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SIMULATION PROGRAM OF NONLINEARITIES APPLIED  
TO TELECOMMUNICATION SYSTEMS

Christian Thomas

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SIMULATION PROGRAM OF NONLINEARITIES APPLIED  
TO TELECOMMUNICATION SYSTEMS

Christian Thomas

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1. INTRODUCTION

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The performance prediction for multicarrier operation of a nonlinear satellite repeater is essential for the optimization of a telecommunications system which operates in the multiple access mode with frequency distribution.

The spectra of the signals running through the equipment undergo deformations due to intermodulation in the progressive wave amplifiers (TOP), low noise amplifiers, mixers or filters. Depending on their mode or operation point, this equipment behaves more or less like nonlinear elements. This produces amplitude distortions and phase distortions which are detrimental to the signal being transmitted. Basically, this equipment has two kinds of nonlinearities [1]:

- a phase nonlinearity;
- an amplitude nonlinearity.

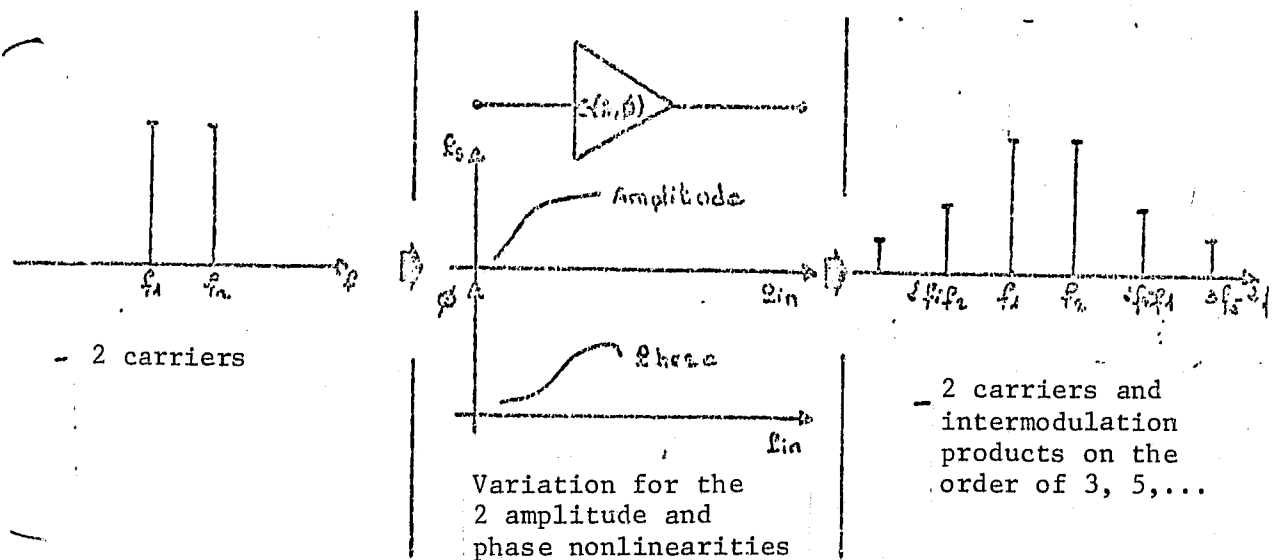
In most cases these nonlinearities have no memory, can be modeled, and can be measured:

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\*Numbers in the margin indicate pagination of original foreign text.

- They have no memory because the output voltage of the element only depends on its direct gain and the input voltage: this is the case of the TOP and the filters;
- They can be modeled because they can be analytically simulated by complex polynomials;
- They can be measured using a network analyzer in the case where the number of carriers is not too large.

Schematically, in the case of an input signal consisting of two carriers of equal amplitude and separated by  $\Delta F$ , the problem can be formulated in the following way:



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In the present report we will describe a "source program" which allows one to calculate the following from subprograms of the COMLIB [2]:

- the complex envelope of the output signal;
- its spectrum;
- the carrier ratios with intermodulation products;
- the carrier ratios with intermodulation noise in a frequency band (C/N).

The input signal can consist of as many lines as necessary.

The results of simulation will be compared to those obtained in experiments. A direct application consists of a low-noise amplifier which operates in the frequency range between 3.7 to 4.2 GHz [3], which is designed for a 6-4 GHz repeater. The input signals consist of two or three carriers and of a noise density associated with a single carrier of low amplitude. The simulation program will also be applied to the case of an input amplifier for a direct TV satellite, for which the transmission/reception coupling introduces parasitic signals at the input which intermodulate with the useful signals. Such a program, therefore, is a working tool which is useful for "conceiving systems" and is also useful to the "circuit engineer" because, from overall systems specifications (for example, minimum C/I), it can help in the selection of the component or the equipment and can also specify its optimum operational point.

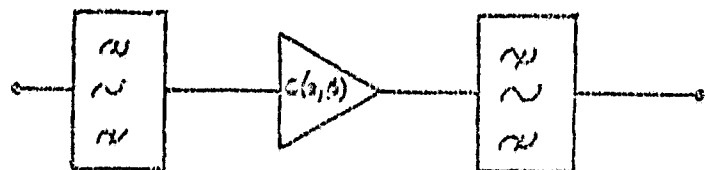
## 2. STARTING HYPOTHESES REGARDING THE 3.7 - 4.2 GHz AMPLIFIER

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The nonlinear model of the amplifier considered consists only of amplitude and phase nonlinearities. In particular, we will ignore the amplitude/frequency nonlinearities which would lead to a numerical resolution system which would be too complicated.

The amplifier is considered as having no memory. That is to say, there are no reactions at the output to the input. In practice, there is a complex inverse gain  $S_{12}$ , but in the first approximation we can assume that it is sufficiently small compared with the direct gain  $S_{21}$ , so that it can be ignored. In addition, the amplifier is assumed to have a limited band so as to neglect the influence of the input signal harmonics.

The following is the equivalent diagram for the amplifier:



passband fil-    amplitude and    passband fil-  
 ter                    phase non-            ter  
                           linearities

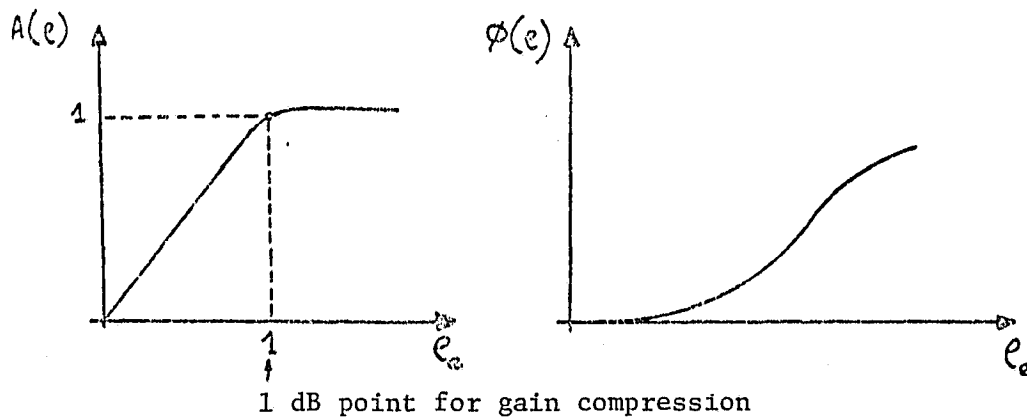
### 3. THEORETICAL ANALYSIS

For more clarity, we seem to prefer to collect several terms about complex signals and complex gains developed by M. Sombrin in Report No. 77/026/CT/TCS/LT [1].

#### 3.1 Concept of Complex Gain

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The amplitude and phase nonlinearities have the following appearance:



where  $e_e$  : input signal amplitude

$A(e_e)$  : output signal amplitude

$\phi(e_e)$  : phase difference between output and input

We can then introduce a complex gain of the amplifier which takes into account these nonlinearities:

$$G(e_\omega) = \frac{A(e_\omega)}{e_\omega} \cdot e^{i\psi(e_\omega)}$$

In the case of a low noise amplifier, we arbitrarily normalize the complex gain to 1 for the 1 dB compression point:

$$e_c = 1 \Rightarrow A(e_c) = 1 \Rightarrow |G(e_c)| = 1$$

This normalization allows one to generalize the method and to apply any kind of nonlinear element: mixer, limiter, assuming that the coordinates of the 1 dB compression point are known.

### 3.2. Representation of Signals as a Function of Time

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The numerical representation of a signal as a function of time requires the sampling of this signal. The number of samplings will become prohibitive at several gigahertz and because of the theorem of Shannon.

It can be shown [1] that the signal can be represented only in the form of its complex envelope, that is if  $S(t)$  is a real signal which is modulated in amplitude and phase:

$$\begin{aligned} S(t) &= E(t) \cdot \cos[\omega_0 t + \phi(t)] \\ S(t) &= \mathcal{R} [ E(t) \cdot e^{i\omega_0 t} \cdot e^{i\phi(t)} ] \end{aligned}$$

The complex envelope  $E_c(t)$  around  $\omega_0$  is:

$$E_c(t) = E(t) \cdot e^{i\phi(t)} = X(t) + i Y(t)$$

We then have:

$$S(t) = \mathcal{R} [ E_c(t) \cdot e^{i\omega_0 t} ] = \mathcal{R} [ (X(t) + i Y(t)) \cdot e^{i\omega_0 t} ]$$



If B is the passband of the signal, the maximum frequency of  $E_c(t)$  is then  $B/2$ . The sampling will be done at the frequency B instead of at  $f_0$ . The calculation time is then reduced accordingly.

### 3.3. More General Case

If we apply the nonlinearity over the complex envelope of the signal, we then find:

- real signal:  $s(t) = E(t) \cos [\omega_0 t + \varphi(t)]$
- complex signal around  $\omega_0$ :  $E_c(t) = E(t) \cdot e^{j\varphi(t)}$
- complex gain of the nonlinearity:  $G(p_e) = B(p_e) \cdot e^{j\psi(p_e)}$
- complex envelope at output:  $S_c(t) = E(t) \cdot e^{j\varphi(t)} \cdot G[E(t)]$   
 $= E(t) \cdot B[E(t)] \cdot e^{j[\varphi(t) + \psi(E(t))]}$
- real output signal:  
 $s(t) = \Re \left[ S_c(t) \cdot e^{j\omega_0 t} \right] = E(t) \cdot B[E(t)] \cdot \cos [\omega_0 t + \varphi(t) + \psi(E(t))]$

In practice, we have access to  $B[E(t)]$  and  $\psi(E(t))$  from measurements.

