SELF MIXING OSCILLATORS AT Q BAND

USING GUNN DIODE SOLID STATE SOURCES

Thesis for the Master of Science Degree
Faculty of Engineering
University of Cape Town

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1982

ACKNOWLEDGEMENTS

The author wishes to thank the management of Plessey South Africa for allowing him to do this research in their laboratories and the Department of Electrical Engineering at the University of Cape Town for making this possible and Mr John Greene of the Department for his supervision.

The author would also like to thank Jan Schreuder and Bob Evans of Plessey and Dr Barry Downing of RECCI for many usefull discussions and encouragement. Also he would like to thank the members of the CA2000 team for their forbearance during the many months it took to prepare this thesis.

Also the author would like to thank his family and friends for their patience when he was unavailable or just simply drowning under a sea of paper.

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SUMMARY.

This Thesis is a report of research done into the use of a GUNN Diode Millimetre Wave Source as both the Reciever and the Transmitter of a system.

The Thesis is devided into three main sections. The first of these sections (Chapter 2) "sets the scene" so to speak. It presents certain preliminary fundamentals that are essential for placing the results obtained later into their proper context. Areas discussed are NOISE FIGURE, TRANSMISSION PATH LOSS and GUNN DIODE operation.

The second section (Chapter 3) considers the operation of the device as a mixer. It's operation as a Microwave Power Source is a well known concept and it will not be discussed here. This section covers such aspects as CONVERSION LOSS, NOISE FIGURE and BIAS CIRCUIT STABILITY.

The third section considers those points arising from the second section which seem to be of interest. These include such aspects as Better Correlation of results and theory and ways of applying the device to modern Waveguide structures.

The Thesis ends with a Conclusion, Appendix and a Bibliography.

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CHAPTER 1

INTRODUCTION

1.1 Justification for the investigation

The need for cheap receivers at Millimeter Wave frequencies has in recent years stimulated interest in the SELF MIXING OSCILLATOR concept. In particular, CW applications such as DOPPLER RADAR and Electronic Distance Measurement need low noise mixers because of their limited power output capabilities. Separate Mixer Diodes for Millimeter Wave applications can be very expensive and the Noise Figures obtainable still leave much to be desired. In addition the application of separate Mixer Diodes requires extra mounting structures and extra design effort and hence extra expense. This means that there remains a whole class of applications where optimum range performance is not at a premium and where price considerations outweigh performance considerations. The use of a GUNN Device as both the source of output power and as the receiver of the system starts to look attractive, especially if it is considered that Noise Factors as good as 12dB could be obtainable.

The GUNN effect was discovered in the early sixties (37). By the end of the sixties the GUNN effect was already in widespread use as a microwave source and HAKKI (38) had as early as 1966 suggested its use as a mixer. It was only in 1970 that ALBRECHT and BECHTLER (8) did any thorough investigations into the self mixing properties of a GUNN source. After this a group at the University of Lancaster in England (2,3,4 & 5) did some investigations into the S.M.O. concept with (5) suggesting threshold powers as good as -93dBM. Subsequent work by CHREPTA and JACOBS (6) suggested Noise Figures as low as 15.5 dB and an even more recent paper from Lancaster (39) still confirmed the threshold powers of approximately -90dBM found by them.

No more recent work has been done to date, and it was thus felt that a new wide ranging investigation was required. The Object of this thesis is to explain the mechanism of mixing and to provide practical information for the Microwave Engineer attempting to use this technique.

1.2 Layout of the Thesis

Chapter 1 is the project overview, of which this section is a part. It is hoped that this section will enable the reader to select only those parts of the thesis which are of interest to him. The various different sections can be viewed independantly.

In Chapter 2 it is hoped to present some preliminary considerations. The subject of Noise Figure will be discussed and what its implications are to range performance will be shown. Some useful graphs and tables will be included with full explanations as to their origin. Also in Chapter 2 is a summary of the GUNN effect and an explanation of a GUNN Oscillator. Noise effects in GUNN Oscillators will also be considered.

Chapter 3 will present a view as to how the device operates as a mixer. The effects of the various parameters will be discussed. The matching of the device to the IF Amplifier will be discussed and the gain and stability effects will be evaluated.

Chapter 4 will attempt to show where simplifications in the theory as presented in Chapter 3 cause results and theory to differ. A computer numerical analysis will be used to obtain a more acurate harmonic analysis of the mixing waveform. The alternative approach of a Negative Resistance Amplifier followed by a non linear device will also be discussed. Other modes of oscillation will be considered and such topics as harmonic mixing will also be discussed.

Chapter 4 will also consider future developements such as Dielectric Waveguide mounting and monolithic circuits mounted directly into Waveguide.

Chapter 5 is a conclusion. A bibliography will be included at the end of the thesis.

<u>Note:</u> The symbol E is used to indicate exponentiation of 10. e.g. 5E-3 is equal to $5x10^{-3}$.

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Chapter 2

PRELIMINARY CONSIDERATIONS.

2.1. Review of the definition of Noise Figure.

A system can be specified in terms of various parameters. Among these parameters are Frequency Response, Dynamic Range, Power output and Noise Figure. The significance of Noise Figure is that it specifies the minimum possible signal power that is usable through the system.

2.1.1 Symbols Used.

S_i = Available input signal power

N_i = Available input noise power

S_o = Available output signal power

N_o = Available output noise power

T = Temperature in Kelvin

k = Boltzmann's constant
= 1.38E-23

B = Bandwidth in Hertz

G = Gain of the system

F = Noise Figure

NF = Noise Factor in dB
= $10\text{Log}_{10}F$

The concepts of Noise Temperature are not dealt with here. A fuller description of the subject of NOISE is available in references 22, 40 and 41.

2.1.2 Noise Figure.

A system can be represented as follows

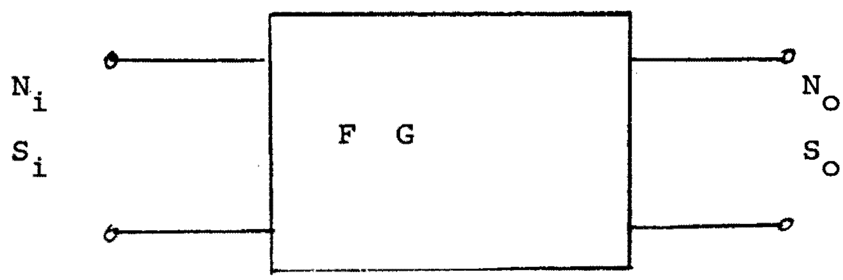

Fig.2.1

The Noise Figure F is then defined as follows.

$$F = \frac{S_i/N_i}{S_o/N_o}$$

This definition implies that the inputs and outputs are correctly matched.

By manipulation of the expression it is possible to show that -

$$F = \frac{N_o}{G(kTB)}$$

This expression is very significant as will be shown later. Its significance is that it gives rise to the noise figure of the SMO directly.

2.1.3 Cascaded Systems

A cascaded system can be represented as shown below.

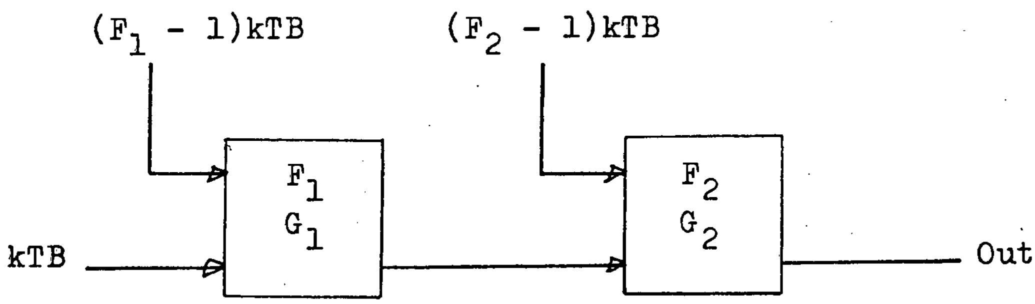

Fig. 2.2

Output noise of the first network = F_1G_1kTB

Output noise of an ideal network = G_1kTB

Noise generated in the first network

$$= F_1G_1kTB - G_1kTB$$
$$= (F_1-1)kTBG_1$$

Thus the noise generated in the first network can be represented as an input to the network.
Similarly for the second network.

Now using a new definition for Noise Figure as follows:-

$$F = \frac{\text{Noise output of the system}}{\text{Noise output of the ideal system}}$$

Therefore:-

$$F = \frac{F_1 kTBG_1 G_2 + (F_2 - 1)kTBG_2}{kTBG_1 G_2}$$

$$F = F_1 + (F_2 - 1)/G_1$$

Or more fully

$$F = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_1 G_2$$

(Friis' Formula)

2.1.4 Direct Mixer Noise Figure.

A variation of the expression in section 2.1.2 can be used to show that the Noise Figure of a reciever using a mixer as the front end is given by:-

$$F = L_c(F_2 + t - 1)$$

where L_c = Mixer conversion loss.

F_2 = Noise Figure of the I.F. Amp.

t = Mixer Noise Temperature.

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2.2 Noise Figure related to System Performance.

The range performance of a system can be directly related to its noise performance. If we return to the expression for Noise Figure we will see that the ratio So/No appears in the denominator. This ratio, together with F can be used to calculate the threshold power input of the receiver.

2.2.1 Threshold Input Power.

If we consider the basic block diagram of the receiver we will see that it consists of a Front End, IF Amplifier, Detector and Low Frequency Amplifiers. Now considering Friis Formula as given in 2.1.3 we note that the Noise Figures of the latter stages become insignificant if the gains of the preceding stages are high. In a receiver system the gain of the IF Amplifier will be quite high and hence the Noise Figure of the detectors and following stages can be ignored.

To calculate the threshold power at the input we must know the worst possible signal to noise ratio that the detectors will be able to use. As a rule of thumb and a good starting point we can assume this ratio to be equal to 1. This implies that the expression for Noise Figure simplifies as shown:-

$$F = \frac{S_i/N_i}{S_o/N_o}$$

But $S_o/N_o = 1$

$$F = S_i/N_i$$

$$S_i(\text{thrsh}) = F.N_i$$

or more completely

$$S_i(\text{thrsh}) = F.N_i(S_o/N_o)$$

2.2.2 Calculation of maximum path loss

Assuming that we now know the transmitter output power (P_o) and the gains of the respective receive and transmit antenna (G_r and G_t) then we can calculate the possible path loss. Where path loss is L_p.

$$S_i = P_oG_rG_t/L_p$$

or possible path loss:-

$$L_p = \frac{P_oG_rG_t}{S_i}$$

Or if the values are in dB.

$$L_p(\text{dB}) = 10\text{Log}(P_o/S_i) + 10\text{Log}(G_r) + 10\text{Log}(G_t)$$

But $S_i = F.N_i$

$$L_p(\text{dB}) = 10\text{Log}(P_o/(FN_i)) + 10\text{Log}(G_r) + 10\text{Log}(G_t)$$

Or $L_p(dB) = 10Log(P_o/N_i) - 10Log(F) + 10Log(G_r) + 10Log(G_t)$

2.2.3 Calculation of Range from path loss

The attenuation of a Microwave signal is generally made up of three components.

a) Attenuation due to free space.

$L_s = 20Log(4PiR/l)$

Where R is distance in meters
and l is wavelength in meters

b) Attenuation due to atmospheric absorbtion. (La).

This is usually due to the oxygen and water vapour in the atmosphere, and it can be read from a table such as that shown in Fig. 2.3. (Hughes wall chart. 1981)

c) Attenuation due to rainfall.
This attenuation starts to be of major importance at millimeter wave frequencies. The particular attenuations can again be read off from an appropriate chart as in Fig 2.4 (Hughes wall chart. 1981)

The overall attenuation over any path can thus be calculated as follows :-

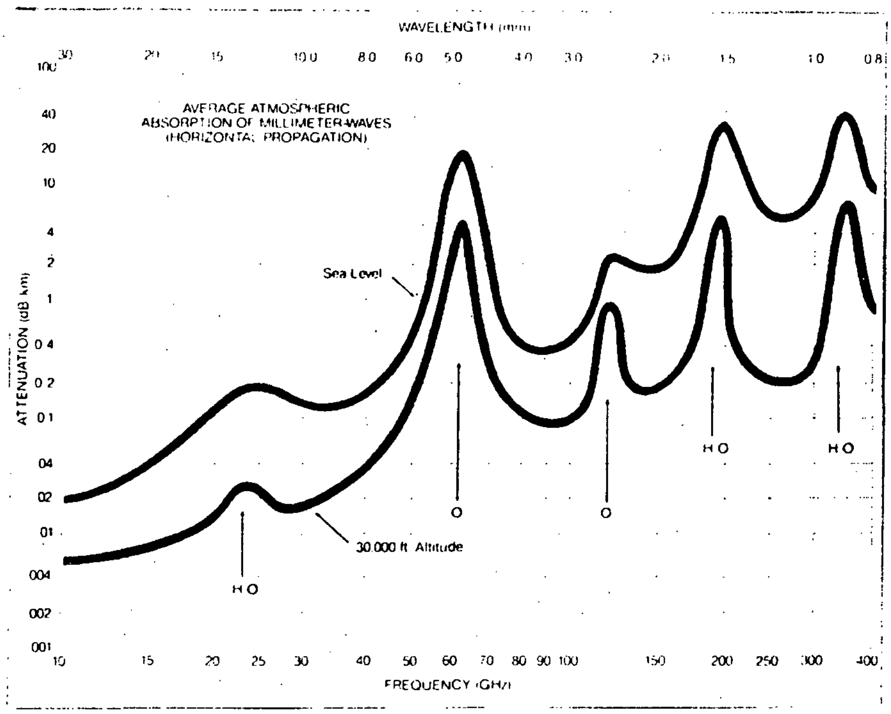

Fig 2.2 Attenuation due to Atmospheric absorbtion.

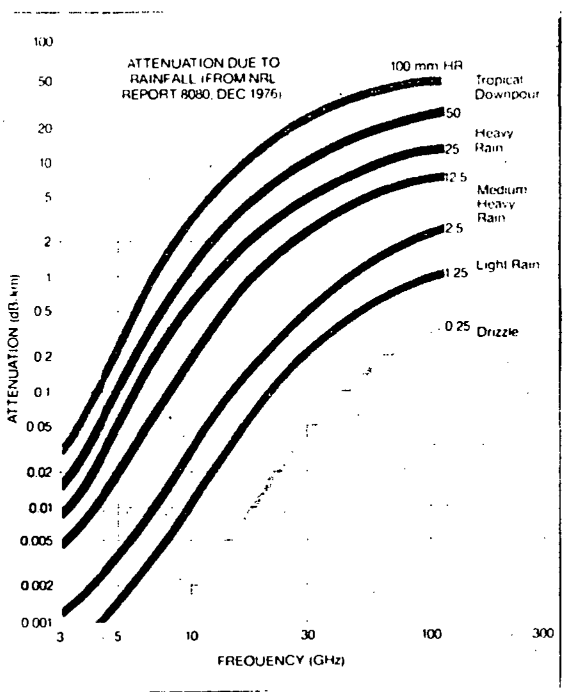

Fig 2.3 Attenuation due to Rainfall.

$$L_p(dB) = 20Log(4PiR/1) + R(L_a + L_r)/1000.$$

Where L_a and L_r are read off from the appropriate table and R is distance in meters.

A set of tables has been compiled which more clearly shows the effects of the various attenuations. Three frequencies are chosen as being three so called "windows".

See overleaf.

