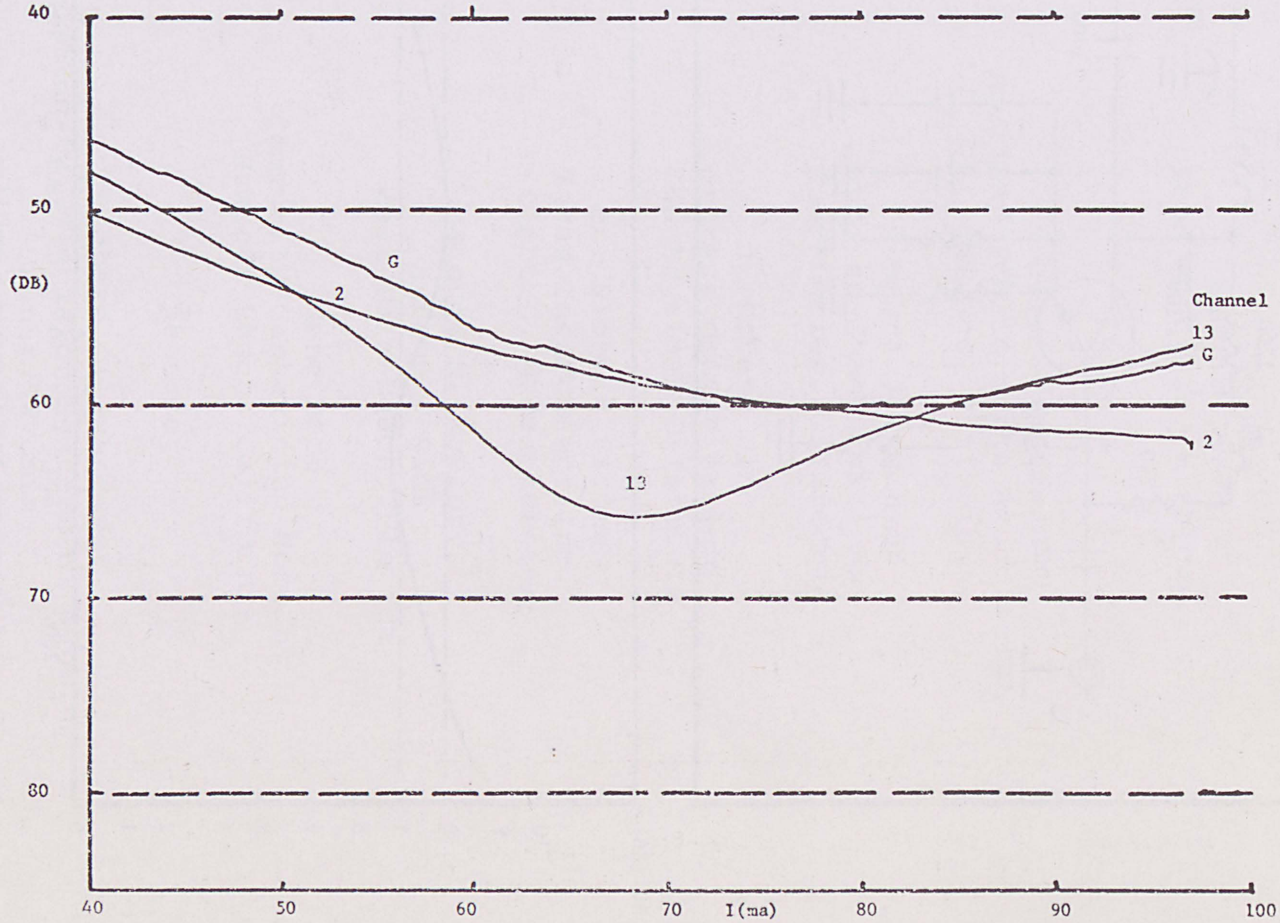
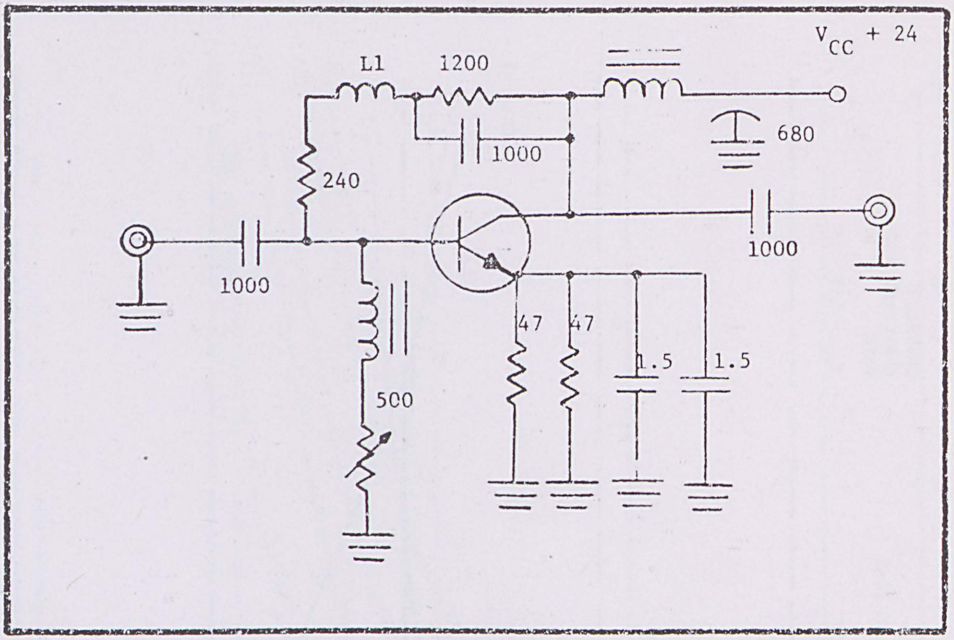
TRW LAB 1 Fig. 20. SECOND ORDER