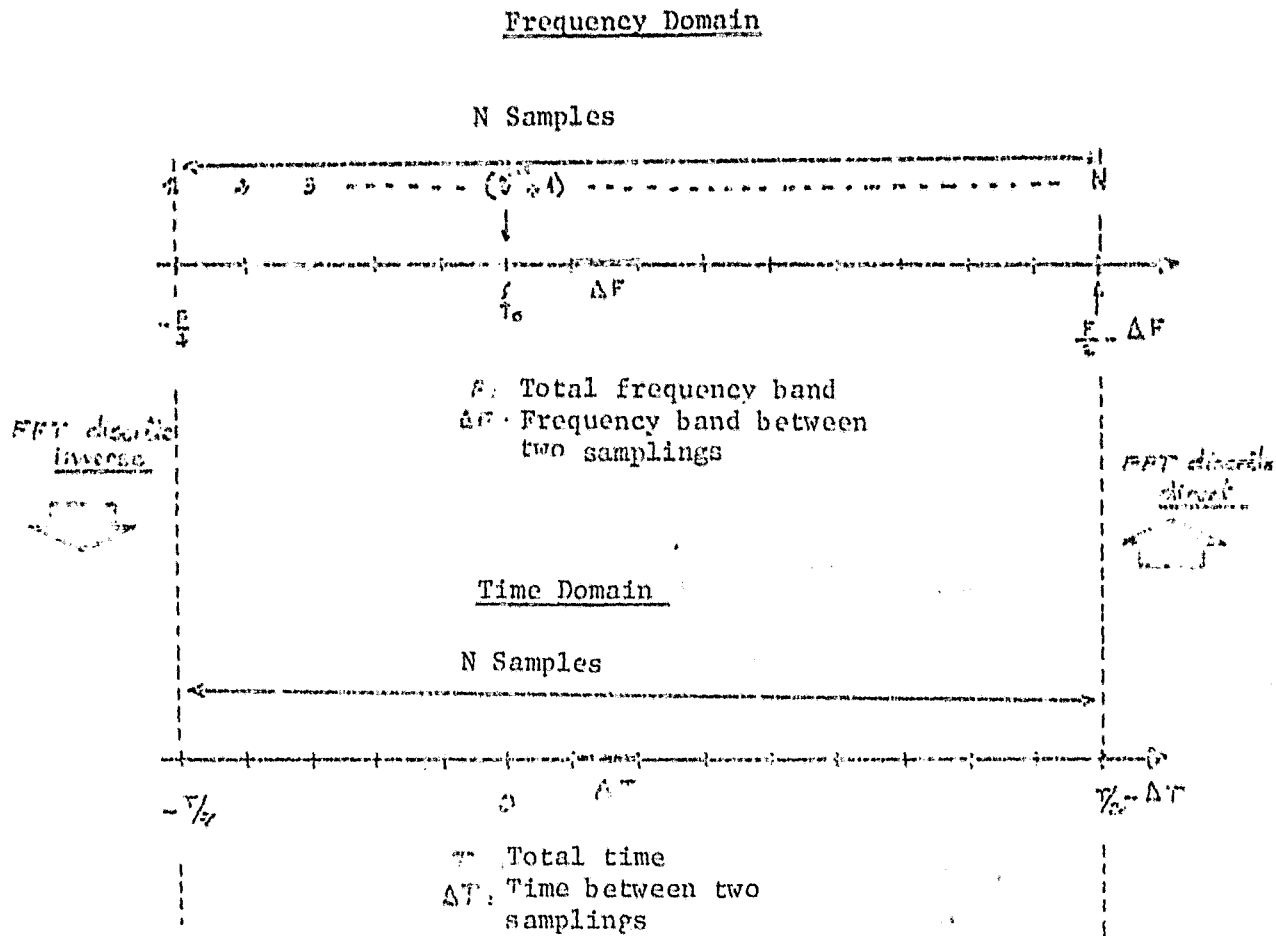
### 3.4. Properties of Fast Fourier Transform

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The transfer from the complex envelope of the signal to its spectrum and inverse operation implies the use of a Fourier transform. In order to reduce the calculation time and increase the accuracy of the results, at the present time numerical calculus methods use the Fast Fourier Transform method (FFT).

### 3.4.1. Time-Frequency Correspondence for FFT

The two diagrams given below give an example of using discrete FFT for calculating discrete spectra and the complex envelope of the signal:



These time-frequency relationships, therefore, satisfy the following relationships:

$$\Delta T = \frac{1}{F}$$

$$\Delta F = \frac{1}{T}$$

$$\Delta F \Delta T = \frac{1}{N}$$

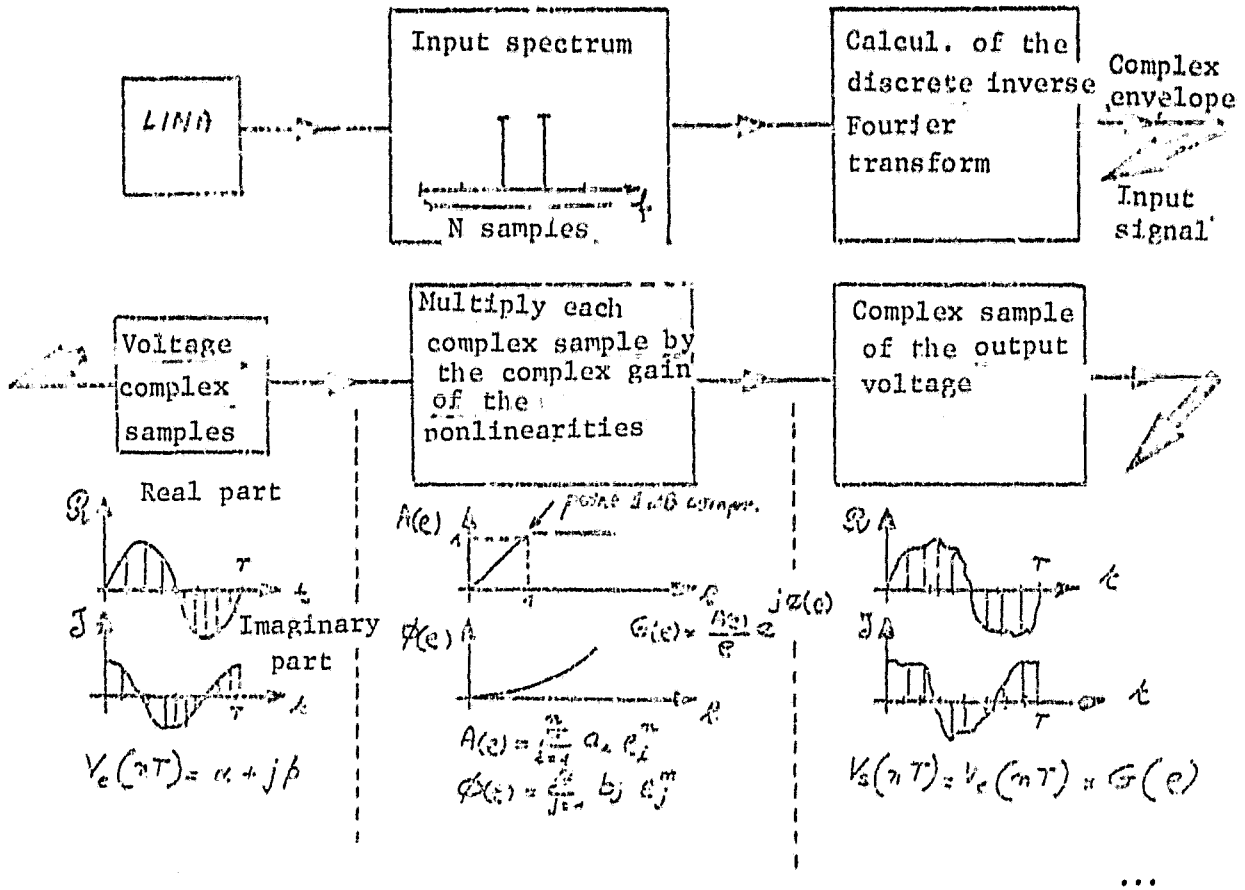
$$F \cdot T = N$$

$$\Delta T = \frac{T}{N}$$

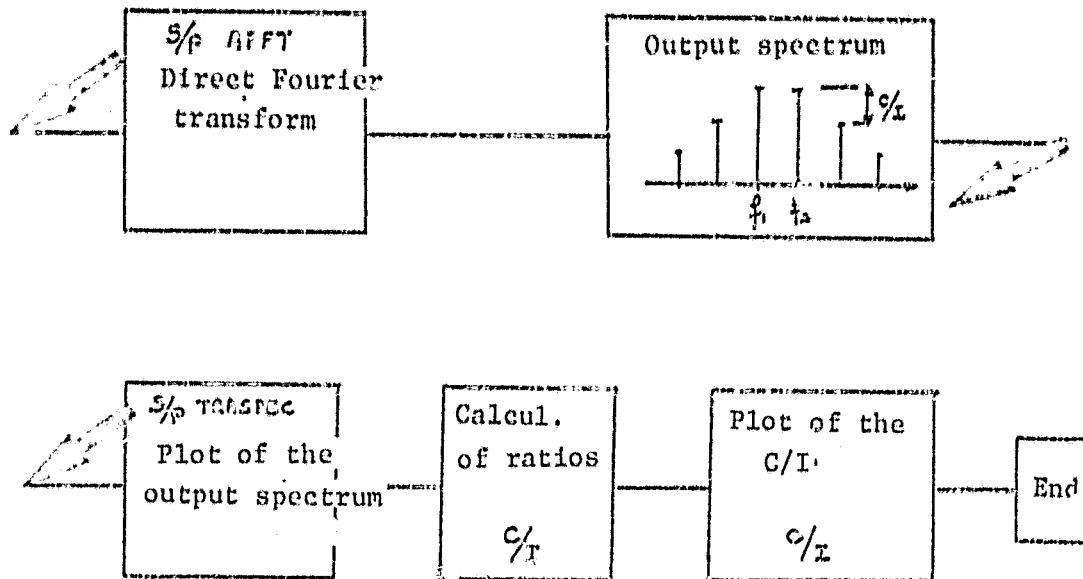
#### 4. NUMERICAL SOLUTION PROCESS

##### 4.1. Synopsis

Below we will give an organization diagram of the subprograms which are called by the main program LINA and their mathematical significance.



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#### 4.2 Description of the Main Program LINA

LINA carries out the operations described above if the user has supplied a certain number of data about the following:

- the number of samples:  $N$ ;
- the input spectrum (phase amplitude and frequency of each line);
- the polynomials which model the nonlinearities;
- the coordinates of the 1 dB compression point;
- the input power range;
- at the output, the various plotting options desired (either as listings or as Benson prints).

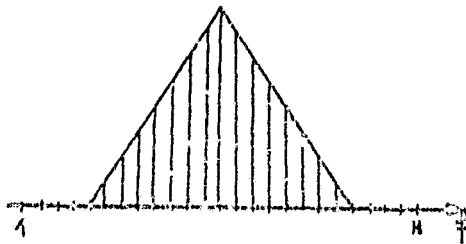
Appendix 1 gives the details about the data cards and their input format, /11 as well as the FORTRAN of the subprograms which do not belong either to COMLIB or to CNESLIB.

### 4.3 Precautions for Use

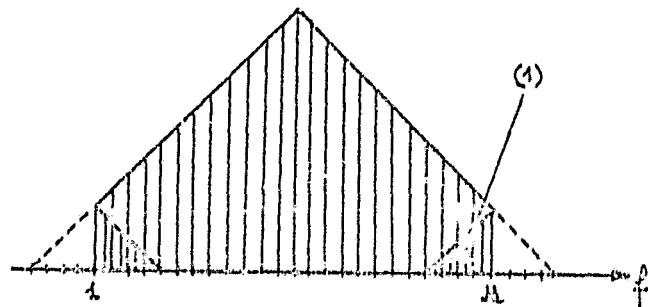
A certain number of precautions must be maintained in order to obtain correct results. First of all, the number of samplings must be sufficient so as to contain the lines of the inlet spectrum and the lines which are suspected in the output spectrum.

An excessively small number of samplings can lead to a folding phenomena of the spectrum which is related to the periodicity of the discrete Fourier transform. This phenomenon is illustrated by the following figures:

Normal spectrum,  
N is sufficient



Incorrect output spectrum,  
M is insufficient



In [1], the folded-back zone of the spectrum is added to the real spectrum. At the end of the spectrum there is then a recombination in amplitude and phase of these lines, which can lead to substantial errors.