Path Attenuation at 10 GHz. (dB)

DISTANCE	a+b	.25	1.25	12.5	50
10	72	72	72	72	72
20	78	78	78	78	78
30	82	82	82	82	82
40	84	84	84	84	85
50	86	86	86	86	86
60	88	88	88	88	88
70	89	89	89	89	89
80	91	91	91	91	91
90	92	92	92	92	92
100	92	92	92	92	93
200	98	98	98	99	99
300	102	102	102	102	102
400	104	104	104	105	105
500	106	106	106	107	107
600	108	108	108	108	109
700	109	109	109	109	110
800	111	111	111	111	111
900	112	112	112	112	112
1000	112	112	112	113	113
2000	118	119	119	119	121
3000	122	122	122	123	125
4000	124	125	125	125	129
5000	126	127	127	128	132
6000	128	128	128	129	134
7000	129	129	130	131	136
8000	131	131	131	132	139
9000	132	132	132	134	141
10000	132	133	133	135	143
20000	138	139	139	143	159
30000	142	143	143	149	173
40000	144	145	146	153	185
50000	146	148	148	157	197
60000	148	149	150	161	209
70000	149	151	151	165	221
80000	151	152	153	168	232
90000	152	154	154	171	243
100000	152	155	155	174	254

Path Attenuation at 35 GHz. (dB)

DISTANCE	a+b	.25	1.25	12.5	50
10	83	83	83	83	83
20	89	89	89	89	90
30	93	93	93	93	93
40	95	95	95	95	96
50	97	97	97	97	98
60	99	99	99	99	100
70	100	100	100	100	101
80	101	101	101	102	102
90	102	102	102	103	103
100	103	103	103	104	104
200	109	109	109	110	112
300	113	113	113	114	116
400	115	115	116	116	120
500	117	117	117	118	123
600	119	119	119	120	126
700	120	120	120	122	128
800	121	122	122	123	130
900	122	123	123	124	132
1000	123	124	124	126	134
2000	129	130	130	134	152
3000	133	133	134	140	166
4000	135	136	137	144	180
5000	137	138	139	148	193
6000	139	140	141	152	206
7000	140	142	143	156	218
8000	141	143	144	159	230
9000	142	144	145	162	243
10000	143	145	147	166	255
20000	149	153	156	194	372
30000	153	159	163	220	487
40000	155	163	169	245	601
50000	157	167	174	269	714
60000	159	170	179	293	827
70000	160	174	184	316	939
80000	161	177	189	340	
90000	162	180	193	363	
100000	163	182	197	386	

Path Attenuation at 90 GHz. (dB)

DISTANCE	a+b	.25	1.25	12.5	50
10	92	92	92	92	92
20	98	98	98	98	98
30	101	101	101	101	102
40	104	104	104	104	104
50	106	106	106	106	107
60	107	107	107	108	108
70	108	108	109	109	110
80	110	110	110	110	111
90	111	111	111	111	113
100	112	112	112	112	114
200	118	118	118	119	122
300	121	121	121	123	127
400	124	124	124	127	132
500	126	126	126	129	136
600	127	128	128	132	140
700	128	129	129	134	143
800	130	130	131	136	147
900	131	131	132	137	150
1000	132	132	133	139	153
2000	138	139	140	152	180
3000	141	143	145	163	205
4000	144	146	149	173	229
5000	146	149	153	183	253
6000	147	151	155	191	275
7000	148	153	158	200	298
8000	150	155	161	209	321
9000	151	157	163	217	343
10000	152	159	166	226	366
20000	158	172	186	306	586
30000	161	182	203	383	803
40000	164	192	220	460	
50000	166	201	236	536	
60000	167	209	251	611	
70000	168	217	266		
80000	170	226	282		
90000	171	234	297		
100000	172	242	312		

Now the maximum Path Loss can be calculated as follows:-

$$L_p(dB) = 10Log(P_o/N_i) - 10Log(F) + 10Log(G_r) + 10Log(G_t)$$

The maximum useable range can now be read off the appropriate table.

Example

Say N_i = kTB = 1E-15
P_o = 50E-3 watts
F = 25.5dB
G_r = 27dB
G_t = 30dB

Therefore $L_p(max)$ = 168dB

Using the charts gives the following ranges.

Distance	a+b	0.25	1.25	12.5	50
10 GHz	>100Km	>100Km	>100Km	>100Km	>100Km
35 GHz	>100Km	50Km	40Km	10Km	3Km
90 GHz	70Km	20Km	10Km	4Km	1Km

2.2.4 Conclusion

Concluding this section it can be pointed out that NF subtracts directly from the maximum usable Path Loss. This fact is independant of either the band width of the receiver or the ratio of signal power to noise power at the detector.

Further it can be noted that the Noise Factor of most Microwave Systems can be directly related to the front end. e.g. The Mixer Diode. The IF Amplifier works at low frequencies and hence can have good Noise figures.

The price of the Microwave Mixer Diode relates directly to the systems performance.

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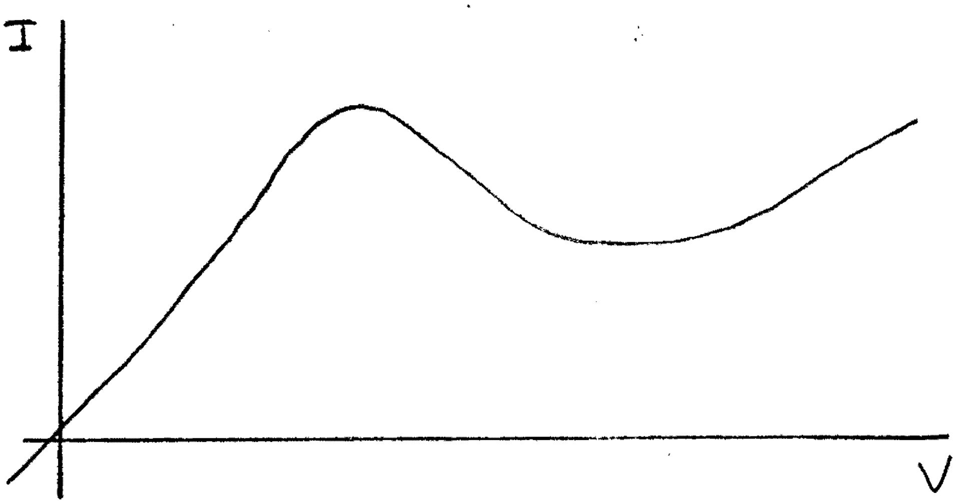
2.3 GUNN Theory :- The Transfered Electron Effect

It is hoped in this section to describe the operation of the GUNN Diode and hence to explain the dynamic characteristic of the GUNN Diodes transfer function.

2.3.1 The Transfer Function

The current to voltage characteristic can be drawn for low frequencies as in Fig 2.4.



The characteristic can generally be approximated by two straight lines as the Bias Voltage added to the Local Oscillator swing will generally never take the voltage into the area of the ultimate bulk resistence.

2.3.2 The Transfered Electron Effect

The shape of the characteristic is due to to what is called the Transfered Electron Effect. This can be summarised as follows.

2.3.2.1 The Two Valley Model Theory

Gallium Arsenide is one of a class of semi-conductors which has unusual properties of the electrons in its conduction band.

i.e. The electrons are to be found in one of two different valleys in the conduction band. The unusual property is that the electrons in the higher energy valley in fact have lower mobility than electrons in the lower energy valley.

The conductivity, given by s, of the n-type GaAs is given as follows:-

$$s = e(m_l n_l + m_u n_u) \quad \text{- - - - - - - - - - -} \quad (1)$$

Where e = electron charged

m_l = mobility in the lower valley

m_u = mobility in the upper valley

n_l = electron density in the lower valley

n_u = electron density in the upper valley

If the applied electric field is increased some electrons will scatter into the upper valley. If some electrons are still present in the lower valley a negative conductance region exists. When all electrons are scattered into the upper valley the conductance is once more positive.

This is known as the Two Valley Model Theory and the mathematical description is as follows.

Differentiating (1) with respect to E gives :-

$$ds/dE = e[m_1(dn_1/dE) + m_u(dn_u/dE)] + e[n_1(dm_1/dE) + n_u(dm_u/dE)] \quad \text{- - -(2)}$$

If the total number of electrons is given by $n = n_1 + n_u$ and it is assumed that m_1 and m_u are proportional to E^p, where p is constant, then :-

$$\frac{d(n_1 + n_u)}{dE} = \frac{d_n}{dE} = 0 \quad \text{- - - - - - - - - - - - - - -(3)}$$

$$\frac{dn_1}{dE} = -\frac{dn_u}{dE} \quad \text{- - - - - - - - - - - - - - - - - - - (4)}$$

Also dm/dE is proportional to dE^p/dE which is equal to pE^{p-1} and also to pE^p/E, which is again proportional to pm/E.- - - - - - - - - - - - - (5)

By substituting equations (3) and (5) into (1) we get :-

$$ds/dE = e(m_1 - m_u)dn_1/dE + e(n_1m_1 + n_um_u)p/E \quad \text{- - - - (6)}$$

Now Ohms Law J = s.E can be differentiated with respect to E as follows:-

$$dJ/dE = s + E.ds/dE$$

This may be rewritten as:-

$$\frac{1}{s}\frac{dJ}{dE} = 1 + \frac{ds/dE}{s/E} \quad \text{-----------} \quad (7)$$

Now in a negative resistance region the ratio dJ/dE must be negative. This would be true if:-

$$-\frac{ds/dE}{s/E} > 1 \quad \text{---------------} \quad (8)$$

Now putting equations (1) and (6) into (8) gives:-

$$\frac{(m_l - m_u)}{(m_l + fm_u)} \times \frac{-Edn_l}{n_l dE} - p > 1$$

Where $f = n_u/n_l$.

Notes:

1) The exponent p is a function of the lattice scattering mechanism and should be negative and large.

2) The term $(m_l - m_u)/(m_l + fm_u)$ must be positive to satisfy the inequality. Thus m_l is greater than m_u. Electrons begin in a low mass valley and transfer to a high mass valley. The maximum value of this term is unity, implying that m_l is much greater than m_u.

3) The term dn_1/dE must be negative. This quantity represents the rate with which electrons transfer to the upper valley with respect to the electric field, and will depend on differences between electron densities, electron temperatures and energy gaps between the two valleys.

In practice this means that the semi-conductor should satisfy three criteria in order to exhibit negative resistance.

a) The energy level between the bottom of the lower and the bottom of the upper valley must be much larger than the thermal energy at normal ambient temperatures. ie greater than 0.026eV.

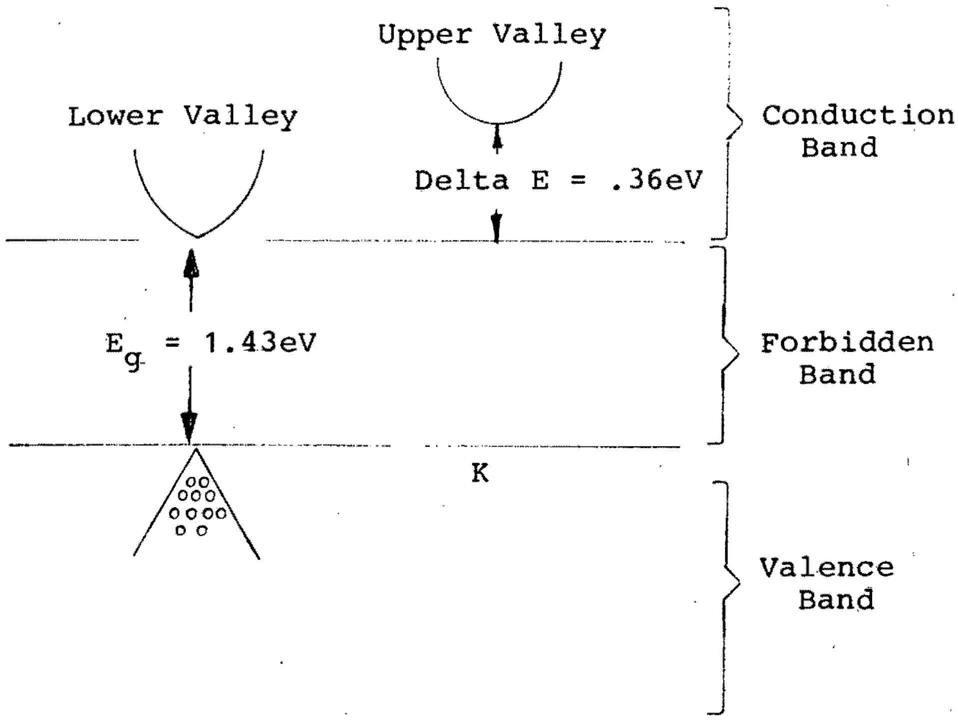
b) The energy difference between the two valleys must be smaller than the difference between the conduction and valence bands.

c) The electron velocities must be much lower in the upper valley than in the lower valley.

Some of the semi-conductors that satisfy these requirements are Gallium Arsenide (GaAs), Indium Phosphide (InP) and Cadmium Telluride (CdTe).

The Transfered Electron Effect can be shown graphically as in Fig 2.5.

In practice the characteristic of GaAs is as in Fig 2.6.



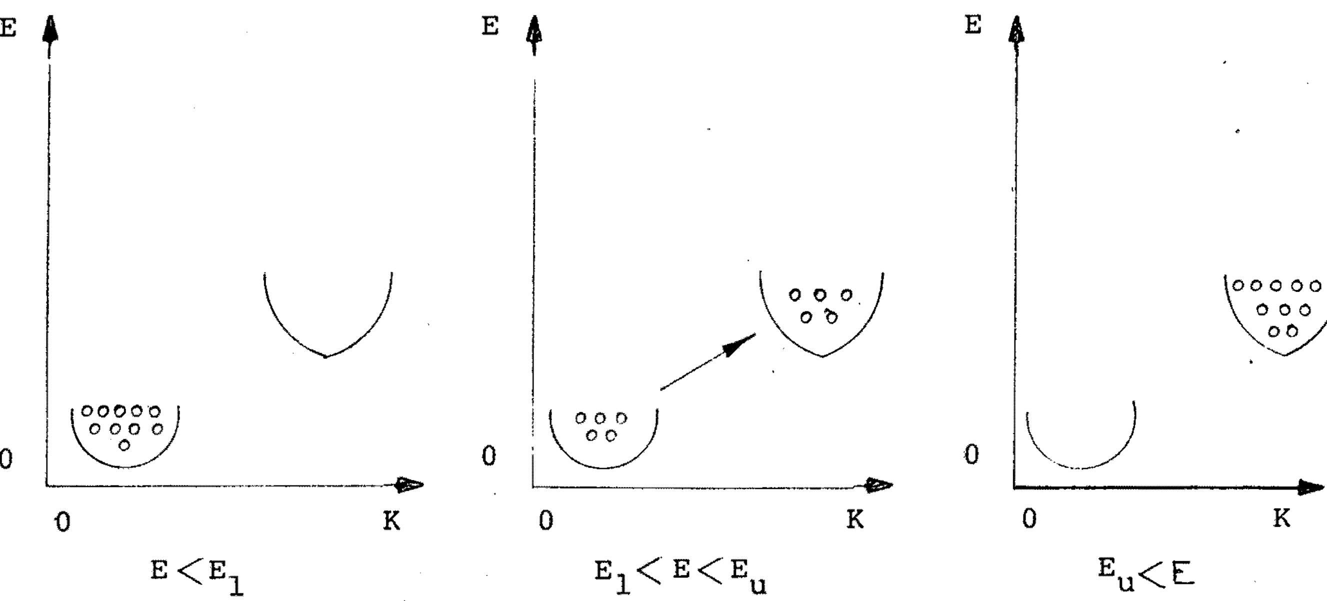
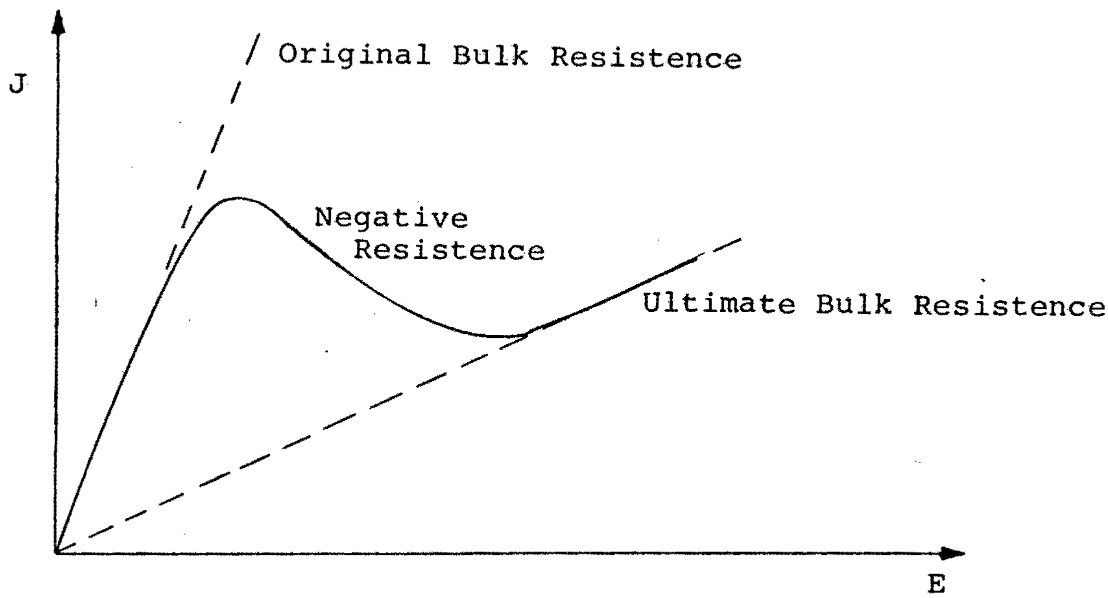