TRANSISTOR TYPE ; XM 28-5

CHANNEL
dbmv(out) +50
DATE

$V_{CC} = 16V$





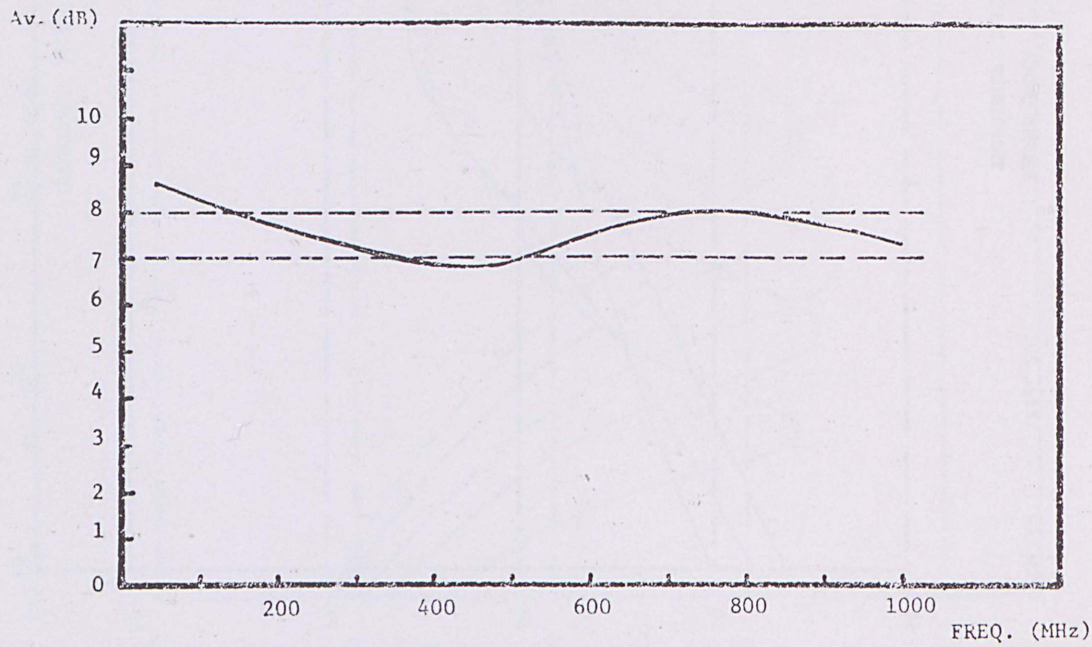


Fig. 21. a) Single-stage transformerless amplifier
b) Voltage gain versus frequency