In practice, the number of samples must cover twice the passband of the input signal. In addition, the number of samples must be a power of 2:

$$N = 2^m \text{ or } 4096$$

5. RESULTS

5.1. Case Where the Input Signal Consists of 2 Carriers of Equal Amplitude

5.1.1 Experiments

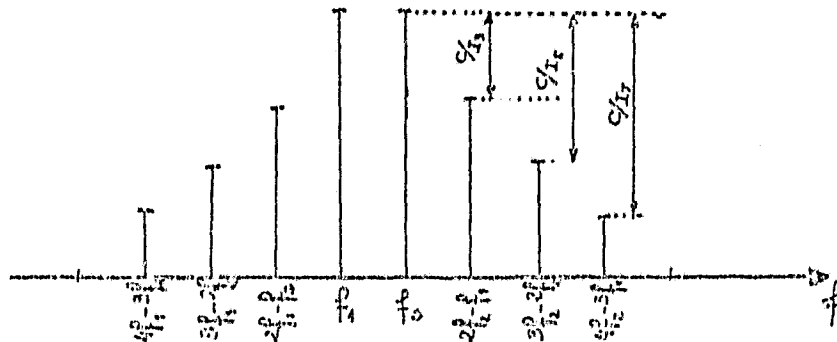
Two carriers spaced by 5 MHz around 4 GHz are reentered in the end stage of an amplifier with an intermediate frequency, operating in the frequency band between 3.7 to 4.2 GHz.

The intermodulation products created by the phase and amplitude nonlinearities have the following form:

$$f_i = \pm m f_1 \pm n f_2$$

- where:  $f_1$  and  $f_2$ : the two carriers
- $m = 0, 1, 2, \dots$  order of intermodulation products
- $n = 0, 1, 2, \dots$  " " "

If we ignore products with an order greater than 7, we obtain an output spectrum having the following form:



The test equipment used is the one described in Figure 1.

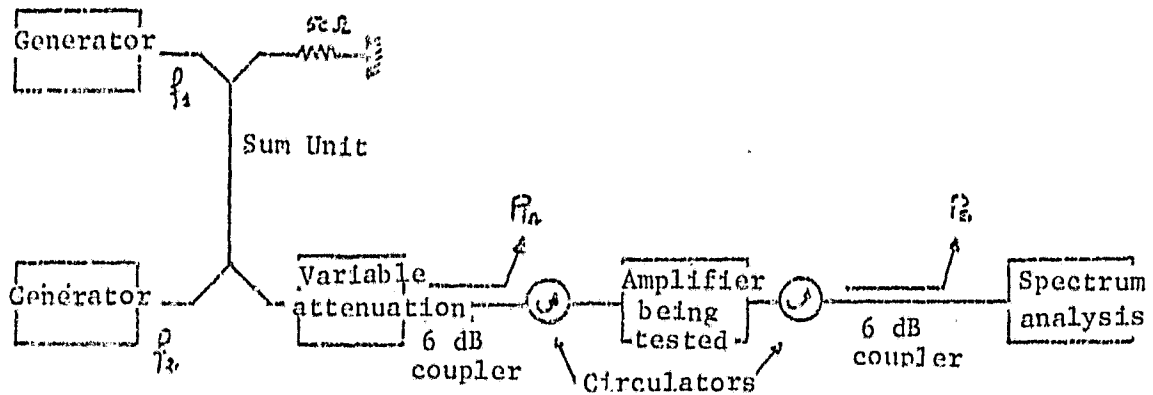


Figure 1

The ratios  $C/I_3$  are measured for different values of total input power and are plotted in the curves of Figure 2.

The test complex allows one to determine the amplitude-phase relationship of the amplifier, and was previously described in Report No. 76/061/DSES/RT/TTI/MO/E2. The amplitude and phase curves for this amplifier are shown in Figure 3.

### 5.1.2 Simulation

We select 256 samples for representing the input signal. The output spectrum is given in Figure 4 for a total input power of - 20 dBm.

The intermodulation products calculated according to the previously described method were plotted in Figure 2 and compared with the measured values.

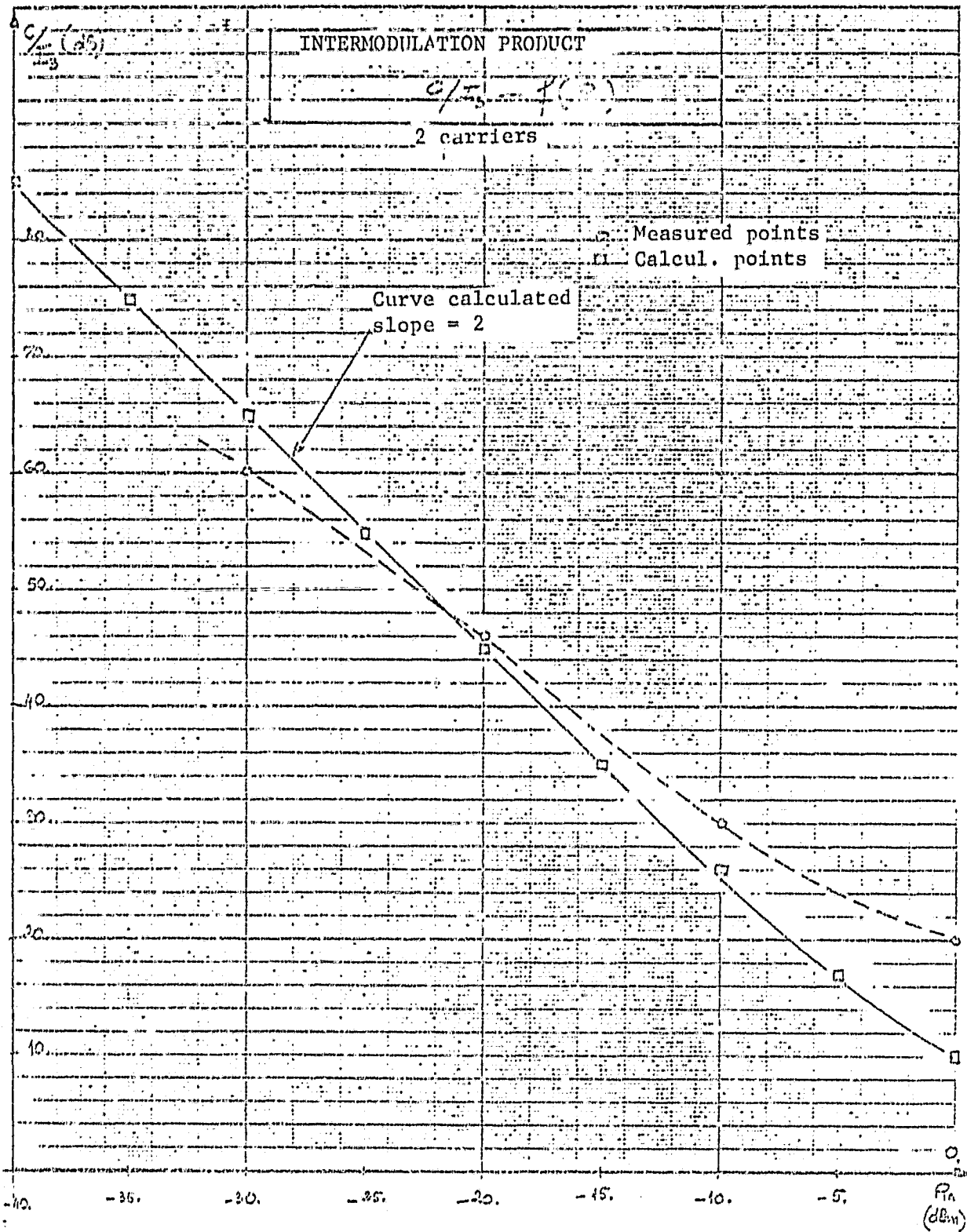


Figure 2



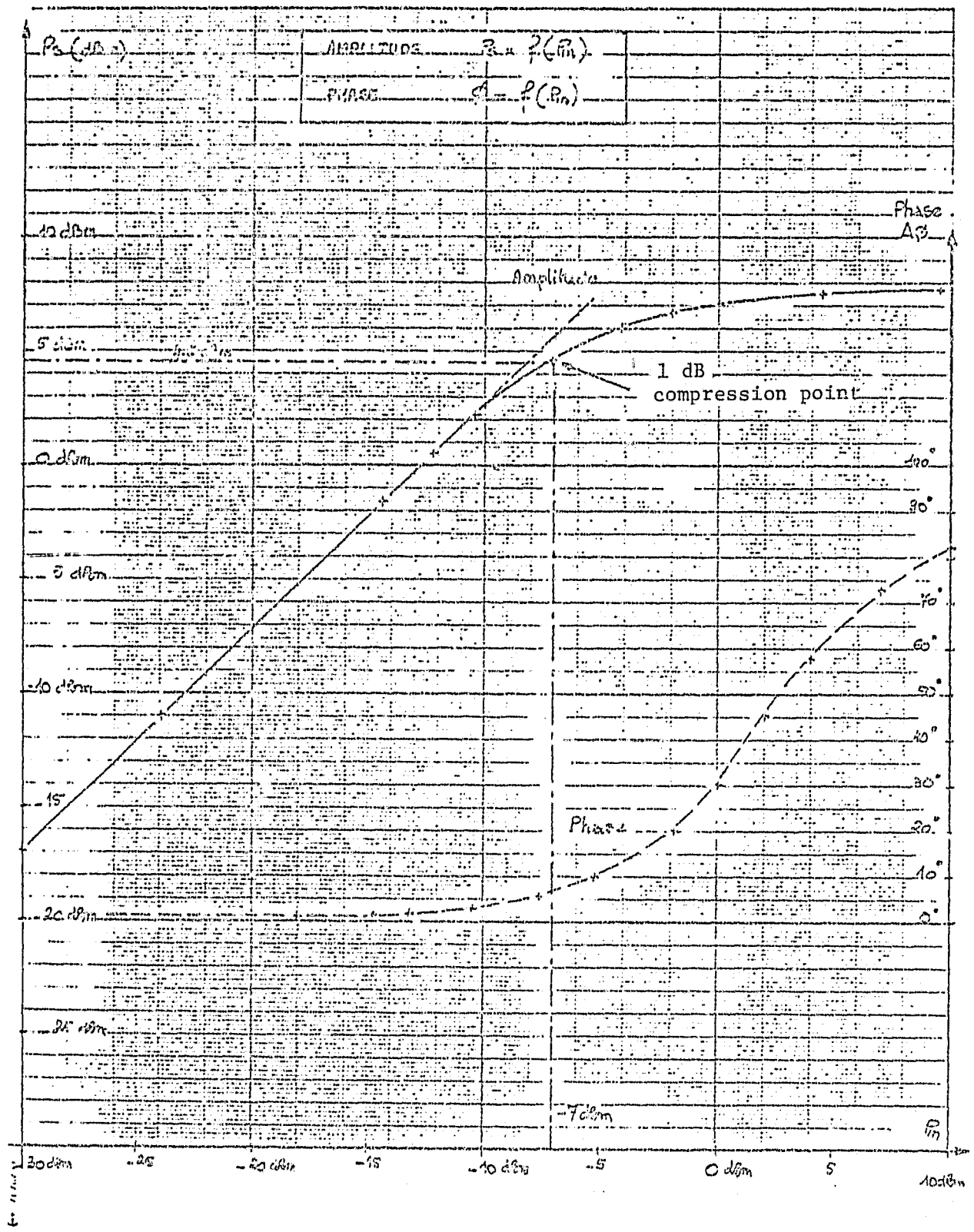
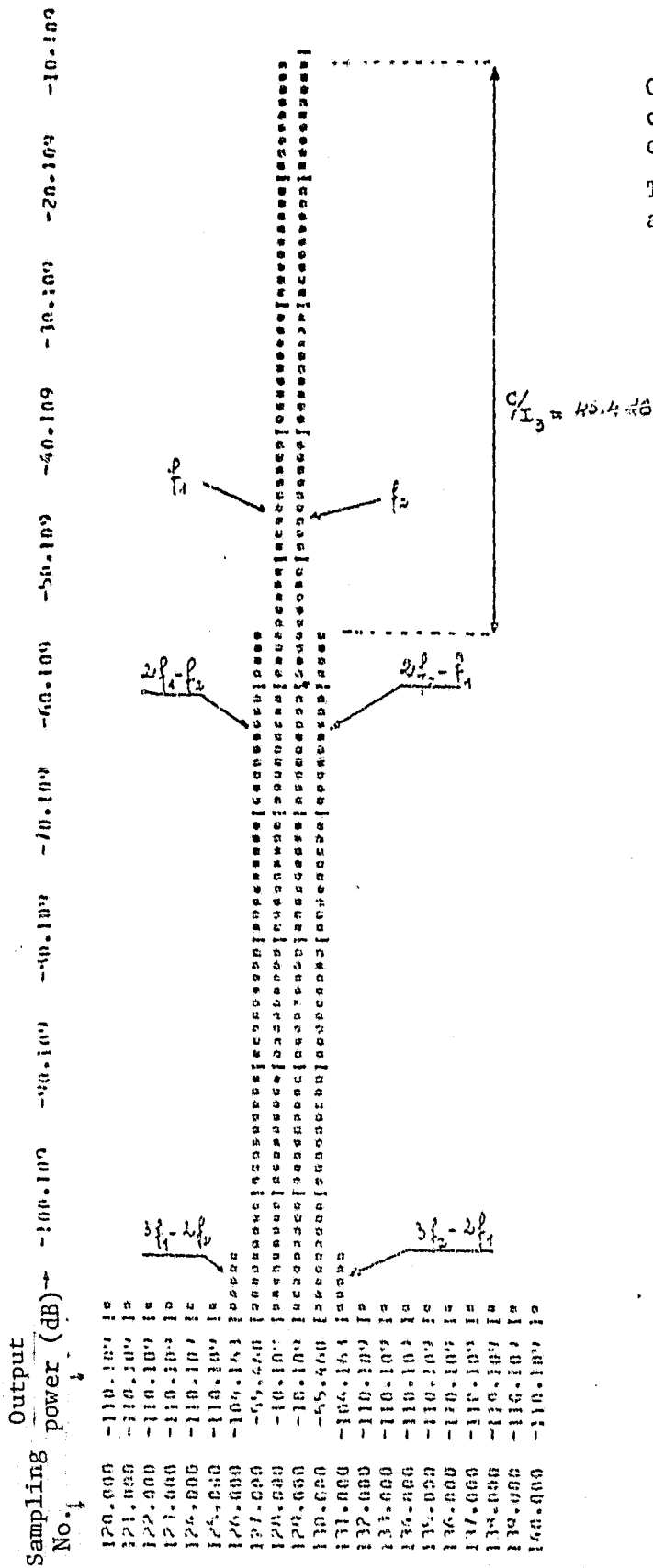