Fig 2.5. <u>The Transfered Electron Effect</u>



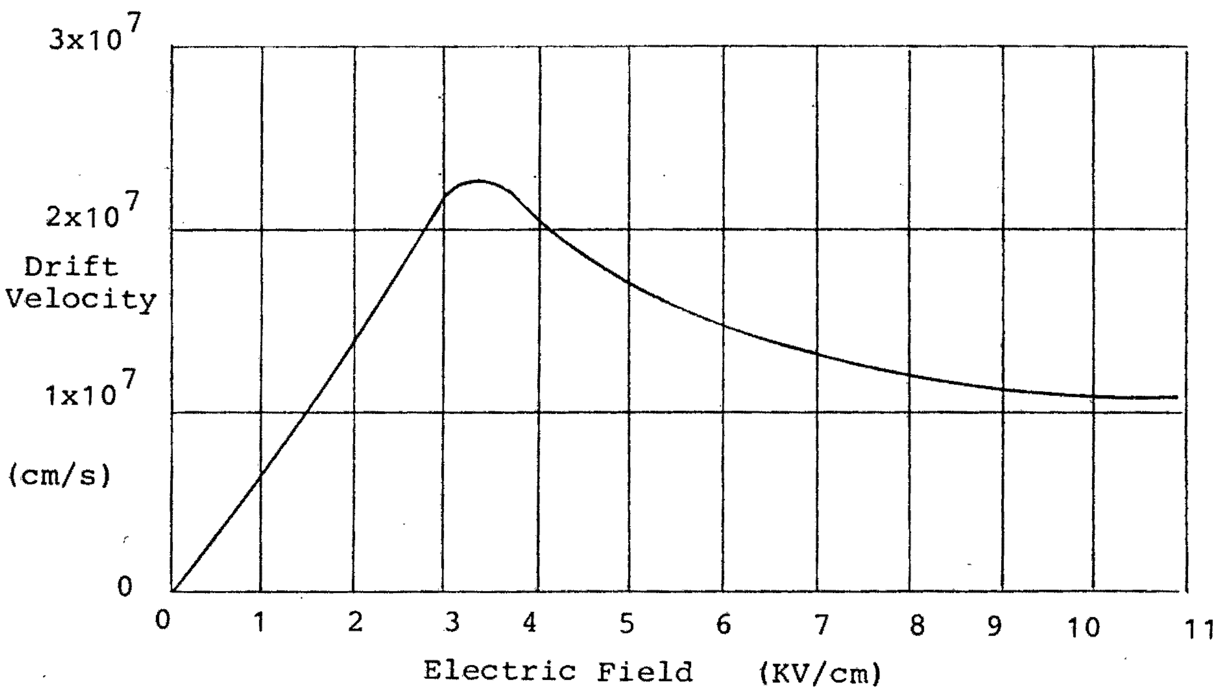


Fig 2.6 Characteristic of GaAs.

2.3.2.2 The High Field Domain Theory.

The previous section, 2.3.2.1 explained why the bulk semi-conductor exhibits negative resistance. It is possible to take this negative resistance as an observed phenomenon and to apply it to an RLC circuit and then to explain the resulting oscillations by means of Circuit Theory. This "circuits" approach will not however help the semi-conductor physisist to explain the oscillations of current in the device on its own, nor will it help him in the design of Solid State Sources. The High Field Domain Theory attempts to do this.

In the n-type GaAs the majority carriers are electrons. When the applied voltage is low the Electric Field and conduction current density are uniform throughout the device. The conduction current density is given by:-

$$J = s.E_x = s.U_x.V/L = p'.v_x.U_x$$

where:-

J = Conduction current density.

s = Conductivity.

E_x = Electric field in the x direction.

L = Length of the device.

V = Applied voltage.

p' = Charge density.

v_x = Drift velocity in the x direction.

U_x = Unit vector in the x direction.

The density of donors less the density of acceptors is called the "Doping". When the space is zero the carrier density is equal to the doping. If the applied voltage is above the threshold value of about 3000 V/cm then a high field domain is formed near the cathode that reduces the electric field in the rest of the material and causes the current to drop to about 2/3 of its maximum value. This can be illustrated by the expression for applied voltage:-

$$V = -\int E_x dx$$

V is constant and hence the electric field is non linear.

The high field domain then drifts with the carrier stream across the electrodes and disappears at the anode contact.

In general the high field domain has the following properties:-

1) A domain will start to form whenever the electric field in a region of the sample increases above a certain threshold field.

The domain will drift through the sample with the carrier drift.

2) On applying additional voltage the domain will increase in size and absorb more voltage than was added and current will decrease.

3) A domain will not disappear before reaching the anode unless the voltage drops appreciably below threshold.

4) The formation of a new domain can be prevented by reducing the voltage below threshold.

5) A domain will modulate the current in the device as it passes through regions of different doping or different cross section.

6) The domain's length is generally inversely proportional to doping.

The formation of an electron dipole layer in GaAs can be shown as in Fig 2.7.

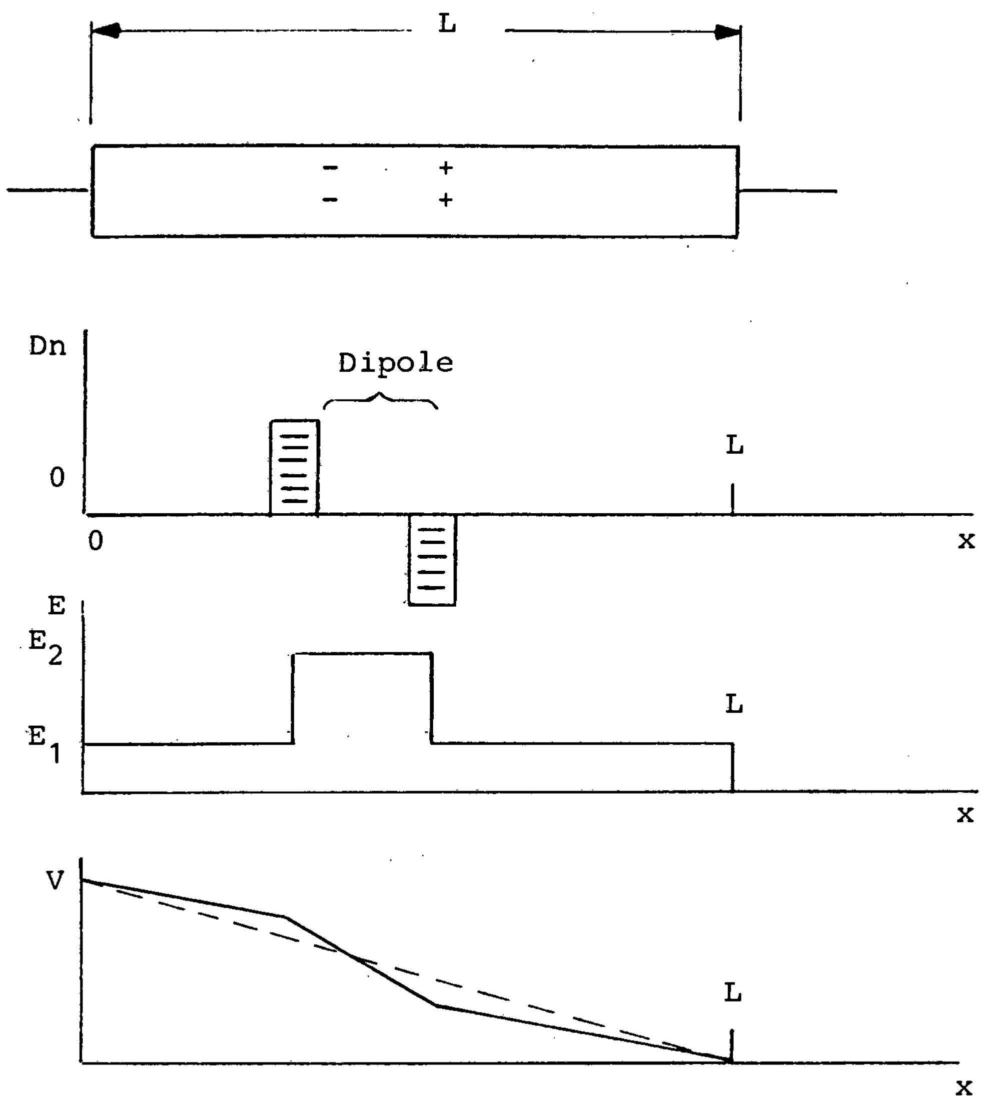


Fig 2.7 Formation of a Dipole Layer.

2.3.3 Modes of Operation

The four modes of operation are as follows:-

First : GUNN Oscillation Mode
Second : Stable Amplification Mode
Third : LSA Oscillation Mode
Fourth : Bias Circuit Oscillation Mode

The first, second and fourth modes are of interest to us in this investigation. The first will be described here, the second in Chapter 4 and the fourth in Chapter 3.

2.3.3.1 GUNN Oscillation Modes

GUNN observed perturbations of current in the bulk semi-conductor at a frequency given by:-

$$F = V_{dom}/L_{eff}$$

Where V_{dom} is the domain velocity
L_{eff} is the effective length of the sample

These oscillations will generally occur if the product n_oL lies in the range of $10E12/cm^2$ to $10E14/cm^2$.

a) Transition Time Domain Mode

This implies that the device is in a non-resonant circuit.

$Vd = vs = f_L = \pm 10E7$ cm/s

b) Delayed Domain Mode

$10E7$ cm/s $> f_L > 10E6$ cm/s

The field would be oscillating more slowly than the transit time of the device. ie. $t_o > t_t$

c) Quenched Domain Mode

$f_L > 2 \times 10E7$ cm/s

This means that t_o is less than t_2. The domain is quenched before it reaches the anode.

2.3.4 Practical GUNN Sources

In general the device will be operated in a resonant cavity coupled to a waveguide. In many cases it will be loosely coupled to a varactor to provide tuning of the centre frequency of the source. (Refer to 30).

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2.4 Noise Generation in GUNN Oscillators.

In this section it is hoped to describe the various forms of noise to be found close to the centre frequency of a GUNN Oscillator. The Noise at a frequency shifted from the carrier by the I.F. frequency is important as it appears as the intrinsic noise input of the Self Mixing Oscillator.

2.4.1 Thermal Noise Generation.

This noise is generally referred to as the A.M. noise of the device. It is only of significance at high IF frequencies as it is usually several orders of magnitude lower than the F.M. noise. It is actually difficult to measure as its amplitude is very low. It can be simply described as follows.

The device has a certain resistance and when operating has a certain noise temperature. This generates the usual White Noise. This noise power amplitude modulates the GUNN in a cavity of a certain Q. The bandpass characteristic is now imposed onto the White Noise generator giving rise to the spectrum appearing as in Fig 2.8.

This diagram excludes the F.M. noise generation. Josenhans in reference 17 measured a noise to carrier ratio of -120dB at a point 100KHz from the carrier. In our application of the Self Mixing Oscillator we are generally interested in frequencies above 10MHz from carrier and hence we can expect N/C ratios of at least as large as -160dB.

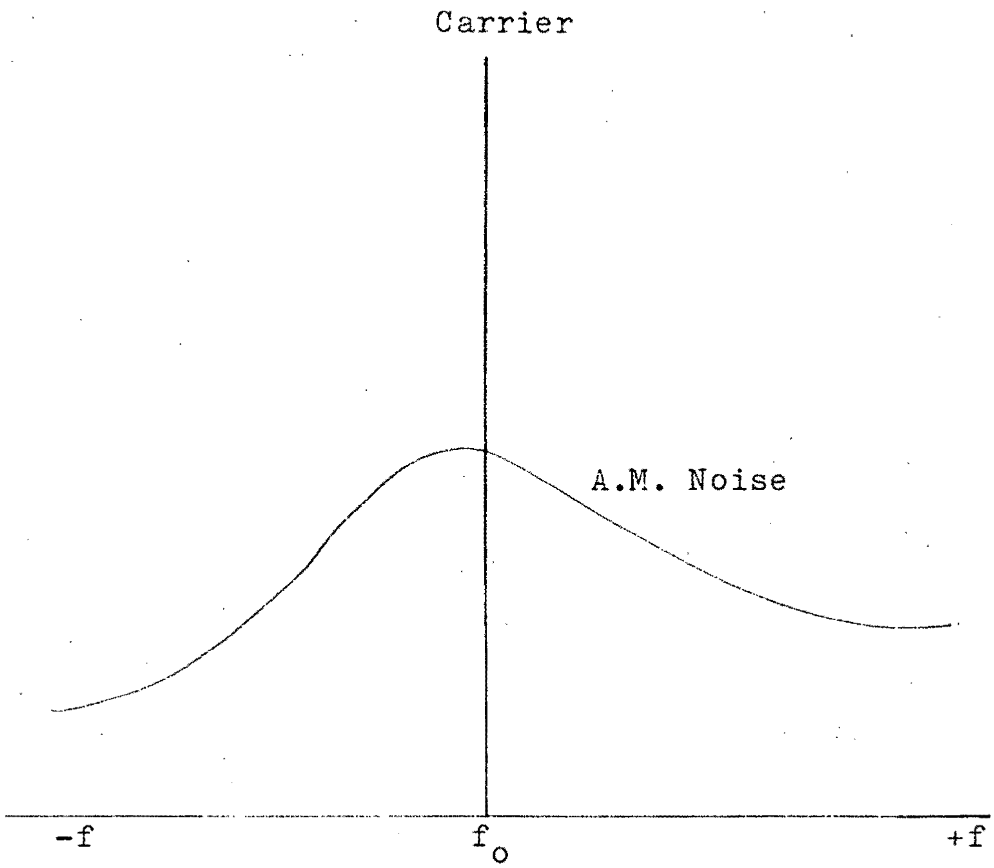

Fig 2.8. White Noise of the GUNN Source.

2.4.2 Upconverted Noise Generation.

This noise spectrum takes on a somewhat different appearance to that of the A.M. spectrum. It is upconverted noise and is generated by the process of F.M. modulation of the GUNN source.

The frequency of the GUNN source has a dependance on the supply voltage and the current flowing in the device. The bulk semiconductor exhibits Shot Noise, Flicker Noise and Thermal Noise in its current flow and hence the center frequency is modulated according to some function of these noise currents.

The mathematical treatment in general involves a statistical analysis of the modulation current and hence an application of this function to the modulation spectrum. This treatment is however not very clear in its explaination of the actual process of upconversion. A clearer description could be as follows.

Given that the current in the device is equal to the quiescent current(I_o) added to the noise current(I_n).

Now $F_o = f(K, I_o, I_n)$

If K and I_o are constant then :-

$d(F_o)/dI_n = f(I_n)$

or more correctly:-

$delta(F_o)/delta(I_o) = f(I_n) = g(t)$

as I_n is itself a function of t. The instantaneous signal output of the GUNN oscillator can be written as follows:-

$$S_o = SCos(w_o t + M\int g(t)dt) \text{ - - - - - - - - } (a)$$

The conversion of such a time domain expression into a frequency domain expression is very difficult unless g(t) can be represented sinusoidaly. In fact I_n can be shown to have a spectral form with a maximum at I_o and tailing off at the rate of 6dB per octave or at another rate dependant on the ratio of $1/f^v$. If this spectrum where applied to expression (a) the modulation process should result in a similar spectral pattern around the centre frequency of the GUNN diode.

Matsuno in reference 25 shows that in fact the ratio of F.M. noise to carrier power drops by as much as 9dB per octave in the region close to carrier. Matsuno measured the noise spectrum of a cavity controlled GUNN oscillator with various Qs and his results are repeated here as in Fig 2.9.

A fuller explanation of GUNN diode noise is beyond this scope of this chapter. The reader is referred to references 16, 17, 19, 22, 23, 25, 33, 35 and to Section 3.4 following in the next chapter.

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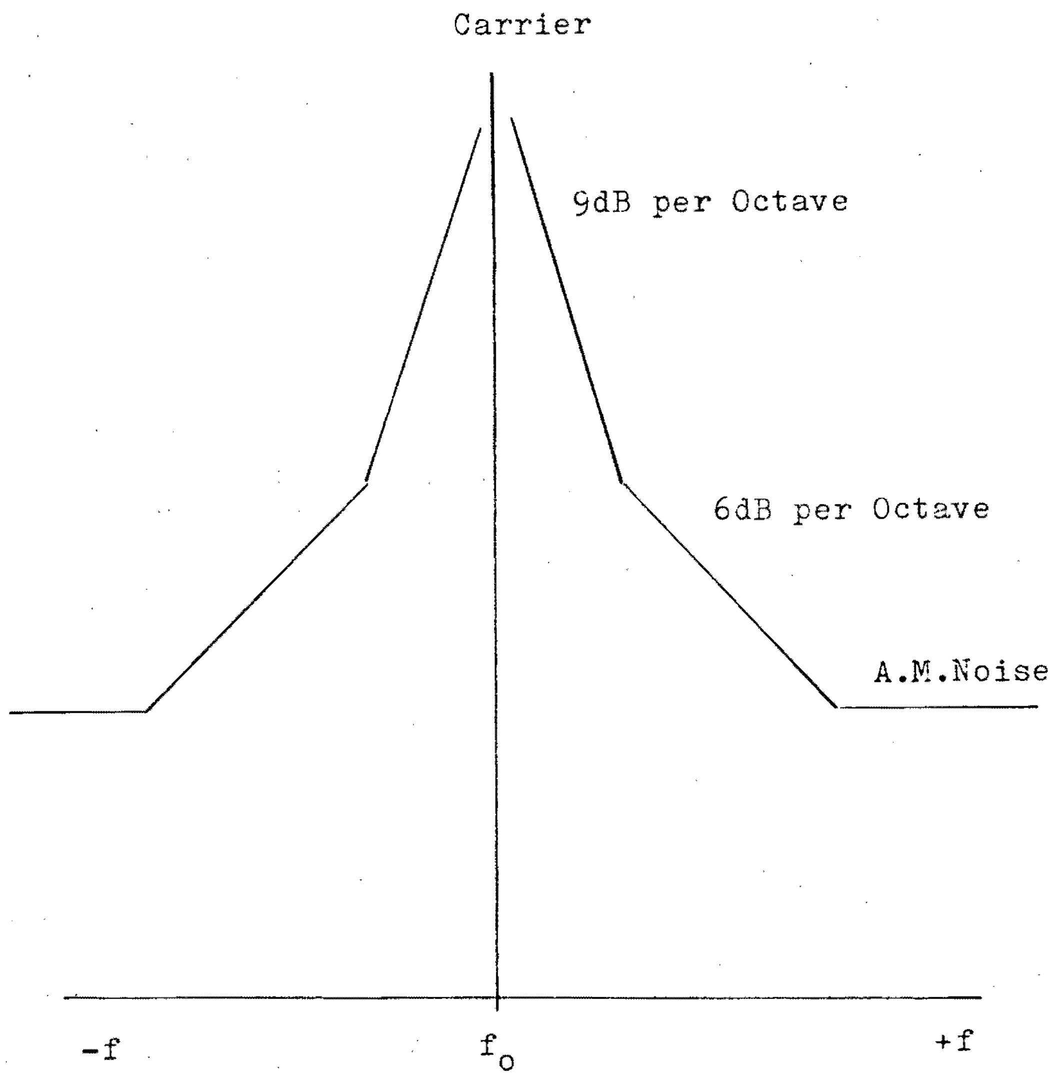

Fig 2.9. FM Noise of the GUNN Source.

Chapter 3

THE DEVICE AS A MIXER.

3.1 Mixer Operation.

In this section an explanation will be given as to how the device operates as a mixer. The approach used is only one of a number of possible approaches that are possible. In the author's opinion this particular approach most readily lends itself to analysis. The main alternative approach will also be discussed but in Chapter 4.

3.1.1 Symbols Used.

L_c = Conversion Loss.
= Power available to IF/Power input.

$L_c(dB)$ = Conversion Loss in dBs.
= $10\text{Log}_{10}(L_c)$

r_p = Positive resistance of the GUNN Diode. ie. In the region before threshold.

r_n = Negative resistance of the GUNN Diode. ie. In the region after threshold.

Note. r_p and r_n are set by the inherent properties of the GaAs semiconductor. See references 20, 21 and 37.

P_p = Reflection Coefficient in the region before threshold.

P_n = Reflection Coefficient in the region after threshold.

$P(t)$ = Time dependant Reflection Coefficient.

P_0, P_1 etc.
= Fourier Coefficients of P(t).

V_b = DC Bias Voltage on the device.

V_1 = Amplitude of the Local Oscillator signal.

V_{th} = Threshold Voltage.

T_c = Conduction Time. This refers to the Fourier Analysis.

Z_o = Charateristic Impedance of the Waveguide.

3.1.2 The Transfer Characteristic.

The characteristic as given in section 2.3.1 can be approximated by two straight lines. This seems to be quite close to the actual situation exept that some hysteresis is apparent when the local oscillator signal is applied. This will not effect the amplitude of the Fourier Coefficients, but will effect their phase. The approximate transfer function together with the local oscillator signal can be drawn as follows as in Fig 3.1.

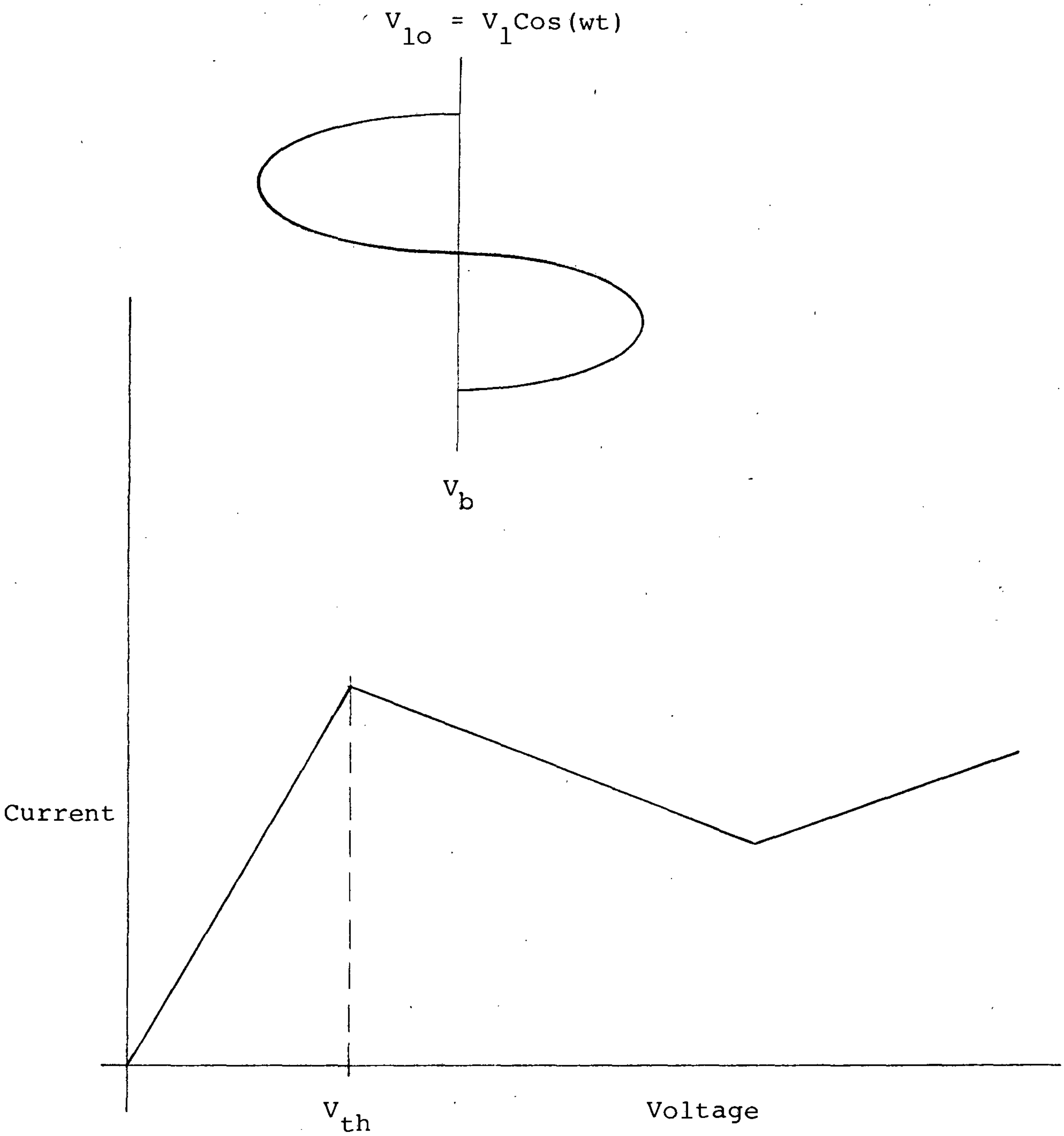

Fig 3.1 The Approximate Transfer Function.

3.1.3 The Reflection Coefficient.

The Local Oscillator signal is due to the oscillatory effect of the device. This voltage gives rise to a time dependant resistance function. If it is now assumed that the device acts as the termination of a waveguide then this also gives rise to a time dependant Reflection Coefficient. The descrete reflection coefficients are as follows:-

$$P_p = \frac{(r_p - Z_0)}{(r_p + Z_0)}$$

$$P_n = \frac{(r_n - Z_0)}{(r_n + Z_0)}$$

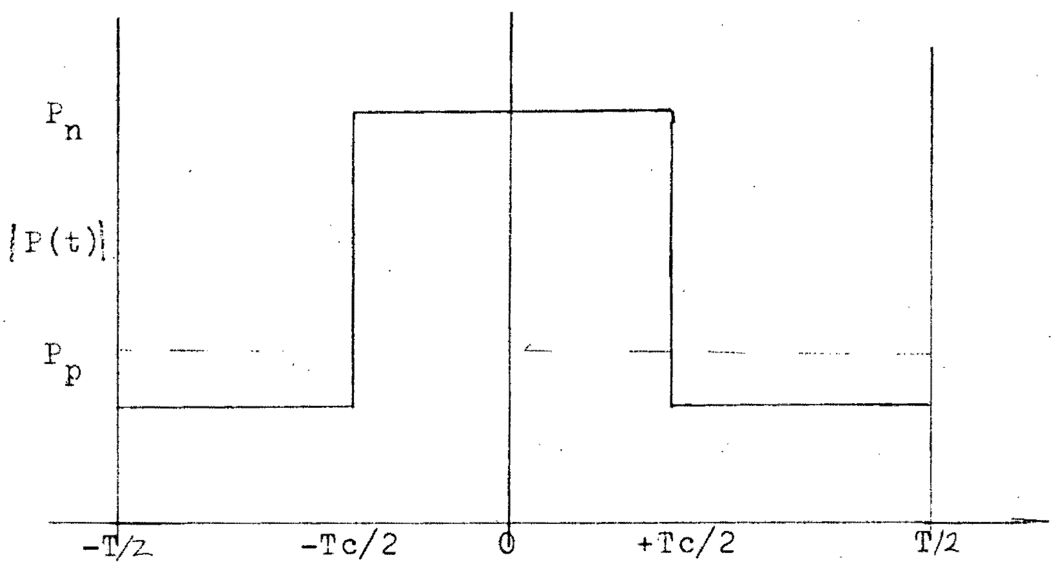

Fig 3.2. Time Dependant Reflection Coeifficient.

Now the ratio T_c/T can be represented as follows:-

The local oscillator voltage reaches threshold when

$V_b - V_{th} = V_1 Cos(PiT_c/T)$

Therefore:

$PiT_c/T = Cos^{-1}((V_b - V_{th})/V_1 = Q$

3.1.4 Sum and Difference components.

The significant reflection coefficient is P_1 and it can be written down as follows:-

$P_1(t) = P_1 Cos(2Pift)$

All other components of P(t) are insignificant as the give rise to signals at microwave frequencies and hence they do not appear at the IF terminals. This also applies to P_0 which is the component that gives rise to the phenomenon of a negative resistance amplifier that is oscillating at a different frequency. operating

If the incomming signal is assumed to be

$V_i = V_s Cos(2wf_s t)$

then the reflected signal is given by

$V_r = V_s Cos(2wf_s t) \times P(t)$

except that the difference signal will not propagate in the waveguide and will be absorbed by the IF terminals. This signal is given by:-

$$2V_{if} = P_1V_s\mathrm{Sin}(2w(f_1 - f_s)t)$$

3.1.5 Conversion Loss.

Conversion Loss is given by:-

$$L_c = P_{if}/P_s$$

Assuming correct matching of the incident and IF signals then:-

$$P_{if} = (P_1V_s)^2/(8R)$$

$$P_s = V_s^2/(2R)$$

$$L_c = P_1^2/4$$

This means that L_c can be given as follows:-

$$L_c = ((P_n - P_p)/w \times \mathrm{Sin}(Q))^2$$

Thus the expression has two distinct components. The component due to conductance of the terminating device and the component due to the conduction angle Q.

3.1.6 Arrival at Noise Figure.

The expression given in section 2.1.2 can be rewritten to give a better form for this section.

$$F = S_i N_o/(S_o N_i)$$

$$F = L_c(N_o/N_i) \text{ - - - - - - - - - - - - - - } (a)$$

$$F = ((P_n - P_p)/w \times Sin(Q))^2 \times (N_o/N_i)$$

Or this can again be rewritten according to the expression in section 2.1.4. Note that the above expression is merely explisive and not analytical.

Now using the noise Temperature definition

$$F = L_c(F_{if} - 1 + t)$$

$$F = L_c t + L_c(F_{if} - 1)$$

$$F = t/G_c + (F_{if} - 1)/G_c$$

Where $L_c t$ is the above expression (a) and $L_c(F_{if} - 1)$ is the second expression of Friis' Formula.

From the expression it can be seen that all terms are a product of L_c. If the IF is ideal then F_{if} is equal to one, and the second term dissapears and we return (a).

Also from section 2.1.3 we see that the equivalent input noise signal can be represented by:-

$$N_i = (F - 1)kTB$$

$$F = N_i/(kTB) + 1$$

To facilitate calculation, N_i can be expressed in terms of Bandwidth so that the term can be reduced to:-

$$F = N_i(B)/(kT) + 1$$

From the expression we see that the Noise Figure of the mixer on its own is independent of the Conversion Loss, however the Noise Figure of the system is very much dependant on Conversion Loss. Using Friis' Formula to collect the terms we get:-

$$F = 1 + N_i(B)/(kT) + (F_{if} - 1)L_c$$

3.1.7 <u>Application to a typical system</u>.

To illustrate the effect of the various parameters on the system performance we can take a few typical values.

Say L_c = 12dB = 16

F_{if} = 3dB = 2

Ni = -170dBc/Hz
= 5E-19watts/Hz

KT = 4.14E-21

This gives:-

F = 147

And

$NF = 22dB$

Note that the value given for N_i is obtained from the curves given in section 2.4.2.

3.1.8 Noise Figure measurement.

This section describes the methods used for measuring the Noise Figures. The methods make use of the expression for Noise Figure as follows:-

$$NF = 10Log(T_e/T_o - 1) - 10Log(N_e/N_o - 1)$$

Where the first part of the expression is called the Excess Noise Ratio and is given by the manufacturer of the noise source, and the second part of the expression N_e/N_o is the ratio of powers measured with the noise source active (N_e) and the inactive (N_o). The expression can be derived very easily as follows:-

$$N_e = KT_oBG_c + (F - 1)KT_oBG_c + K(T_e - T_o)BG_c$$

and

$$N_o = KT_oBG_c + (F - 1)KT_oBG_c$$

This means that N_e/N_o can be reduced to:-

$$N_e/N_o = (FT_o + T_e - T_o)/(FT_o)$$

Solving for F gives:-

$$F = (T_e/T_o - 1)/(N_e/N_o - 1)$$

and hence taking the logs of both sides gives the expression for NF.

If the reader requires further information the he is referred to section 2.1 and to references 40 and 41.

A further note can be made concerning errors due to mismatch of the noise source into the device under test. This will effectively reduce the Excess Noise Ratio. If for example the input Standing Wave Ratio of the DUT is 1.2:1 then the mismatch loss will be 0.83dB and hence the Excess Noise Ratio will be reduced by this much. In practice the sorts of Noise Figures that we are measuring are so high that this mismatch loss will not be significant. eg. On a Noise Figure of 27dB a mismatch loss of 0.83dB will be about 10% in the actual ratios, and this will generally be of the order of the measurement errors at these sorts of power levels.

3.1.9 Procedure for Noise Figure measurement.

Measurements were done using both a Manual technique and an Automatic technique.

The Manual setup was as follows as in Fig 3.3.

The Automatic setup was very similar to the manual one, except that an Automatic Noise Figure Bridge was used. The type is a Saunders No. 5440c. The unit switches the noise source on and of at a set rate and takes the appropriate power levels automaticaly, then calculates the Noise Figure and displays it on a digital display.

Swept frequency measurements were made by changing the Local Oscillator frequencies as follows:-

<u>Case 1</u>. 10MHz to 110MHz

f_{lo} = 55MHz to 155MHz

f_{if} = 45MHz

Low Pass Filter in position.