Figure 3

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Case where the input signal consists of two carriers with equal amplitude.

Total input power at the amplifier input:

$$P_{in} = -20 \text{ dBm}$$

Figure 4. Output spectrum of the amplifier

A first analysis of these curves shows a deviation between the measured and calculated values. As far as the "low level" points ( $P_{in} > -10$  dBm) are concerned, the mathematical model selected is found to be insufficient. In effect, the amplifier in this range behaves like a memory element, a hypothesis which has been ignored from the start. In effect, this is not objectionable because the input power range in which we are interested is lower.

Also, points measured for C/I greater than 60 dB have errors because of the fact that we do not know the dynamics of the spectrum analyzer used.

Nevertheless, we find the general variation of the theoretical curve with a slope of 2, which results from the ratio of a curve having the slope 1 ( $P_s = f(P_{in})$ ) and a curve having a slope 3 ( $P_{I_3} = f(P_{in})$ ).

In Figure 5 we show the curves calculated from the program for different intermodulation products of order 3, 5, 7 and 9 as a function of the total input power.

We can note that the interception point is the same for all of them and that its coordinates are:

$$\text{Interception point} \begin{cases} P_{in} = 5 \text{ dBm} \\ P_n = 18 \text{ dBm} \end{cases}$$

Products of order greater than 3 are negligible, which justifies that we only consider products on the order of 3 in this study.

For the hypothesis of a ratio of C/I of 50 dB with 2 carriers equivalent to 28 dB in the favored channel among the 12 equidistant channels with identical levels, we obtain the maximum admissible power of the component utilized (here, a bipolar transistor CG126) or, in other words:

$$P_{s_{max}} = -10 \text{ dBm}$$

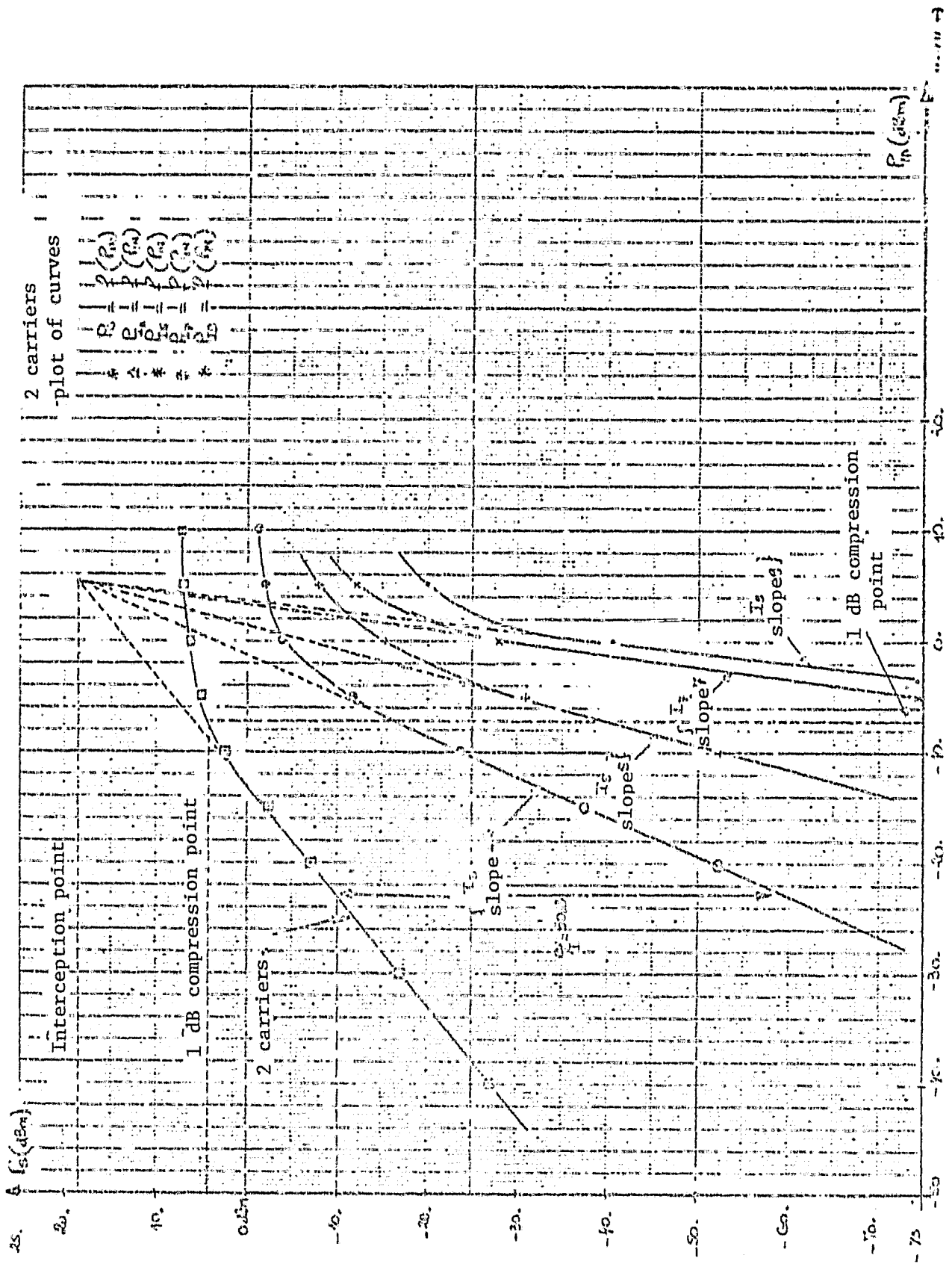
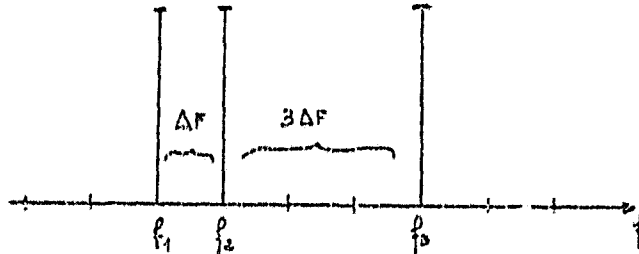


Figure 5

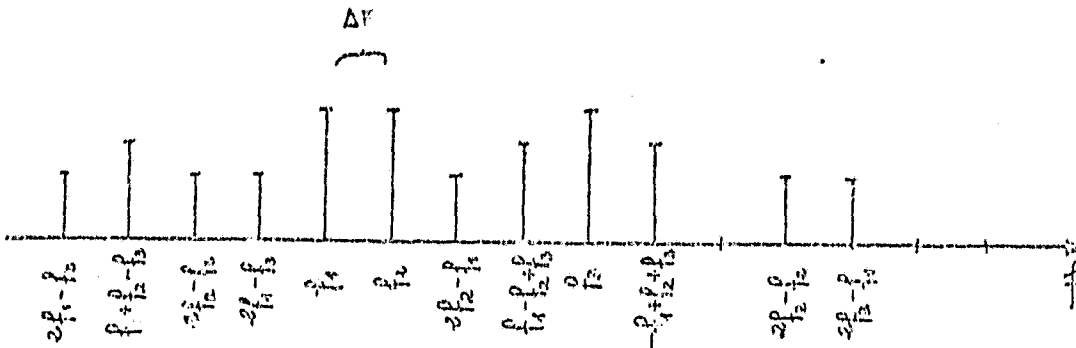
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5.2 Input Signal Consisting of 3 Carriers of Equal Amplitude

The input spectrum has the following appearance:



This configuration of lines allows one to dissociate products on the order of 3 and to add their power levels in the calculation of the global C/I for these products. The following figure shows an output spectrum obtained under these conditions if we are limited to products on the order of 3.



5.2.1 Experiments

The test equipment used for this experiment is similar to the one described before. In addition, the wave generators must have exceptional stability in terms of frequency. The frequency separation  $(f_2 - f_1) = \Delta F$  was selected equal to 5 MHz in order to provide good discrimination of the rays for the analyzer, without excessively widening the frequency spectrum. Nevertheless, the total frequency band is still sufficiently small so that one can neglect the gain-frequency nonlinearity.

### 5.2.2 Simulation

The three lines have the same amplitude but a phase which is equally distributed over the interval  $(0.2 \pi)$ . Figure 6 shows the spectrum calculated for an input power of -25 dBm.

Figure 7 can be used to compare the calculation results and the experimental results for total input powers of -22 dBm and -13 dBm.

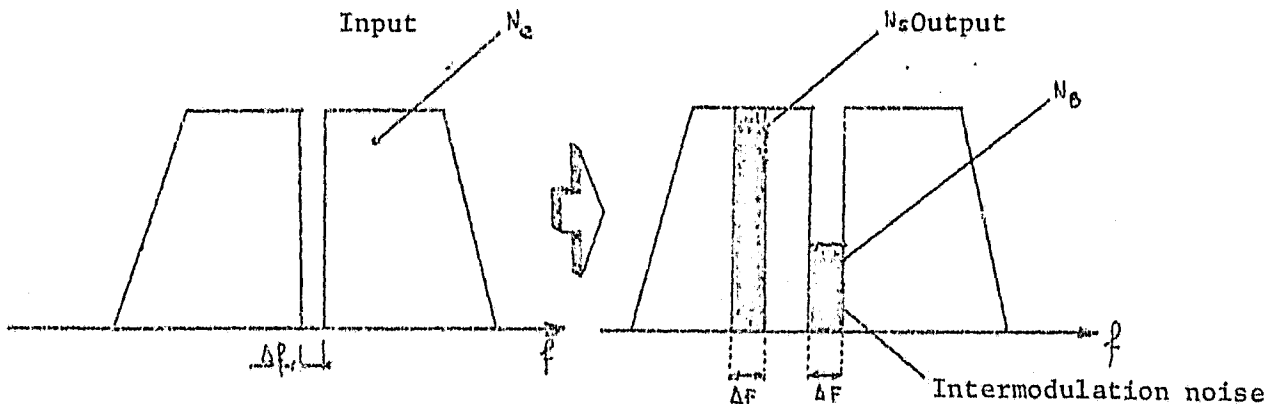
We can see that these results are quite close and, as predicted, the largest product is the one of order 3 having the form  $(\pm f_1 \pm f_2 \pm f_3)$ .

Figure 8 shows the curve  $(C/I) = f(P_{in})$ , where I is the total power of the intermodulation products located between  $f_2$  and  $f_3$ . This allows one to evaluate the ratio C/I for the hypothesis where the difference between  $f_2$  and  $f_3$  is only  $2 \Delta F$ ; this is the least favorable case.

### 5.3 Case Where the Input Signal is a Noise Density

The multicarrier behavior of a nonlinear element can be derived from the behavior obtained in the presence of white noise which is limited within the frequency band of interest.

The nonlinearity is then expressed as a ratio of noise powers. This amounts to finding the ratio of the noise density in a certain frequency band  $\Delta F$  and the noise density in a window with the same width. The following figure describes this operation:



Amplifier Output Spectrum

The input signal consists of 3 carriers of equal amplitude.  
 Total input power  
 $P_{in} = -25 \text{ dBm}$