<u>Case 2</u>. 100MHz to 200MHz

f_{lo} = 55MHz to 155MHz

f_{if} = 45MHz

High Pass Filter in position.

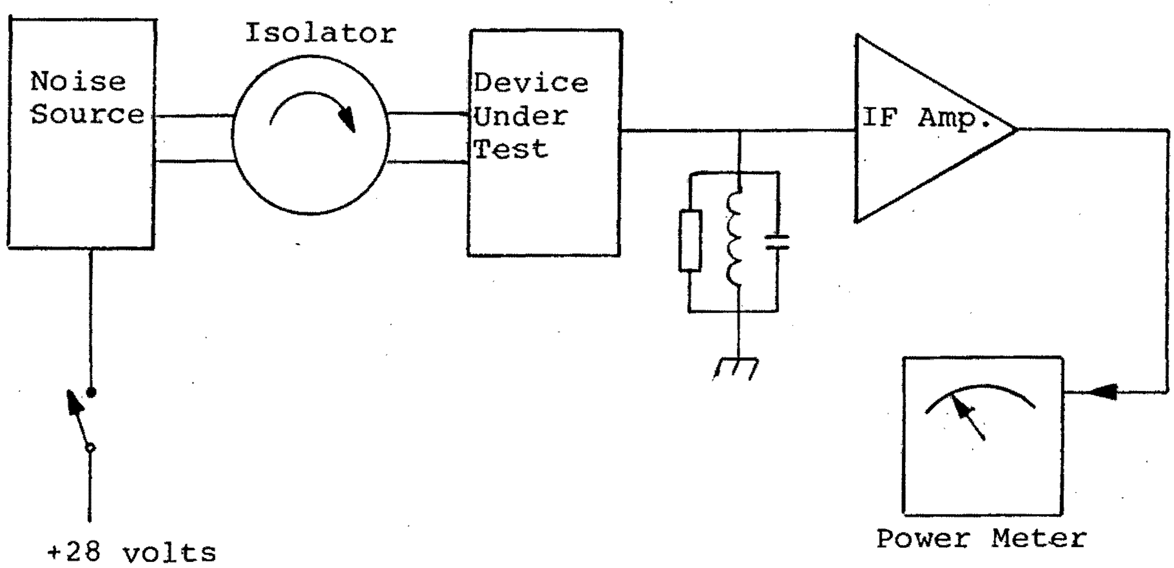

Fig 3.3. Manual Setup.

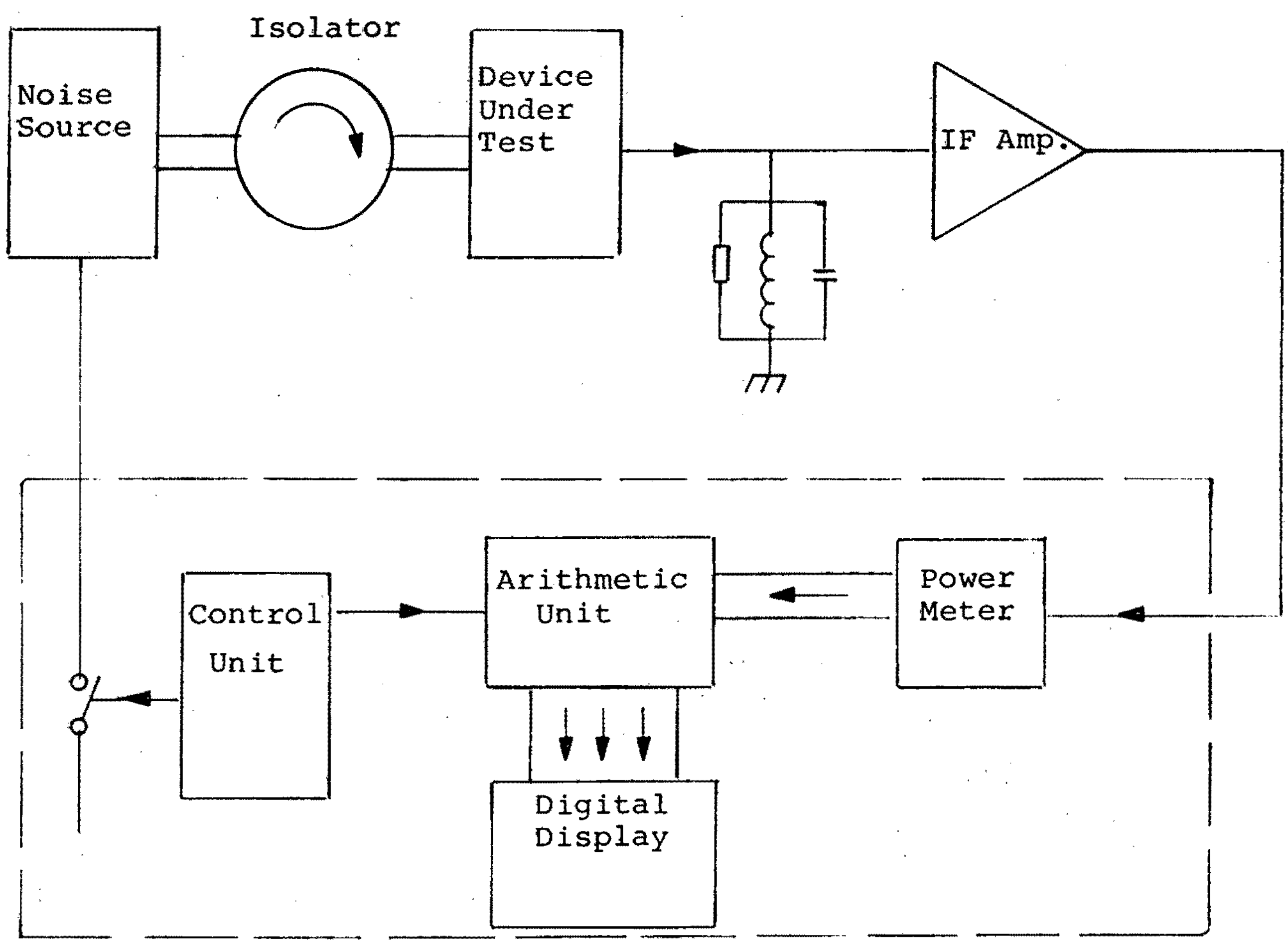

Fig 3.4. Automatic Setup.

3.1.10 Results on 4 Sources.

The graph in Fig 3.5. shows theoretical Noise Figures of 4 different sources plotted over an IF frequency of 10MHz to 200MHz. The actual measured Noise Figures are then shown.

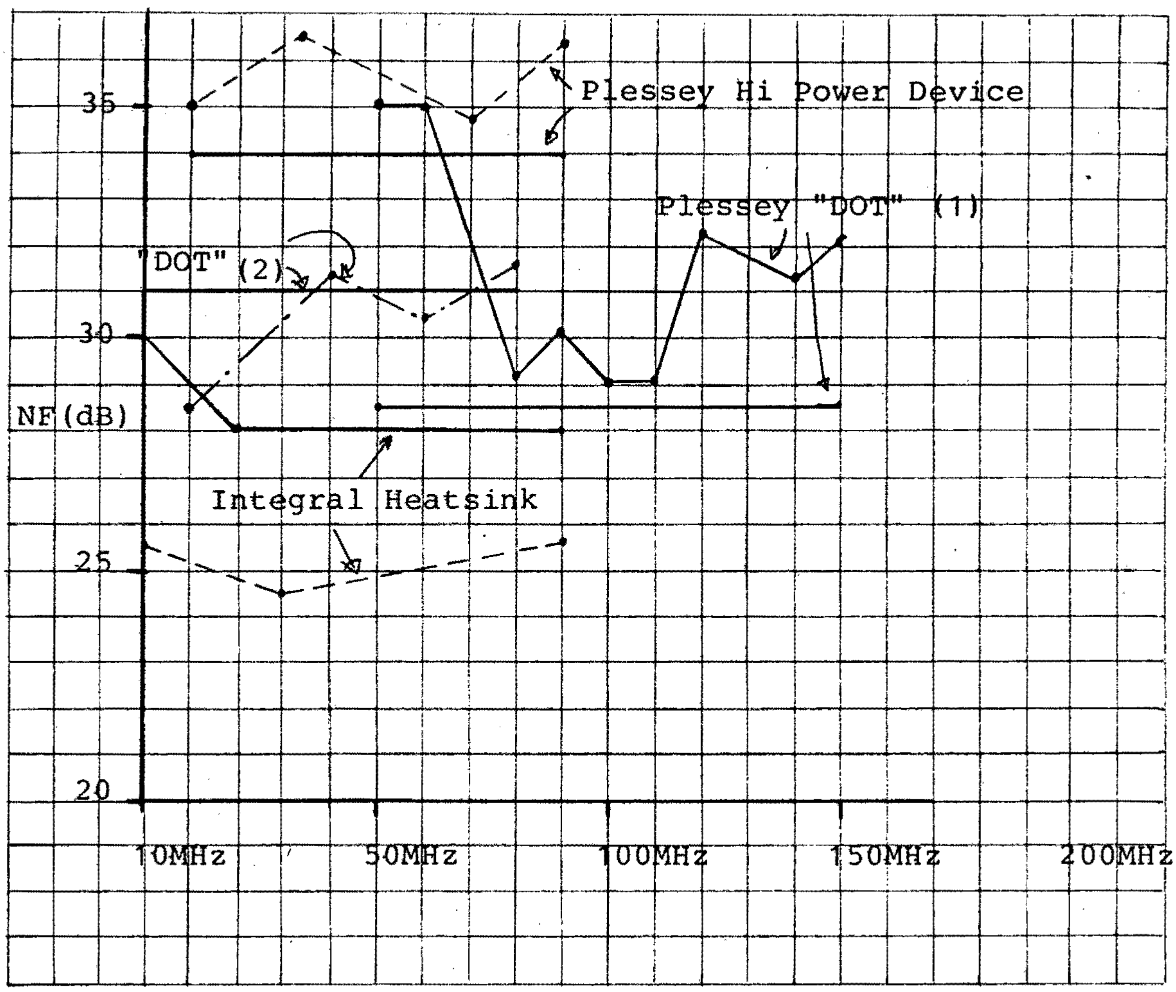

Fig 3.5. Noise Figures against Frequency.

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3.2 Effects of D.C. Bias Voltage.

The bias voltage as related to the threshold voltage affects certain parameters of the device. The parameters it affects are Noise Generation, Conversion Loss and Power Output. All these parameters affect the Noise Figure. This section will show these effects.

3.2.1 Effect on Noise Generation.

Matsuno in reference 25 shows that the Noise Power Spectrum is much dependant on the RMS amplitude of the fundamental oscillations. The RMS amplitude gives the power output of the source and this is much dependant on the D.C. bias voltage.

Matsuno shows that the Noise Power increase at any particular frequency shift from the carrier is largely proportional to the increase in the RMS Amplitude of the oscillations. This in turn is proportional to D.C. Bias and hence also to power output. This leads to the conclusion that Noise to Carrier Ratio remains substantialy constant with bias voltage. On the other hand Matsuno showes that current fluctuations and hence upconverted F.M. Noise increased sharply as the applied electric field approaches the threshold field.

This means that the Noise to Carrier ratio is not much dependant on the Bias Voltage unless the Bias Voltage is brought close to the Threshold Voltage.

3.2.2 Affect on Conversion Loss.

The expression for Conversion Loss as derived in section 3.1.5 can be written as follows:-

$$L_c = ((P_n - P_p)/w \times \mathrm{Sin}(Q))^2$$

Where Q can be written as:-

$$Q = \mathrm{Cos}^{-1}((V_b - V_{th})/V_1)$$

From the expression it seems that $\mathrm{Sin}^2(Q)$ is at a maximum when $V_b = V_{th}$. In practice this would appear to be correct as this would indicate a Time Dependant Reflection Coefficient with a mark to space ratio of 1:1. In reality there are some difficulties in achieving this. On the one hand the amplitude of V_1 is somewhat dependant on V_b and decreases as V_b decreases. On the other hand it is very difficult to operate a GUNN Source with a low Bias Voltage as the bias circuit becomes unstable.

The expression for Q can be re-written to aid analysis as follow:-

$$Q = \mathrm{Cos}^{-1}((V_b/V_{th} - 1)/V_1/V_{th})$$

The expression can be plotted on a set of axis to show the effect of the various parameters. Say $L_c' = \mathrm{Sin}^2(Q)$ then L_c' can be plotted as as in Fig 3.6.

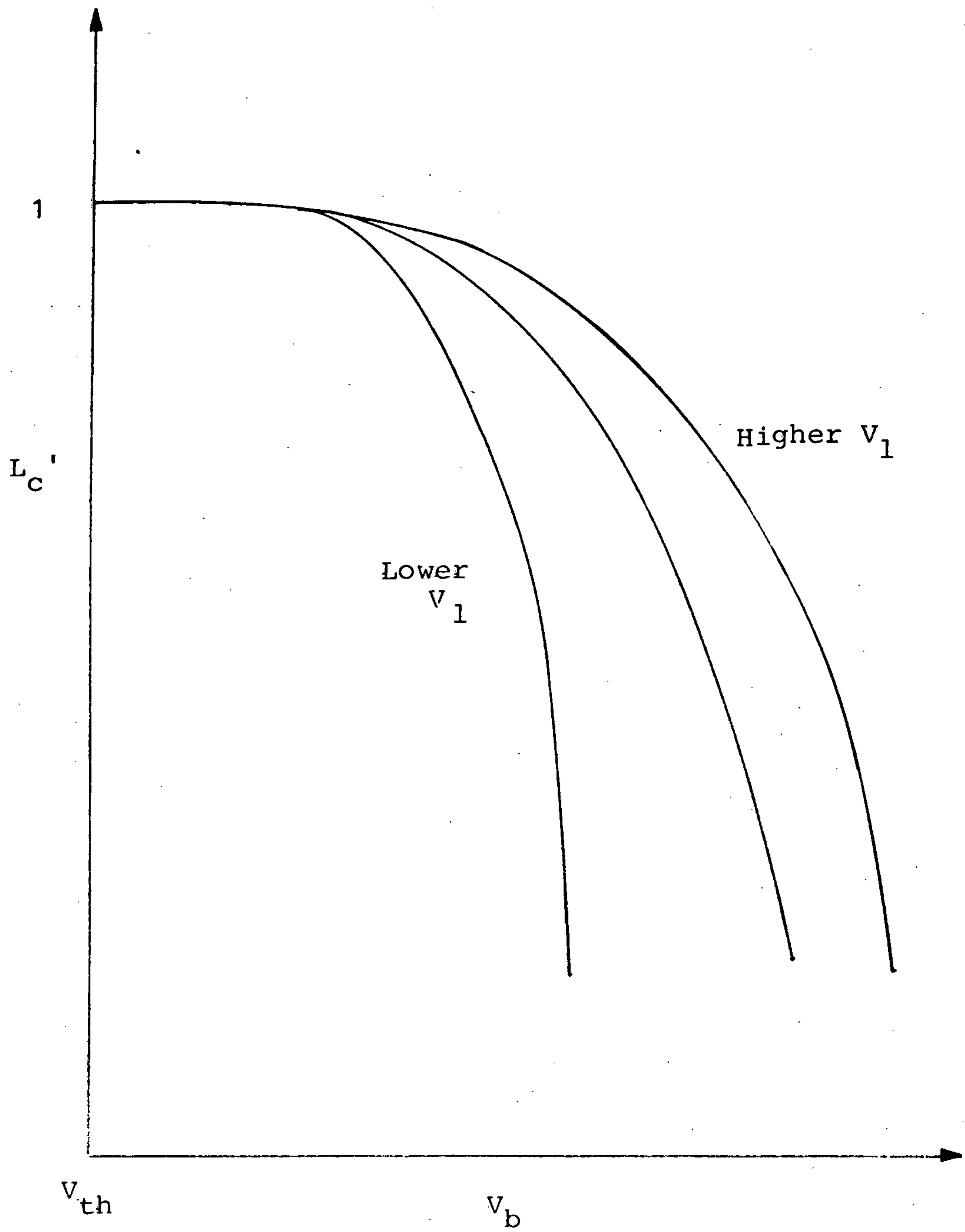

Fig 3.6.

To illustrate the theory a number of sources have been measured for Conversion Loss against Bias Voltage, and the results plotted as follows:-

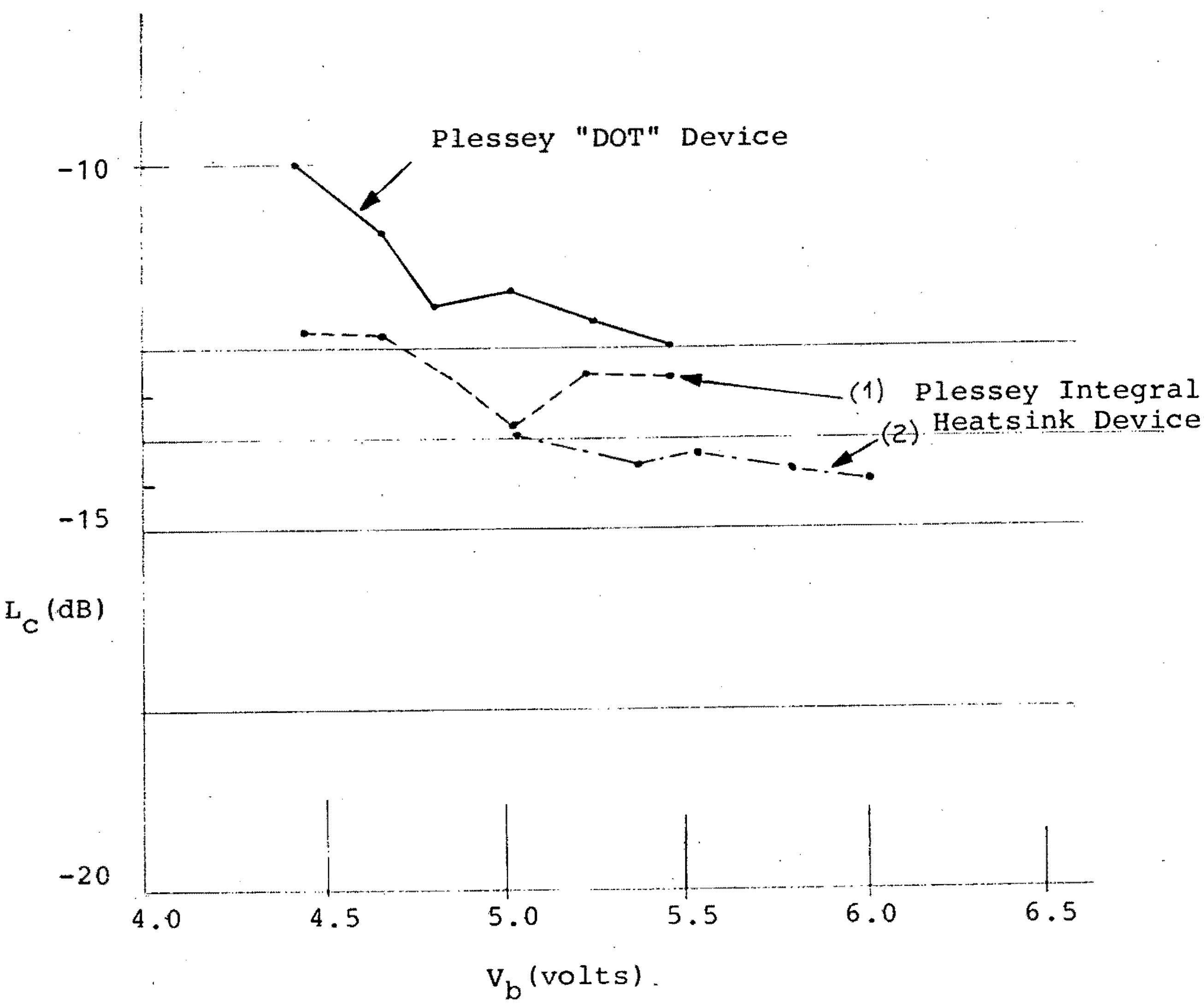

Fig 3.7.

3.2.3 Effect of Bias Voltage on Power Output.

Bias voltage effects power output and hence it a effects the amplitude of the Local Oscillator Voltage. This in turn a effects the Conversion Loss which has an effect on the Noise Figure. The main influence on Noise Figure is however caused by the change in the amplitude of the noise input due to the Local Oscillator. If it is assumed that the Noise to Carrier ratio remains essentually c constant with bias voltage, then the amplitude of the added noise will have the same characteristic with respect to bias voltage as does the power output. This in turn will affect the Noise Figure almost directly as shown in section 3.1.6 equation (c). The contribution to Noise Figure due to added noise is far more significant than that due to the Noise Figure of the IF in combination with the Conversion Loss.

The diagram in the previous section shows the effect of bias voltage on Conversion Loss and hence on Noise Figure. As another illustration of this effect a number of sources with widely ranging output powers are plotted for Noise Figure. All are Q Band sources and all are operated at recommended bias voltages. (Fig 3.8.)

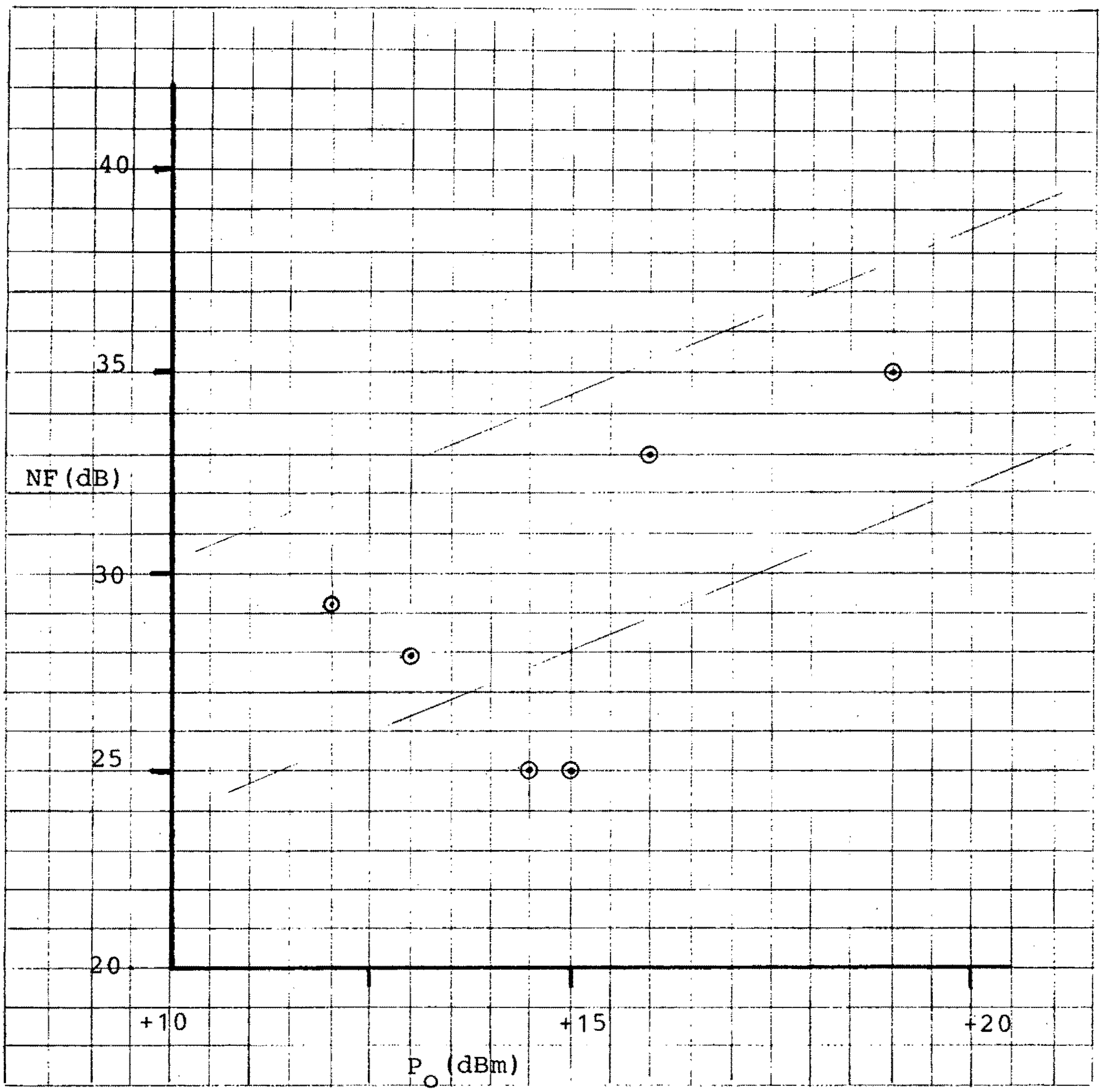

Fig 3.8.

3.2.4 Conclusion.

It seems that Noise Figure is much dependant on Bias Voltage but not because of change in Conversion Loss. It is affected by the amplitude of added noise at the I.F. frequency of interest.

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3.3 <u>Effect of Rp and Rn on Conversion Loss.</u>

The intrinsic factor $(P_n - P_p)^2$ plays a very important part in the Conversion Loss of the device.

$$P_p = \frac{R_p - Z_o}{R_p + Z_o}$$

$$P_n = \frac{R_n - Z_o}{R_n + Z_o}$$

To find the optimum values of R_p and R_n for the best Conversion Loss, we could differentiate the term w.r.t the variable of interest.

P^2 = Loss Term.

$$= \left(\frac{R_p - Z_o}{R_p + Z_o}\right)^2 - 2\left(\frac{R_p - Z_o}{R_p + Z_o}\right)\left(\frac{R_n - Z_o}{R_n + Z_o}\right) + \left(\frac{R_n - Z_o}{R_n + Z_o}\right)^2$$

However it is probably easier to simply observe the expression by inspection as the gain term varies with its variable.

<u>First as regards Rp.</u>

This term is usually not very variable and in any case it only provides the loss component of the conversion loss, whereas R_n can also provide a gain term. Generally R_p will always be of the order of +2 to +4 ohms, and hence P_p will always range between -1 and +1 depending on Z_o. However Z_o can generally be assumed to be of the order of hundreds of ohms - positive, and hence P_p will always be close to -1.

Secondly as regards Rn

This term can range quite widely in magnitude but will always be of the order of -100 to -200 ohms. This means that P_n will always be of the magnitude of say -1.5 to -2.6.

We can now re-inspect the gain expression in terms of its three parts. The first term $(P_n)^2$ is the gain term and will always be positive. The third term is a loss term and will always be less than one and will always be positive.

The second term may represent gain or loss but will generally be negative and will subtract from the sum of the other two terms.

Case 1.

$P_p = -0.99$

$P_n = -2.0$ ($R_n = -150$ $Z_o = 450$)

$P^2 = 1.020$ (= 0dB)

Case 2.

$P_p = -0.99$

$P_n = -1.57$ ($R_n = -100$ $Z_o = 450$)

$P^2 = 0.33$ (= - 4.8dB)

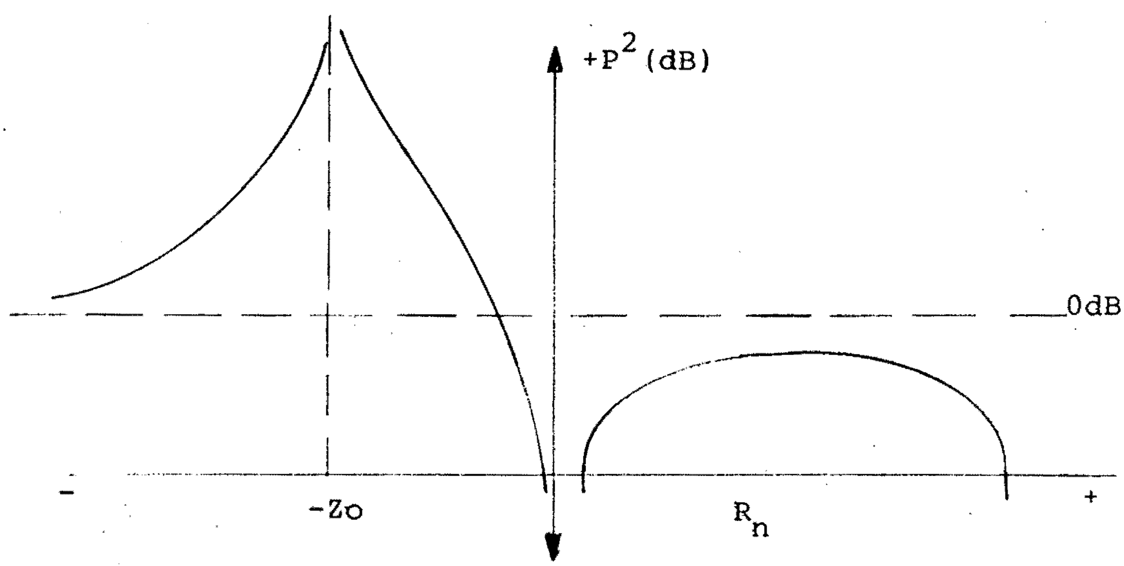
Case 3.

$P_p = -0.99$

$P_n = -2.6 \quad (R_n = -200 \quad Z_o = 450)$

$P^2 = 2.6 \quad (= 4.1\text{dB})$

This indicates a practical range of values for P^2 (dB) of about 8. In theory we should be able to obtain infinite gain when $R_n = -Z_o$. The curve in the following diagram indicates the calculated range of P^2(dB) against a range of R_n.



3.3.2 Practical significance of Conversion Loss

If we return to the expression for the noise figure of the system we may see that for the SMO scheme, converion loss does not play an important part in the overall system performance.

Overall noise figure is given by:-

$$F = F_{SMO} + L_{SMO}F_{IF}$$

In general the significance of the various terms will be as follows:-

$F_{SMO} = 316$ (25dB)

$L_{SMO} = 10$ (10dB - say worst case)

$F_{IF} = 2$ (3dB - easily obtainable at IF frequencies of between 10 and 200MHz)

This means that the overall noise figure will be made up as follows:-

$$F = 316 + 10\text{x}2 = 336.$$

The most significant part by order of magnitude is thus the SMO noise figure.

The conversion loss can always be catered for by increase gain in the IF amplifier.

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3.4 Arrival at Noise Figure from Conversion Loss and added Noise.

In this section we will try to show the form of the noise skirts of a Gunn Source and how they give rise to the Noise Figure.

3.4.1 Analysis of F.M. Noise of GUNN V.C.O.

Generally the instantaneas expression for an F.M. modulated wave can be given as follows:-

$$e(t) = E_c \mathrm{Cos}(w_c t + \int B \mathrm{Sin}(w_m t))$$

Where E_c = Carrier Amplitude.

w_c = Carrier Frequency.

w_m = Modulation Frequency

B = Modulation Index

$= f_{peak}/f_m = w_{peak}/w_m$

Expanding the expression and setting E_c to 1 gives:-

$$e(t) = \mathrm{Cos}(w_c t).\mathrm{Cos}[B\mathrm{Sin}(w_m t)] - \mathrm{Sin}(w_c t).\mathrm{Sin}[B\mathrm{Sin}(w_m t)]$$

Now for N.B.F.M. , that is with a modulation index of much less than 1.5 we can approximate the following:-

$$\mathrm{Cos}[B\mathrm{Sin}(w_m t)] = 1$$

$$\mathrm{Sin}[B\mathrm{Sin}(w_m t)] = B\mathrm{Sin}(w_m t))$$

The F.M. Wave now becomes:-

$$e(t) = \mathrm{Cos}(w_c t) + (B/2)\mathrm{Cos}(w_c - w_m)t + (B/2)\mathrm{Cos}(w_c + w_m)t$$

The expression is very similar to that for a Phase Modulated wave, except that B is replaced by Q/2, where Q is Phase deviation.

A general way of specifying the noise level of an oscillator is to use the concept of the Single Sideband to Carrier Ratio. This is defined as:-

$$\frac{SSB}{Carrier}(dB) = 10\mathrm{Log}\,\frac{SSB\ Power}{Carrier\ Power} = \frac{N}{C}$$

Given that the NBFM approximations hold then the expression ca be written as:-

$$\frac{SSB}{Carrier}(dB) = 20\mathrm{Log}\left(\frac{B}{2}\right)$$

or:-

$$\frac{SSB}{Carrier}(dB) = 20\mathrm{Log}\left(\frac{f_{pk}}{2f_m}\right)$$

Now given that $f_{pk} = 1.4\,\Delta f_{rms}$ the expression becomes:-

$$\frac{SSB}{Carrier}(dB) = 20\text{Log}\left(\frac{\Delta f_{rms}}{1.4f_m}\right)$$

Now in terms of phase deviation this is:-

$$\frac{SSB}{Carrier}(dB) = 20\text{Log}\left(\frac{Q}{2}\right)$$

The above expressions apply to a single tone FM modulation and its resultant sidebands. In general the phase noise spectrum of a GUNN oscillator looks something like Fig 3.9.