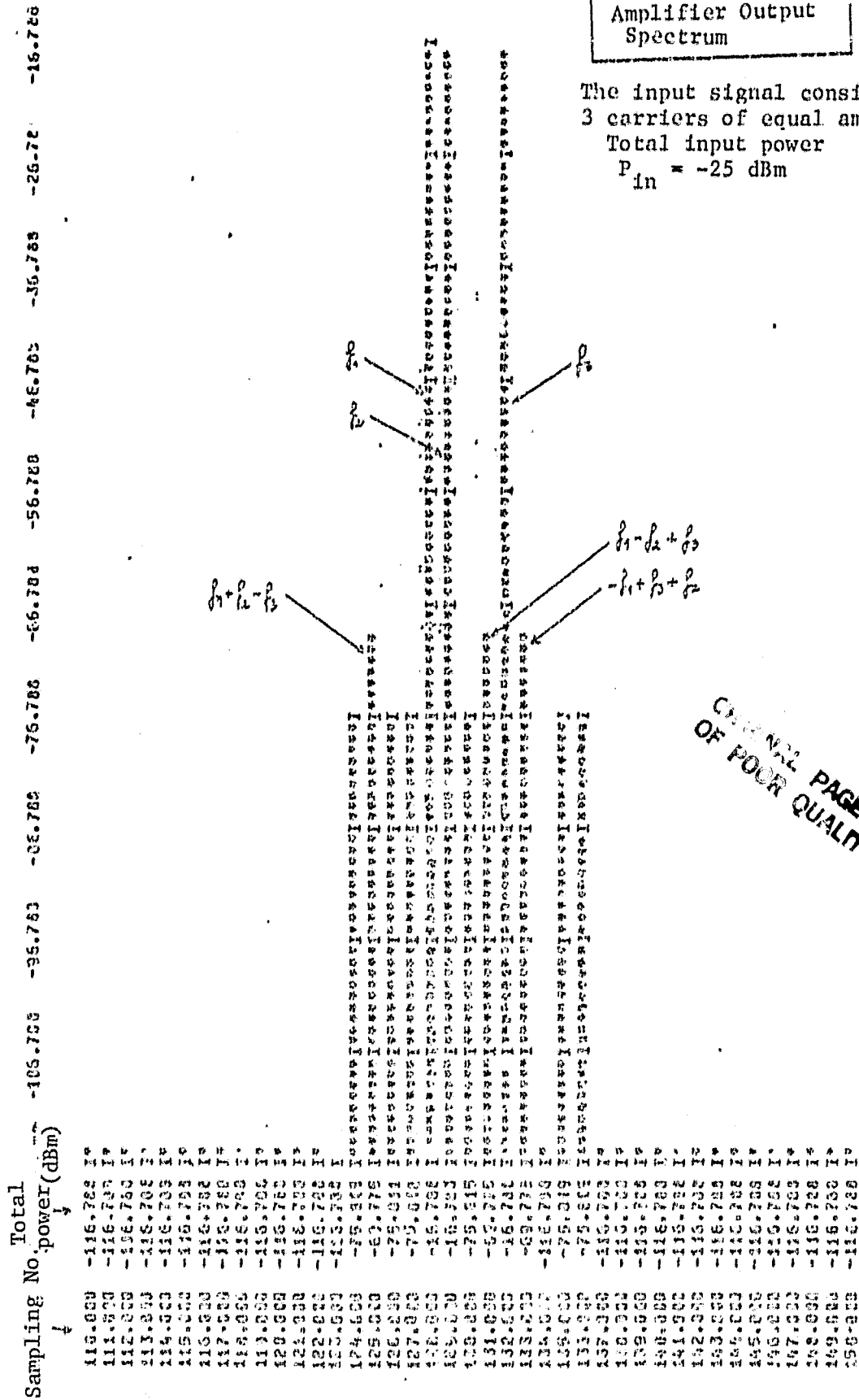


Figure 6

Output Spectrum with 3 Carriers

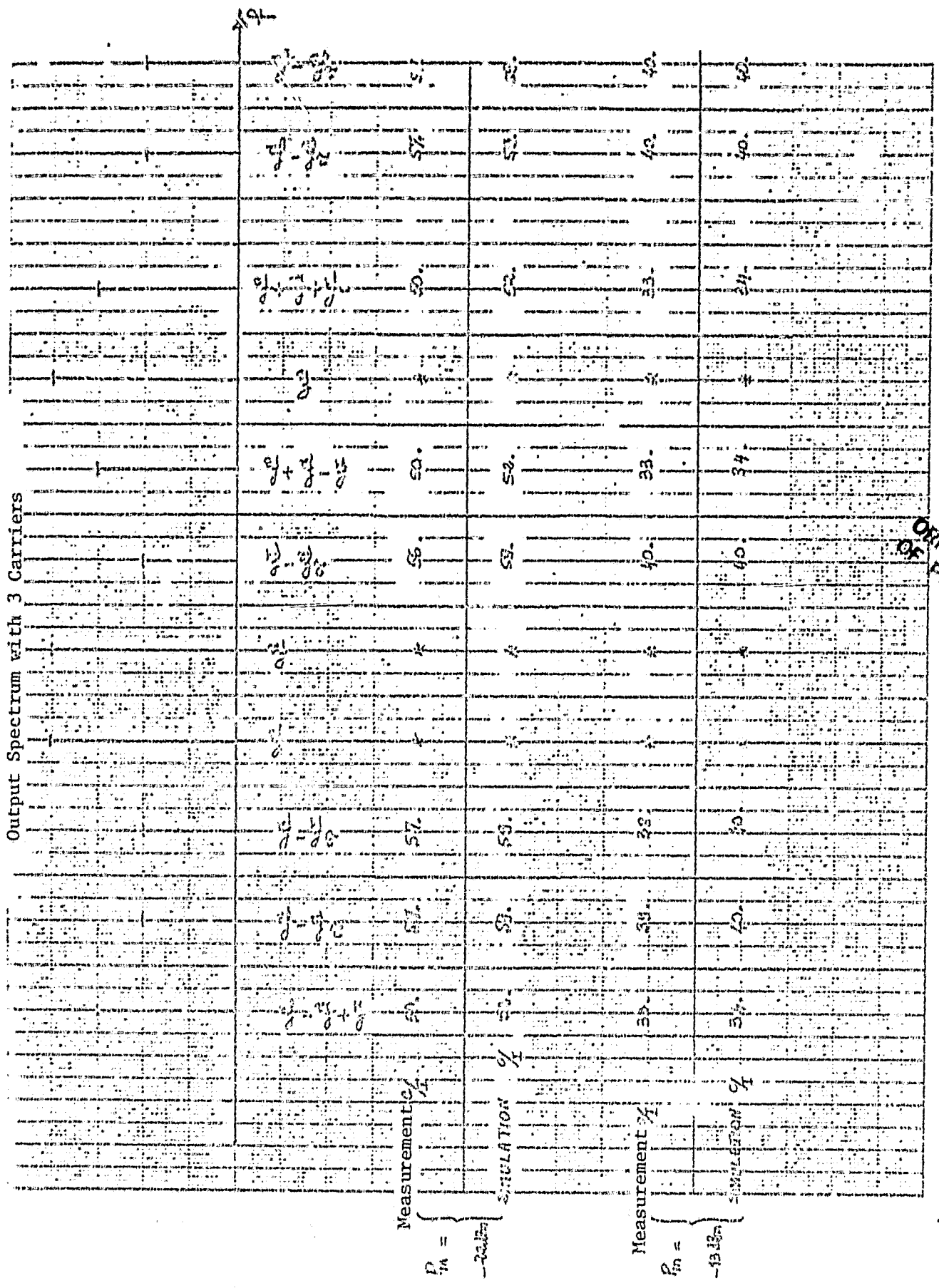


Figure 7

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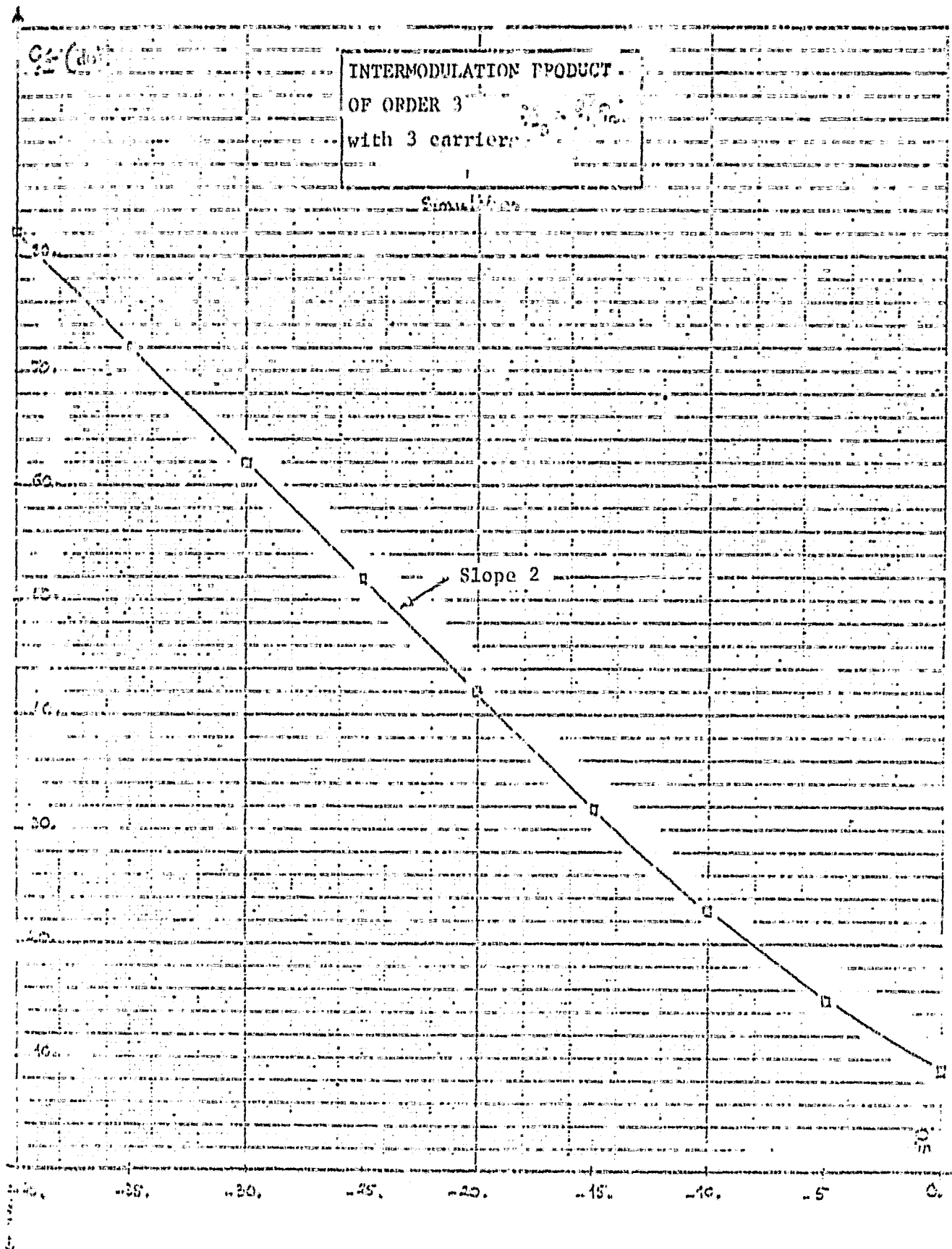


Figure 8

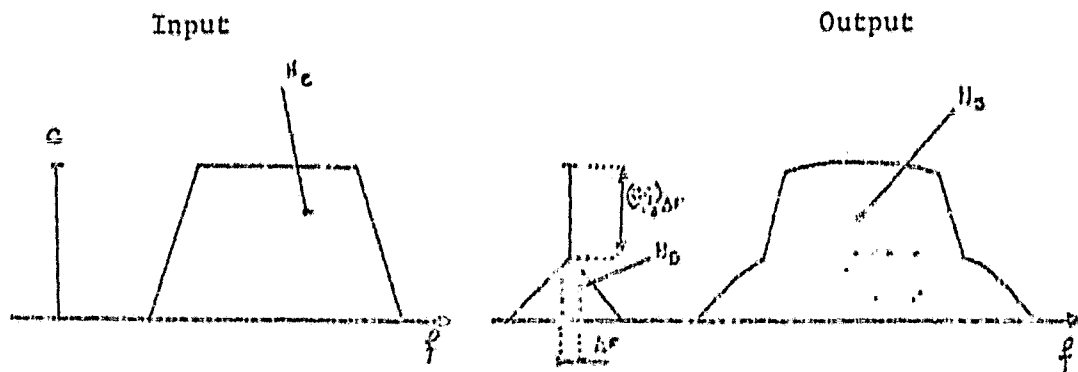
The carrier ratio for intermodulation products in this case is then:

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$$\left(\frac{C}{N}\right)_{\Delta F} = \frac{N_3}{N_3}$$

This method requires the use of a very selective filter of the stop-band type in order to obtain a narrow window with steep edges. Such a filter will require a special study.

We have preferred to associate a sufficiently weak carrier with the noise density, compared with the total noise power around which the intermodulation noise is located.



We then have:

$$\left(\frac{C}{N}\right)_{\Delta F} = \frac{c}{N_3}$$

### 5.3.1 Experiments

Elliptical filters [4] centered around 3960 MHz deliver a noise band of 36 MHz when the carrier is located at 4085 MHz.

The test equipment is shown in Figure 9.

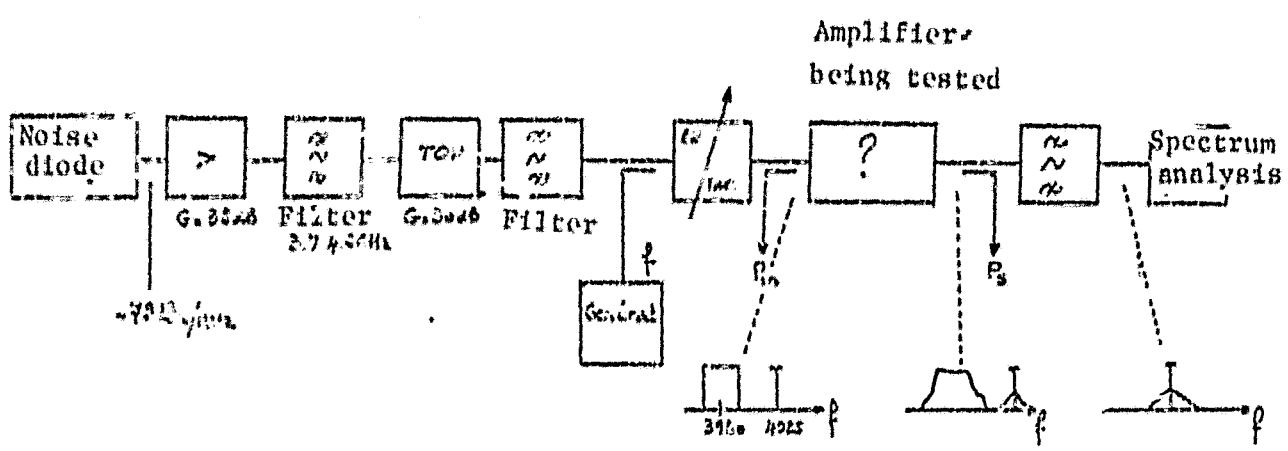


Figure 9

The measurements were made in a 100 KHz analysis band, a value usually used for this type of measurement. This offers an acceptable compromise between the rapidity of acquiring a measurement and the minimum function dynamics of the video band of the spectrum analyzer.

The measurement results are shown in Figure 10.

### 5.3.2 Simulation

In our program, the noise density is simulated by 15 carriers of equal amplitude and having random phases. In order to reduce the number of samplings and therefore the calculation time, the frequency sampling step was selected equal to 1 MHz.

Figure 11 shows the spectrum obtained at the output for a total input power of -20 dBm. Figure 10 shows the results of (C/N) measurements and the (C/N) calculated values, but refer to a 100 KHz band.

The deviation between these two curves for the one hand is caused by the noise model used (15 carriers with a random phase do not represent a sufficiently accurate representation of the noise). Also, it is due to the fact that the noise power measurement of the spectrum analyzer has errors due to the fact that the exact noise bands of the receiver are not known.

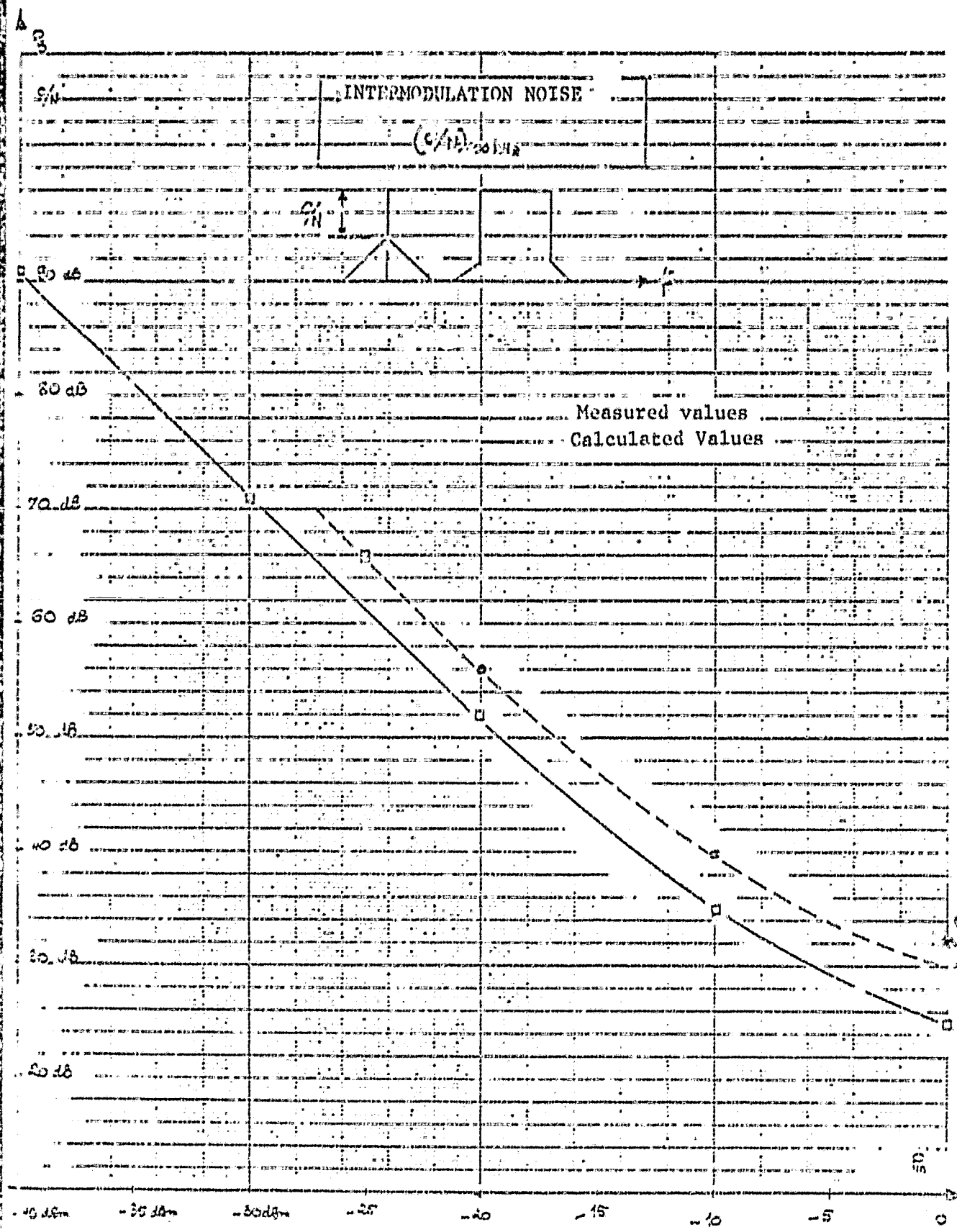


Figure 10

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Input signal consisting of a weak carrier compared with noise intensity (15 carriers with random phase)