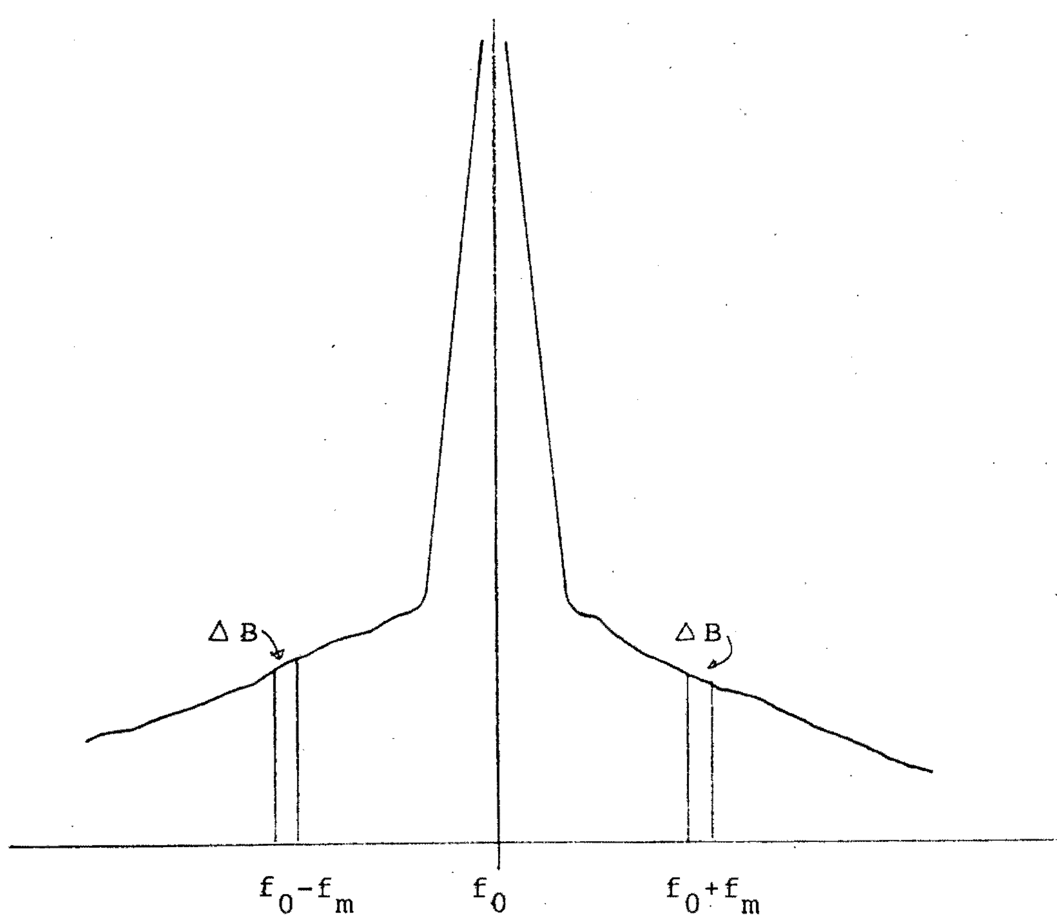


Fig 3.9. Phase Noise Spectrum of typical GUNN.

The noise spectrum on each side of the carrier is continuous, but it is possible to devide it into strips of width B and then to generate a new function as follows:-

If we first assume that the ammplitude of the noise is constant over the width of the strip and the that the strip is placed at f_m removed from the carrier then we may define a noise density function as follows:-

$$D(f_m) = \frac{\text{Power Density(One Sideband, Phase only)}}{\text{Power(Total Signal)}}$$

The dimensions are Watts/Hz/Watts.

Using the single tone analogy we can see that:-

$$D(f_m) = 20\text{Log}\left(\frac{\Omega}{2}\right) \quad \text{dB/Hz}$$

$$D(f_m) = 20\text{Log}\left(\frac{\Delta f_{pk}}{2f_m}\right) \text{dB/Hz}$$

If $D(f_m)$ is measured and now plotted it will be plotted against a measuring bandwidth (BW). e.g. A spectrum analyser will have an input filter set to a certain bandwidth. To convert this spectrum to $D(f_m)$ in 1Hz of bandwidth we can use an expression as follows:-

$$D(f_m)_1 = \frac{\text{SSB Noise Power}}{\text{Signal Power}}(\text{in BW}) - 10\text{Log(BW)} \quad \text{dB/Hz}$$

We can now define the spectral function from the density function as $S(f_m)_{BW}$. The Phase Noise spectrum of the GUNN Oscillator will generally appear to be as in Fig 3.10.

Hobson in reference 15 defines an expression for the power spectral density of a GUNN Oscillator as follows:-

$$S(f_m)_{BW} = \frac{b.BW}{6.28\left[\left(\frac{b}{2}\right)^2 + f_m^{\ v}\right]}$$

Where v is between 2 and 3. ie 2 indicates 6dB/Octave and 3 indicates 9dB/Octave. The term b is given as:-

$$b = \frac{0.94KT}{A\,\varepsilon_o \varepsilon_r E_t^2} \times \frac{fo}{Q}$$

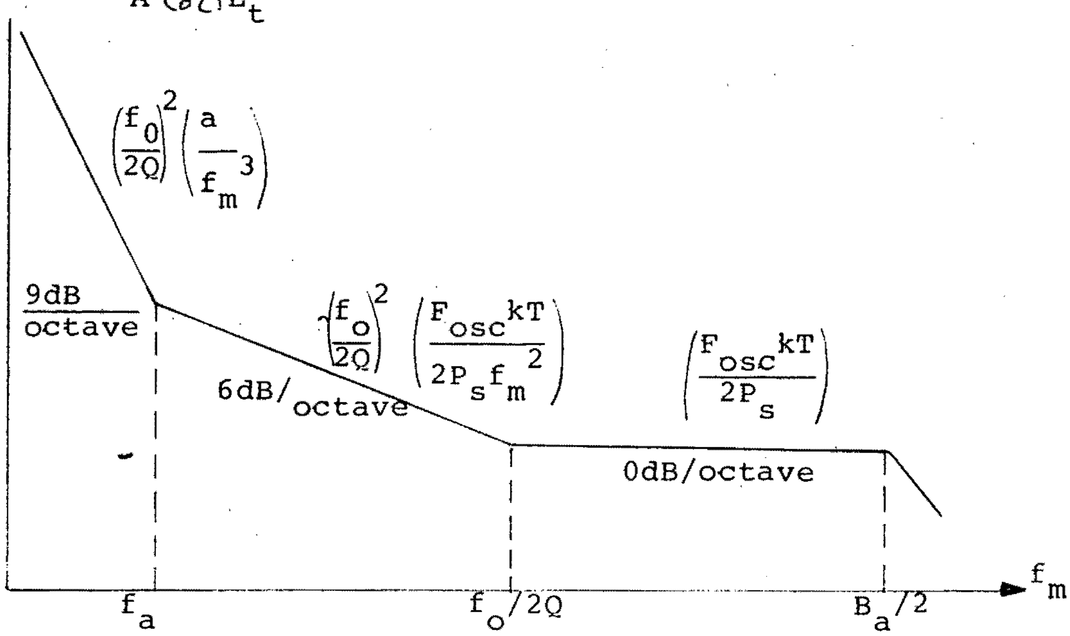


Fig 3.10. Phase Noise Spectrum of GUNN Osc.

Hobson calculates the term appropriate to 1 ohm per cm GaAs as follows:-

$$b = 5.3E-7(f_o/Q)$$

For a typical Q band device with a cavity Q of say 100, b will calculate to:-

$$b = 185.5$$

Now for f_m greater than 1000Hz the term f_m^v becomes much more significant than $(b/2)^2$, and the expression can be written as:-

$$S(f_m)_{BW} = \frac{b.BW}{6.28f_m^v}$$

We can thus derive an expression for the "a" indicated in the previous diagram and repeated here

Slope in the Flicker Noise region is given by:-

(Given that BW = 1)

$$\text{Slope(Flicker)} = \left(\frac{f_o}{2Q}\right)^2 \frac{a}{f_m^3}$$

$$\frac{b}{6.28f_m^v} = \left(\frac{f_o}{2Q}\right)^2 \frac{a}{f_m^v}$$

We can solve for a as follows:-

$a = 3.37E\text{-}7(Q/f_o)$

Now for an X band device we can calculate a value for f_a.

f_a occurs when:-

$$\frac{a}{f_m^3} = \frac{F_{osc}KT}{2P_s f_m^2}$$

This means that f_a is given as:-

$$f_a = \frac{2aP_s}{F_{osc}KT}$$

For $Q = 100$

$f_0 = 10E9Hz$

$P_s = 50E\text{-}3Watts$

$F_{osc} = 10$

This gives f_a as

$f_a = 8.4KHz$

Manassawitch in his book on Frequency Synthesis~~o~~ers predicts an f_a of 300KHz for an X band GUNN source. On the one hand this would suggest a v of about 2.5 or alternatively a cavity with a Q much higher than the 100 we used in our calculation.

If we now apply the expression to a Q band device we may obtain a new value for b. We will use a value of "a" proportionately smaller because of the higher frequency.

$b = 1.8E-6(f_o/Q)$

$a = 5.3E-15$

$f_a = 13.25KHz$

If we now assume that for frequencies above f_a the N/C ratio increases by 6dB per octave then we may calculate the N/C ratio at any particular frequency.

At about 12KHz Hobson suggests a N/C ratio of -60dB in a bandwidth of 70Hz. This means that in 1Hz of bandwidth the N/C ratio is -78dB.

Tabulating by octave we get:-

12Khz	----	-78dB
24Khz	----	-84dB
48Khz	----	-90dB
96KHz	----	-96dB
192KHz	----	-102dB
384KHz	----	-108dB
768KHz	----	-114dB
1536KHz	----	-120dB
3072KHz	----	-126dB
6144KHz	----	-132dB
12288KHz	----	-138dB
24576KHz	----	-144dB
49152KHz	----	-150dB
98304KHZ	----	-156dB
196608KHz	----	-162dB

We also know that the slope becomes 0dB/Octave at a frequency given by:-

$$f_o = QB_{osc}$$

$$= 175\text{MHz}$$

This suggests that the Noise Floor is reached at about -160dBc.

There does not seem to be much gain by going to even higher values of f_m. On the other hand an improvement of cavity Q reflects directly on the Noise to Carrier ratio. e.g. If we had assumed a cavity Q of 1000 then then the floor noise level would have been at -180dBc, and at 17.5MHz. Using the expression for floor noise level this would have calculated to -183dBc.

3.4.2 Noise Contribution due to the Reactance Tuning of the cavity.

We can assume that this contribution is due to thermal excitation in the equivalent resistance of the varactor.

Say $R_{eq} = 150E3$

$K = 25\text{MHz/V}$

$$f_{rms} = K\sqrt{4kTBR}$$

But B = 1

$\Delta f_{rms} = 1.22$

$N/C = 20Log(\Delta f_{rms}/(2f_m))$

At say f_m = 80MHz:-

N/C = -162dBc

This is quite close to the floor level found before and hence it can be ignored.

3.4.3 Noise Figure from Added Noise.

Using the expression previously derived for Noise Figure we can obtain a result by applying the floor value for Added Noise.

$F = 1 + N_i/KT$

Say

$N_i = P_s N/C$

$= 100mW \times (-160dBc)$ W/Hz

= 1E-17 W/Hz

KT = 4.14E-21 W/Hz

This means that F calculates to:-

F = 2416

NF = 33.8dB

Fig 3.10 gives a set of typical devices as plotted.

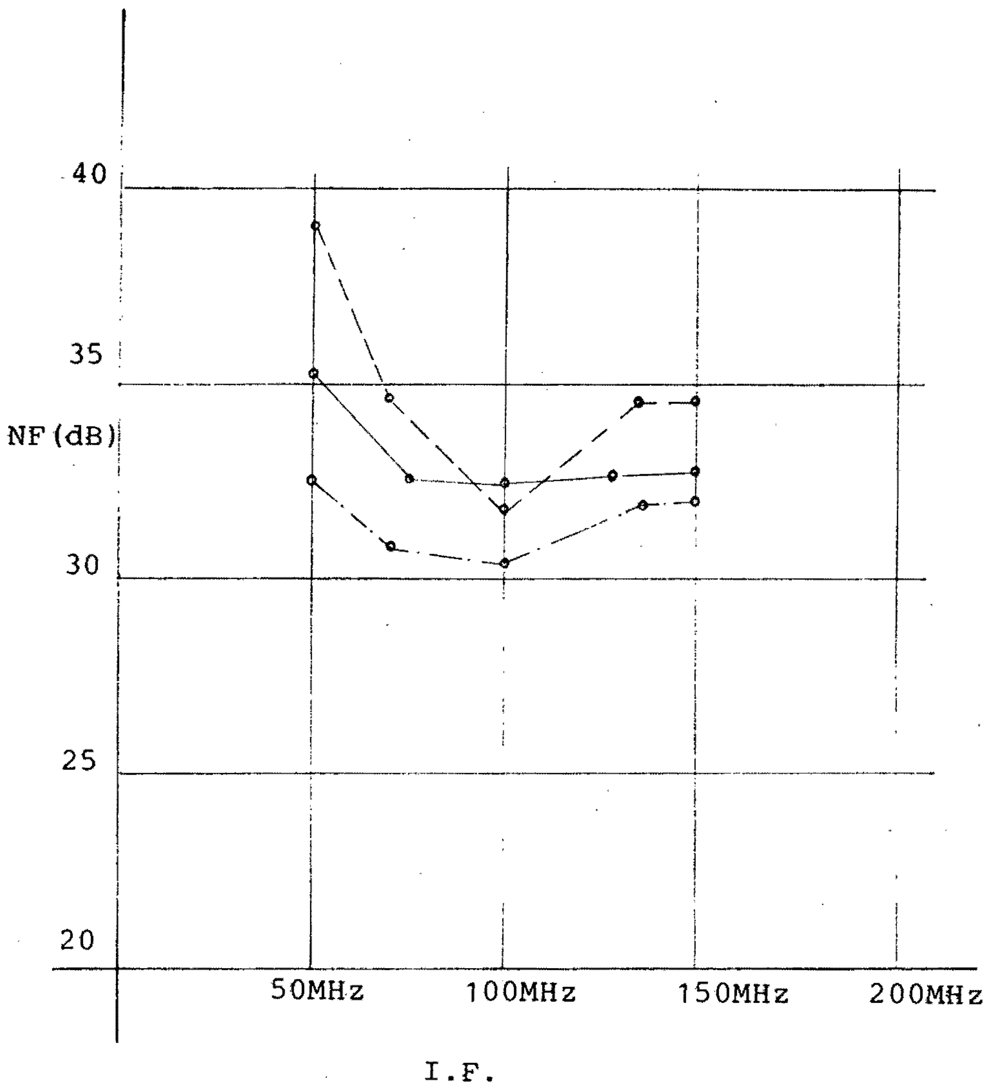


Fig 3.10.

3.5 Effect of Matching Impedance at IF on Noise Figure and System Performance

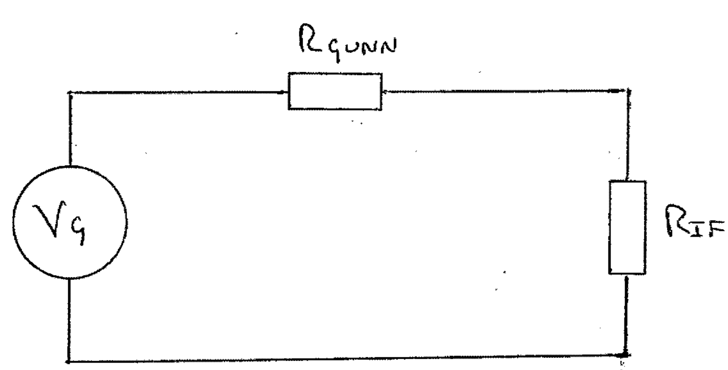
It is convenient to consider the effect of matching impedance from two different aspects. The first aspect being the way in which it affects the intrinsic noise figure of the divice, and the second aspect being the way it affects system noise figure.

3.5.1 Effect on the device Parameters

The two device parameters considered are Noise Figure and Conversion Loss.

Noise Figure appears to be substantially unaffected by matching impedance. This matching impedance seems to play very little part in the added noise and hence very little part in the Noise Figure. On the other hand it plays a very large part in the Conversion Loss of the device.