$P_{in} = -20 \text{ dBm}$

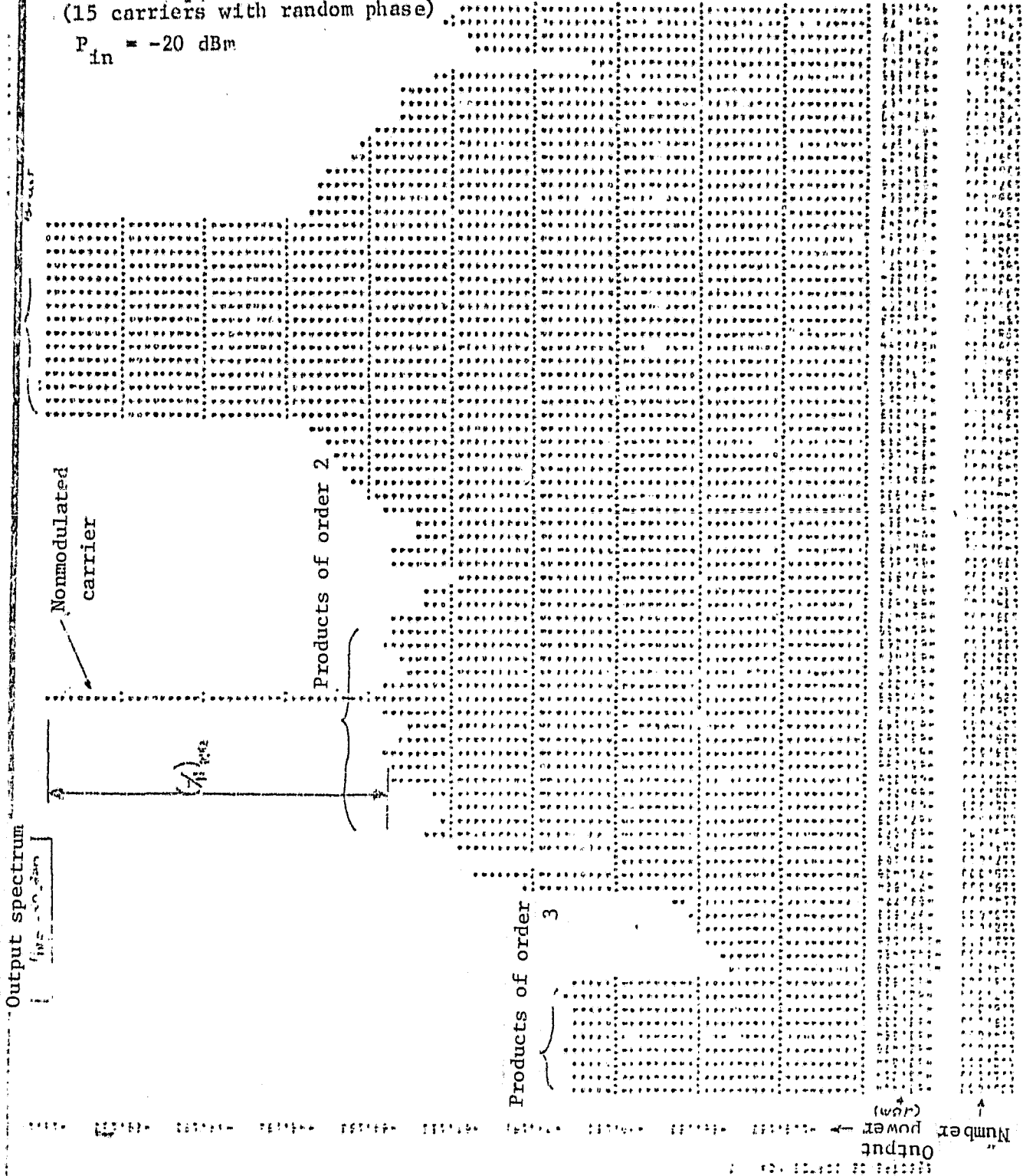


Figure 11

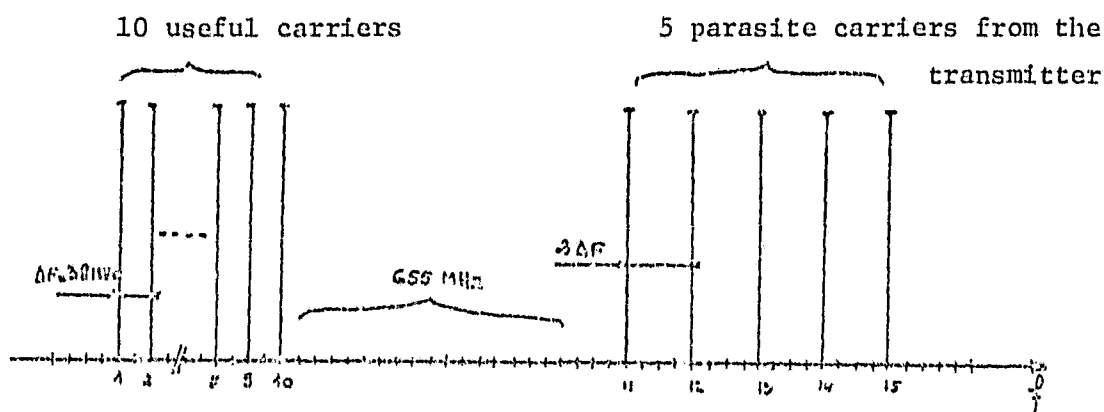
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## 5.4 Application to the Case of a Low Noise Amplifier Which Processes TV Signals

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At the input of a direct TV repeater there are useful signals to be treated, as well as parasite signals which come from the coupling of the transmitter. Even they are strongly attenuated by input filtering elements, these signals retain an amplitude which is sufficiently large to produce inter-modulation products at the input amplitude level.

Simplified, the input spectrum can be represented in the following way:



Among the 10 first carriers, 5 belong to the French network and the others belong to the Luxembourg network.

We are interested in the spectrum at the output of the input amplifier, as well as the (C/N) around each useful carrier which has been shifted beforehand.

The amplifier characteristics were selected similar to those of a parametric amplifier or an FET amplifier, which has the following operating point for a C/I of 61 dB for 2 carriers:

$$\begin{aligned} P_{in} &= -59.1 \text{ dBm} \\ P_S &= -44.35 \text{ dBm} \end{aligned}$$

From these values, the 1 dB compression point calculated is then the following:

$$\left\{ \begin{array}{l} P_{in\_comp} = -38.1 \text{ dBm} \\ P_{s\_comp} = -24.35 \text{ dBm} \end{array} \right.$$

A simulation was made using 4096 samples, random phases for 15 carriers and a frequency sampling step of 1 MHz.

Figure 12 shows the ratio (C/I) of the amplifier for two carriers and the (C/N) for 15 carriers around carrier No. 7, which is the least favorable.

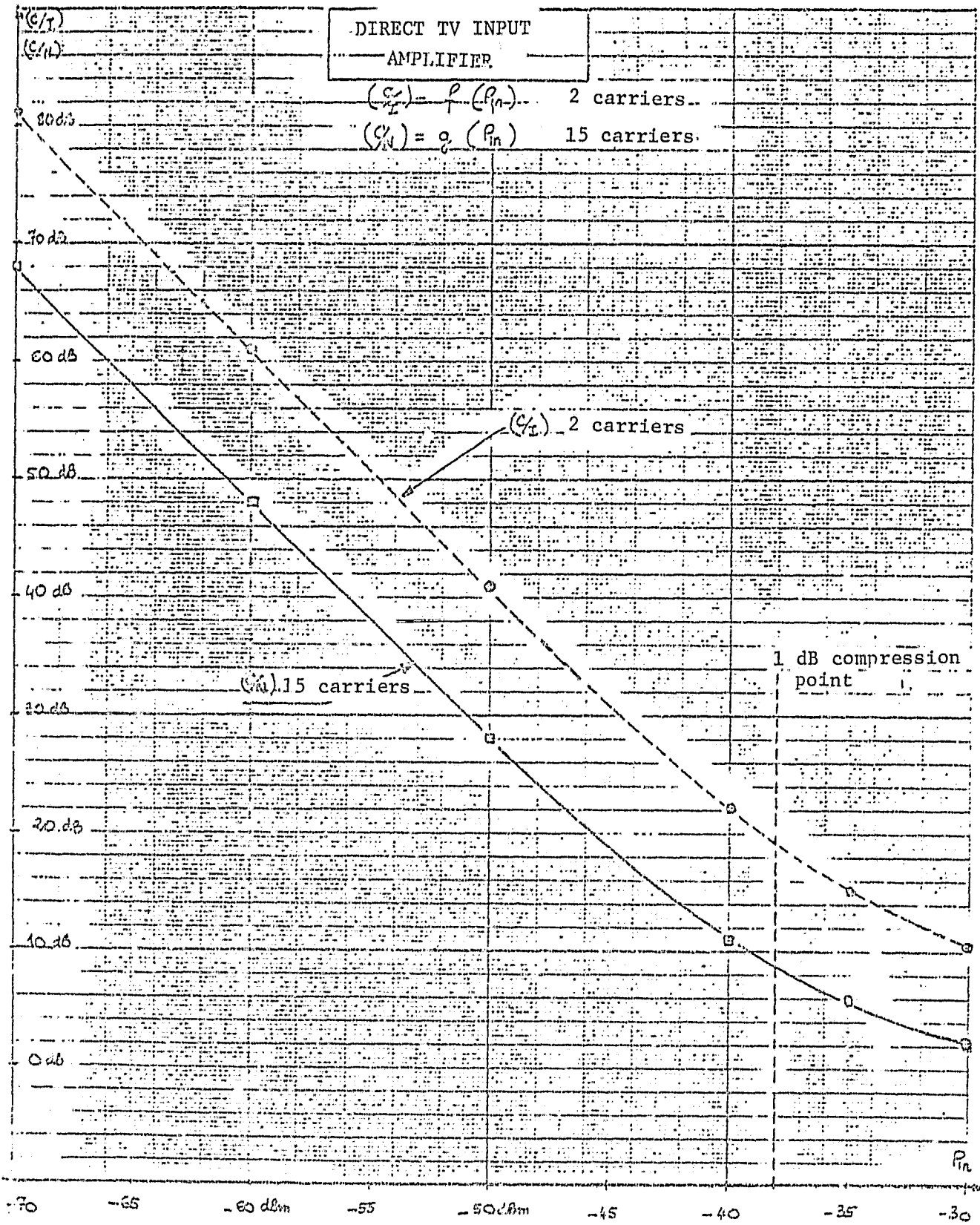


Figure 12

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## 6. CONCLUSION

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The calculation program described within a study of nonlinearities in transistor amplifiers allows one to determine the output spectrum and the carrier ratios with intermodulation products (C/I) in the case where the signals are known in the form of their line spectra.

This program can then be used to evaluate nonlinear elements in the amplitude-phase sense, such as TOP, FET amplifiers, filters, mixers, for which one ignores the multicarrier behavior or when there is noise. In addition, if necessary, one can discriminate between the two kinds of nonlinearities, amplitude and phase.

The tests carried out with an amplifier having an intermediate frequency of a 6-4GHz repeater showed, compared with simulation results, that the selected nonlinearity model agrees especially well with low input power levels.

An additional study carried out jointly with the Mathematics Division will allow a refinement of this model in the case of a nonlinearity with memory.

Finally, application to a case of an input amplifier used in a direct TV transponder showed the influence of the parasite signals due to the emission on the C/I ratio.

The author wishes to thank J. Sombrin and H. Donat for their discussions and the formulation of the problem as well as its programming.

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/32

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Meaning of elements contained in the data cards of  
the nonlinearity simulation program LINA.

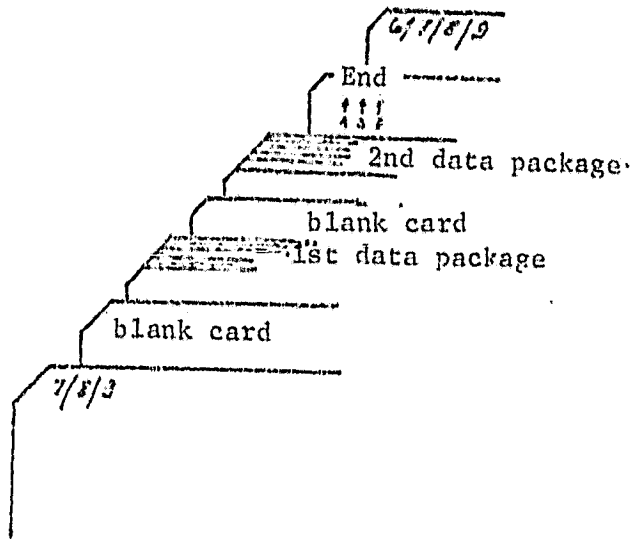
Data cards

Number	Contents	Meaning
1	N, NP, AIN, ASO, DEL, FREQ, FO, N1, N2, PI Comp, P2 Comp	N: number of samples NP: number of carriers AIN: total inlet power min. (dBm) ASO: total inlet power max. (dBm) DEL: increment in input power. These three latter items are referred to the calculation and plotting of the output spectrum. FREQ: total frequency band considered (for the N samplings) (MHz) FO: central frequency of sampling (MHz) N1) output spectrum, will be plotted as a ) listing of the sampling N2) of the sampling N1 with [illegible] N2 PI Comp: input power at 1 dB of compression (dBm) PS Comp: output power at 1 dB of comprssion (dBm)
2	NIP, AMP, PHI	NIP: subscript of the first carrier AMP: amplitude of the carrier (in volts) PHI: phase of the carrier (radians)
3	NIP, AMP, PHI	Other carriers
NP + 2	A(1), A(2), A(3), A(4)	A(4)...A(4): coefficients of the curve output amplifier = f (input amp.)
NP + 3	A(5),... A(9)	A(5)...A(9): continuation of preceding table
NP + 4	B(1),... B(4)	B(1)...B(4): coefficient of the curve phase = g(input amplifier)
NP + 5	B(5),... B(8)	B(5)...B(8): continuation of preceding table

Precautions about use:

Note 1 : We will give a maximum amplitude of 1 to the carrier numbered  $(2^{No. 4} + 1)$

Note 2 : Several data packages can be entered once. For each data package, a [illegible] card is needed; at the end of the last card, a card FIN is required:



Note 3 : The variables (givens) start with N and must have the following format  $I_1$  :

Example: 12, 30, 25 and not : 12., 20., 30.

All the other cards must have the free format.

Example: 2,5, 0.25 E + 01, etc.

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PROGRAM LINA (INPUT,OUTPUT,TAPES=INPUT,TAPES=OUTPUT)
C
C   STUDE CES NON-LINEARITES O, UN AMPLI FE A
C   TRANSECTIONS BIFOLAZES
COMMON N1,N2
COMPLEX Z(4096),ZZ(4096)
DIMENSION X(4096),Y(4096)
DIMENSION A(9),B(9),IN(1024),S(1024)
DIMENSION INLM(27),KALM(27)
DIMENSION PUI(27),OSLFI(27)
COMPLEX G1,G0
INTEGER FIN
ICASE=5
ICPRINT=6
DATA FIN/10HFIN
99  READ (ICASE,33) ICASEN
59  FORMAT (A2J)
    IF (ICASEN.EQ.FIN) STOP
    PI=3.14159
    FACT=1.3
    AY=1.
    UJJ=C.3
    EB=1.0
    EGB=1.0
    IA=579621
    CALL RANSET(IA)
    READ *,N,NP,AIN,ASO,CEL,FREQ,F0,M1,N2,PICCMP,FSCOMP
    WRITE *,N,NP,AIN,ASO,CEL,FREQ,F0,M1,N2,PICCMP,FSCOMP
    NN=N/4
    DO 1 I=1,N
      ZZ(I)=(0.7,0.0)
1    CONTINUE
    DO 2 I=1,NP
      READ *,NIP,AMP,PHI
      ALEA=RANF(IA)
      PHI=(PHI+2*PI)*ALEA
      WRITE *,NIP,AMP,PHI
      EB=AMP*SIN(PHI)
      EB=AMP*CCS(PHI)
2    ZZ(NIP)=CMPLX(EB,EB)
      CONTINUE
      READ *,(A(I),I=1,4)
      WRITE *,(A(I),I=1,4)
      READ *,(A(I),I=5,9)
      WRITE *,(A(I),I=5,9)
      READ *,(B(J),J=1,4)
      WRITE *,(B(J),J=1,4)
      READ *,(B(J),J=5,6)
      WRITE *,(B(J),J=5,6)
      READ *,ITRA1,ITRA2,ITRA3,ITRA4,ITRA5,ITRA6
      WRITE *,ITRA1,ITRA2,ITRA3,ITRA4,ITRA5,ITRA6
      READ *,AMIN,AMAX,STEP,NCI,N3,N4,N5
      WRITE *,AMIN,AMAX,STEP,NCI,N3,N4,N5
      IF(NCI.EQ.1) GO TO 131
      NCI=NCI
      IF(NCI.EQ.-1) NCI=1
      DO 40 J=1,NCI

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READ *,NLM,NLMP
WRITE *,NUP,NUMP
41  INUM(J)=NUM
    KNUM(J)=NUMP
    JJJ=J
47  CCONTINUE
131  CCONTINUE
C
C    CALCUL ENVELOPPE COMPLEXE DU SIGNAL ENTREE
C
VV=( (PI*CCNR-30.)/20.)*ALCG(10.)
VV=1./EXP(VV)
VC=( (PI*SCMP-30.)/20.)*ALCG(10.)
VC=EXP(VC)
M=3
ICPT=0
CF=PREC/N
TR=1./DF
IF (ITRA1.EQ.1.07,ITRA2.EQ.1.08,ITRA3.EQ.1) CALL FLCTS(1,1,1)
AY=(ASC-AIN)/CEL+1.1
JJ=IFIX(AY)
AM=3.0
CCFF=0.
FP=PFMCO(N,ZZ)
WRITE(6,10) SP
17  FORMAT(10X,#FLIS. TOT. INITIALE(CCN)=#,F11.4)
CCFF=AIN-PP
CC 410 I=1,JJ
CC 11 J=1,N
Z(J)=ZZ(J)
11  CCONTINUE
CCFF=AIN-30.
IF(I.EQ.1) 12,13
13  CCFF=CCFF+DEL*(FLCAT(I)-1.)
12  CALL GAG(N,Z,CCFF)
14  FAPA=PFMCO(N,Z)
    AFA=AAPOC(N,Z)
    WRITE(6,120) AFA,FAPA
120  FORMAT(10X,#AMPL. TOT. ENTREE =#,F11.4,#PUIS. TOT. ENTREE =#,
1F11.4)
X1=FLOAT(N)
AM=ALOG10(X1)/ALCG11(2.0)+).1
PP=INT(4M)
IC=0
1072  FORMAT(5(2(1X,1PE12.3)))
    CALL AFFT(Z,INV,S,MM,IC)
    IC=-1
    CALL AFFT(Z,INV,S,MM,IC)
    NC=I
    WRITE(6,1010) NC,FAPA
1010  FORMAT(10X,#COUBRES NC=#,I4,##*#FLISSANCE TOTALE ENTREE P=#,
1F11.4, #DEM?)
IF (ITRA4.EQ.1) 30,31
30  PRINT 22
22  FORMAT(10X,#ENVELOPPE COMPLEXE SIGNAL C,ENTREE#)
    CALL TRACENV(N,Z,FACT,*,*)
31  IF (ITRA1.EQ.1) 60,70

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60 WRITE(6,17) NC,PAPA
10 FORMAT(17Y,*,ENVELOPPE COMPLEXE SIGNAL ENTREE NC=#,I4,
1*TRACE SUR BENSON POUR PUIS. ENTREE TOT. CEP,F11.4,*,CBM*)
C
C
C CURBE LE L'ECCHANTILLON DE SIGNAL
C
C CALL OCCURB1(N,Z,NC,X,Y,T)
C
C PASSAGE DANS S/P AMPLI FI
C
C
C
70 CONTINUE
C0 123 L=1,N
123 Z(L)=Z(L)*VV
CALL AMP1(N,Z,M,A,B)
C0 124 L=1,N
124 Z(L)=Z(L)*VO
1007 FORMAT(10X,I4,I4,F11.3)
IF(ITRAB.EQ.1) 32,33
32 PRINT 21
21 FORMAT(10X,*,ENVELOPPE COMPLEXE SIGNAL SORTIE*)
CALL TRACEN1(N,Z,FACT,*,Y)
33 IF(ITPA2.EQ.1) 30,90
90 PRINT 21,NC,PAP2
21. FORMAT(10X,*,ENVELOPPE COMPLEXE SIGNAL SORTIE NC=#,I4,
1*TRACE SUR BENSON POUR PUIS. TOT. CE #,F11.4,*,CBM*)
CALL OCCURB1(N,Z,NC,X,Y,T)
C
C
C CALCUL DU SPECTRE DE SORTIE
C
C
C
90 CONTINUE
IC=1
CALL APFT(Z,INV,S,MM,IC)
IF(ITRAB.EQ.1) 34,35
14 WRITE(6,36) NC
16 FORMAT(10F1,10X,*,SPECTRE DE SORTIE NC=#,I4,/)
CALL TRASPEC(N,Z,FACT,*,Y)
35 IF(ITRAB.EQ.1) 100,210
C
C
C TRACE DU SPECTRE DE SORTIE
C
C
C
100 CONTINUE
PRINT 171,NO
101 FORMAT(10X,*,SPECTRE DE SORTIE NC=#,I4,*,TRACE SUR BENSON*)
PT=0.0
DF=1.0
CALL SPECTRE(N,Z,NO,*,Y,F0,DF,ICPT,FT,EF)
200 CONTINUE
400 CONTINUE
IF(NC.EQ.1) GO TO 116
AMP=1.0
YY=0.0
C0 110 K=1,JUU
KK=INUM(K)
JK=KNUM(K)
YY=(AMAX-AMIN)/STEP+1.0
KKV=IFIX(YY)
CCFF=)

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CC 90 I=1, KKK
CC 10 L=1, N
Z(L)=ZZ(L)
14 CCITITLE
CCFF=AMTA-77.
IF (I.EG.1) A2=03
45 CCFF=CCFF+UTSP*(FLOAT(I)-1.)
42 CALL GAG(N,Z,CCFF)
17 FAP=CF+CS(N,Z)
FA=FA+DC(N,Z)
FLZ(I)=DAPA

C
C   CALCUL ENVELOPPE SIGNAL ENTREE
C
ID=1
CALL AFFT(2,INV,S,MM,IC)
ID=1
CALL AFFT(2,INV,S,MM,IC)
C   NORMALISATION
CC 125 L=1, N
125 Z(L)=Z(L)*VV
CALL AMF(N,Z,P,A,B)
CC 126 L=1, N
126 Z(L)=Z(L)*VV
ID=1
CALL AFFT(2,INV,S,MM,IC)
GX=Z(KK)
GC=Z(JK)
GC=GC/GX
CC=REAL(GC)
CC=AIMAG(GC)
CSUR(I)=SQRT(CO*CO+CC*CC)
CSUR(I)=SQRT(ALOS10(CSUR(I)))
PRINT
IF(NOT.CC.-1) GO TO 133
GO TO 85)
137 ROTOT=CC
CC 88 IC=NT, L4
IF (IC.EQ.N5) GO TO 86
XN=REAL(Z(IC))
YN=AIMAG(Z(IC))
ROTOT=ROTOT+XN*YK+YN*YK
60 CONTINUE
ROT=10.*ALOG10(ROTOT)
XN=REAL(Z(NS))
YN=AIMAG(Z(NS))
ROK=XK*XN+YK*YK
ROK=10.*ALOG10(ROK)
COU(I)=ROK-ROT
500 CONTINUE
PRINT 132, JK, KK
102 FOR=AN(1,1),#TRACE DU RAPPORT C/2 POUR PORTEUSE N°#J, IN,
1#ET PRODUIT C(INTERMCS, NCH#,24)
CALL TRAC(CSLE I, RUI, KK, JK, KKK)
110 CONTINUE
115 CONTINUE
IF (ITRAZ.EG.1.OF.ITRAZ.EG.1.OR.ITRAZ.EG.1) CALL PLOT(D, I, G, 999)
GO TO 85
END

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