If we assume an equivalent circuit as follows:-



Where R_{GUNN} is the output resistance to the IF of the GUNN terminals and R_{IF} is the loading resistance of the IF.

$$P_{tot} = P_l + P_{IF}$$

$$= i^2 R_{GUNN} + i^2 R_{IF}$$

$$\frac{P_l}{P_{IF}} = \frac{R_{GUNN}}{R_{IF}}$$

$$P_l = \frac{P_{IF} R_{GUNN}}{R_{IF}}$$

$$P_{tot} = P_{IF} \quad \frac{R_{GUNN}}{R_{IF}} + 1$$

$$\frac{P_{IF}}{P_{tot}} = \frac{R_{IF}}{R_{GUNN} + R_{IF}}$$

This expression is a maximum when $R_{IF} = R_{GUNN}$

The conversion loss of the device can now be written as follows:-

$$L'_c = L_c + 10\text{Log}\,\frac{R_{IF}}{R_{GUNN} + R_{IF}}$$

Where L_c is the intrinsic loss of the device in dB.

This means that the conversion loss will in any case always have a minimum of 3dB. Unless the time dependant reflection coefficient becomes such that its gain is greater than 3dB.

3.5.2 Effect on System Performance

The overall system Noise Figure is given by:-

$$F = F_{GUNN} + (F_{IF} - 1)L_C$$

This means that the added conversion loss adds directly to the Noise Figure, but is scaled by the factor $(F_{IF} - 1)$. If F_{IF} tends to 1 then this factor dissapears.

The practical situation shows the significance of the various terms.

If the designer were requiring optimum range performance from his system then he would choose an IF amplifier with the best possible Noise Figure and match it to the GUNN for optimum Noise Figure rather than for optimum gain.

Taking two cases with NF_{GUNN} = 25dB.

Case 1.

NF_{IF} = 2.5dB

L'c = 11dB

Therefore

$F = 316 + (1.778 - 1)12.59$

$F = 326$ $NF = 25.1dB$

<u>Case 2.</u>

$NF_{IF} = 6.0dB$

$L'c = 7dB$

Therefore

$F = 316 + (3.98 - 1)5.0$

$F = 330.9$ $NF = 25.2dB$

This shows the very marginal effect that the overall conversion loss L'c has on the system Noise Figure. Suffice it to say that this should generally be kept at atleast 10dB below the Noise Figure of the GUNN.

In general there are so many other systematic uncertainties that the matching impedance of the IF plays only a marginal part in the system performance.

3.5.3 Measured Results

A GUNN source was set up with an IF amplifier turned to 73MHz with a bandpass of about 3MHz. The GUNN device was then loaded with three different resistors. 50ohms, 100ohms and 500ohms. Change in Noise Figure was unmeasureable and was about 27.5dB with this particular device.

Note that the bias voltage was kept constant at the nominal rated value. On the other hand if the bias voltage were changed then wild variations are found in impedance presented to the matching circuit and oscillations are possible. This gives rise to what seemed to be oscillations at the IF frequency which cannot be distinguished from received IF power and could thus lead to a quite erroneous Noise Figure measurements.

For interest a set of curves is included showing the change of GUNN impedance to the IF over a range of bias voltages and frequencies. These results were measured according to the technique given in section 3.6.2.

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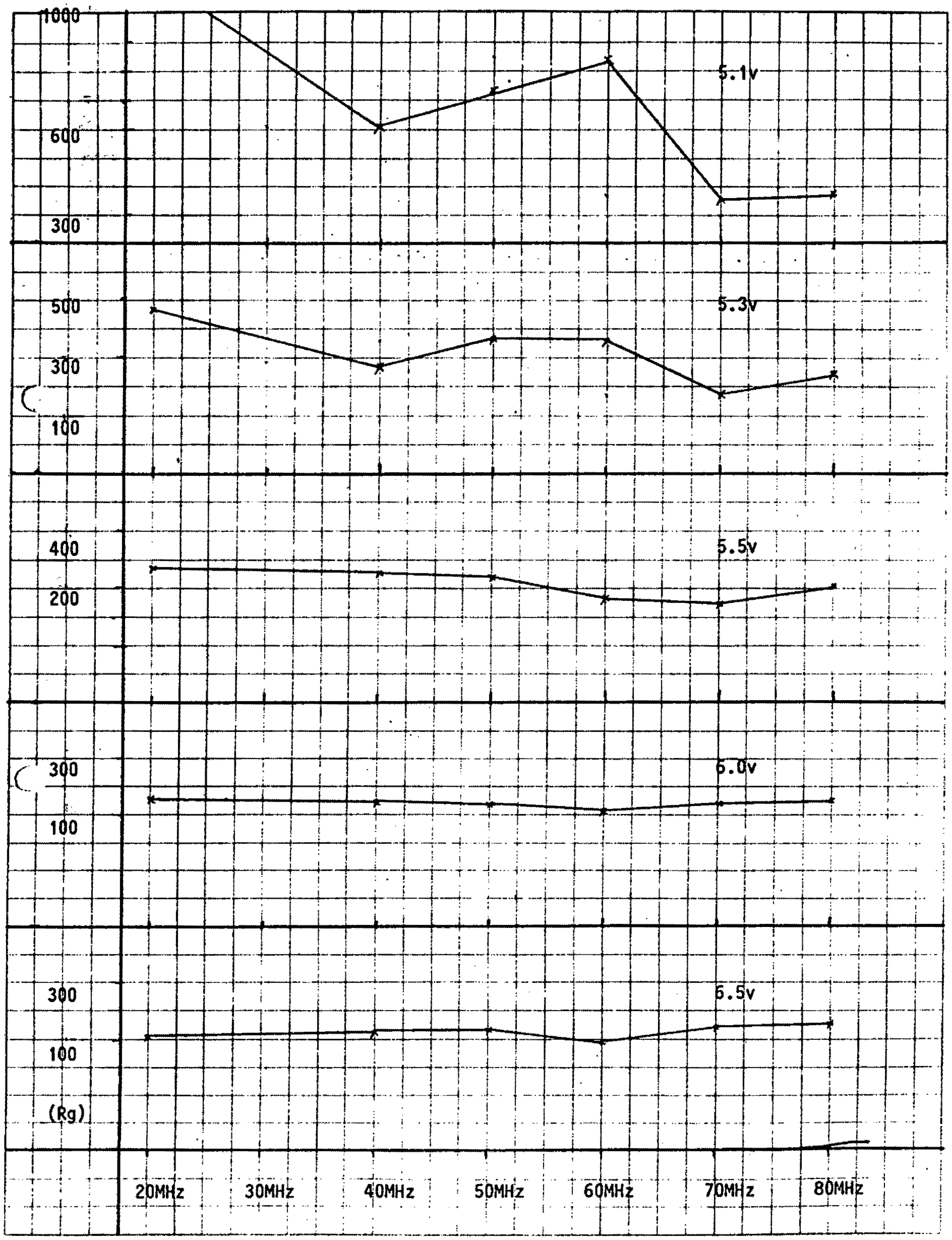
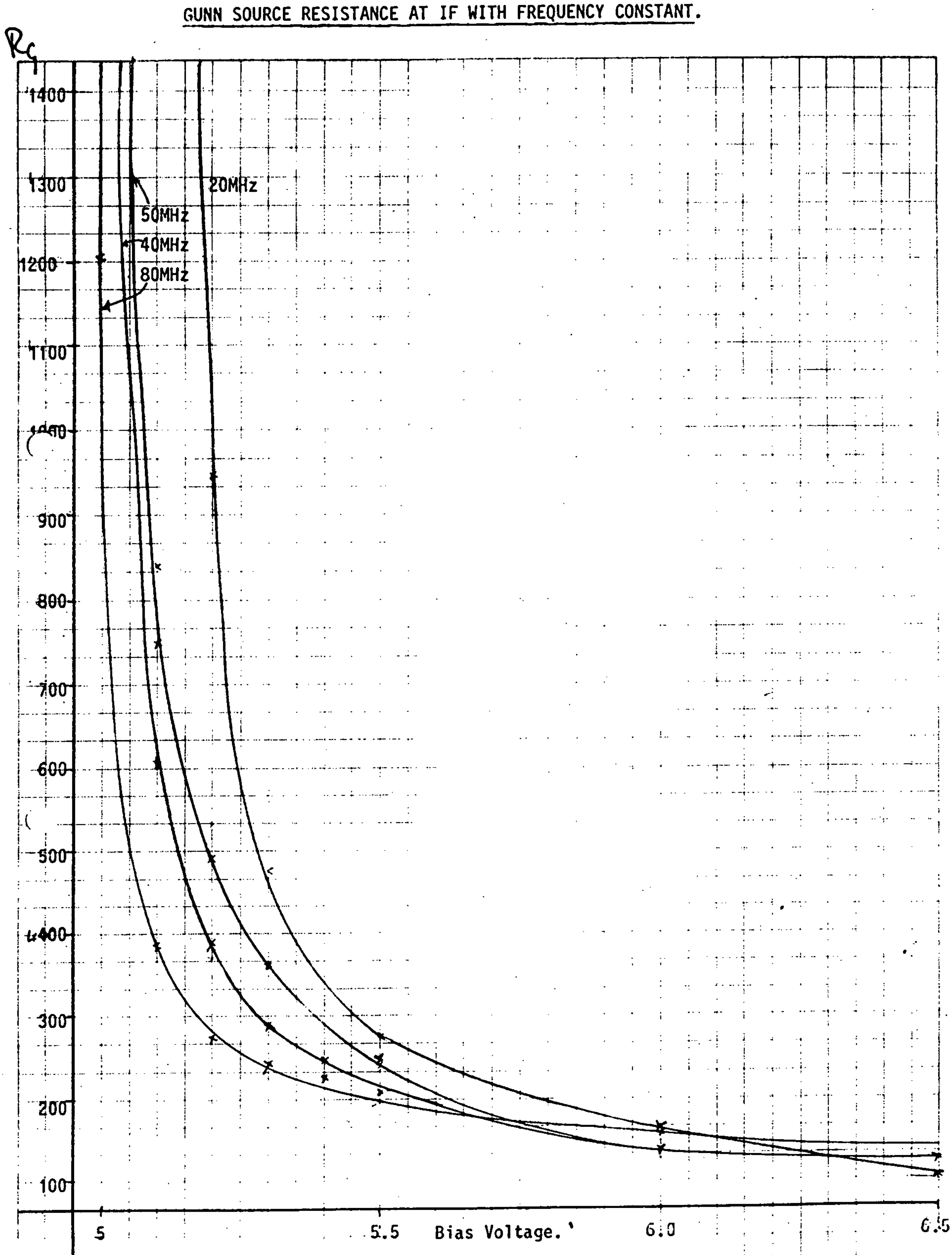
GUNN SOURCE RESISTENCE AT IF WITH VOLTAGE CONSTANT.

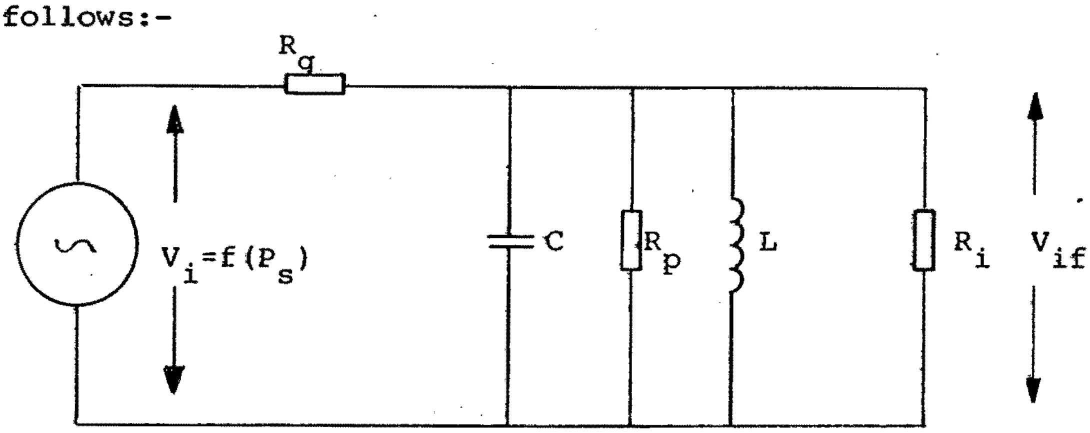


GUNN SOURCE RESISTANCE AT IF WITH FREQUENCY CONSTANT.



3.6 Bias Circuit Stability.

The IF output of the source can be represented as a Voltage Source (Thevenin) in series with a resistance where the voltage is a Function of the input power. The equivalent circuit would be as follows:-



R_g = Source Impedance of the GUNN.

R_p = Parallel resistance of the inductor.

R_i = Input resistance of the IF Amplifier.

3.6.1 Stability.

The interface is a Second Order resonant circiut. If it is first assumed that the Q of the inductor is very high then the total parallel resistance becomes:-

$$R_t = R_i R_g/(R_i + R_g)$$

The total parallel impedance can now be written as:-

$$Z_p = \frac{1}{C}\,\frac{S}{\left(S - \frac{1}{2R_tC} + j\sqrt{\frac{1}{LC} - \left(\frac{1}{2R_tC}\right)^2}\right)\left(S - \frac{1}{2R_tC} - j\sqrt{\frac{1}{LC} - \left(\frac{1}{2R_tC}\right)^2}\right)}$$

The real and imaginary portions of the complex poles can now be extracted and plotted on an S plane diagram. Stability analysis is most easily done by plotting the root loci of the poles. The root locus is a circle as shown in Fig 3.11

The diagram shows four distinct regions, and it can be analysed in these four.

Region A = The overdamped case.

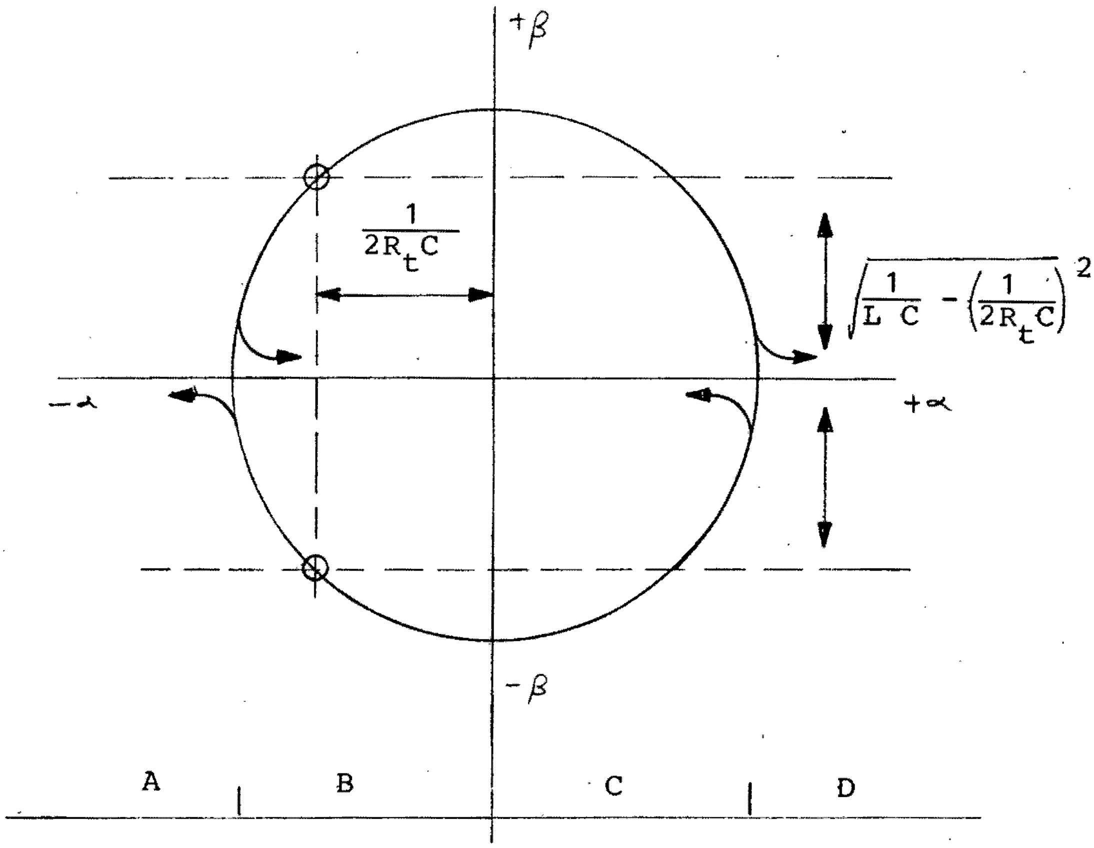
Region B = The underdamped case.

Region C = Sustained oscillation. (Unstable)

Region D = Marginally stable. (Amplification)

Regions C and D can easily be reached if the lossy components in the system are more than counteracted by the energy sources in the system.

Positive resistance can be regarded as an energy sink while negative resistance can be regarded as an energy source. More simply, R_g and R_i can be regarded as a potentiometer and then V_{if}/V_i is given as follows :-

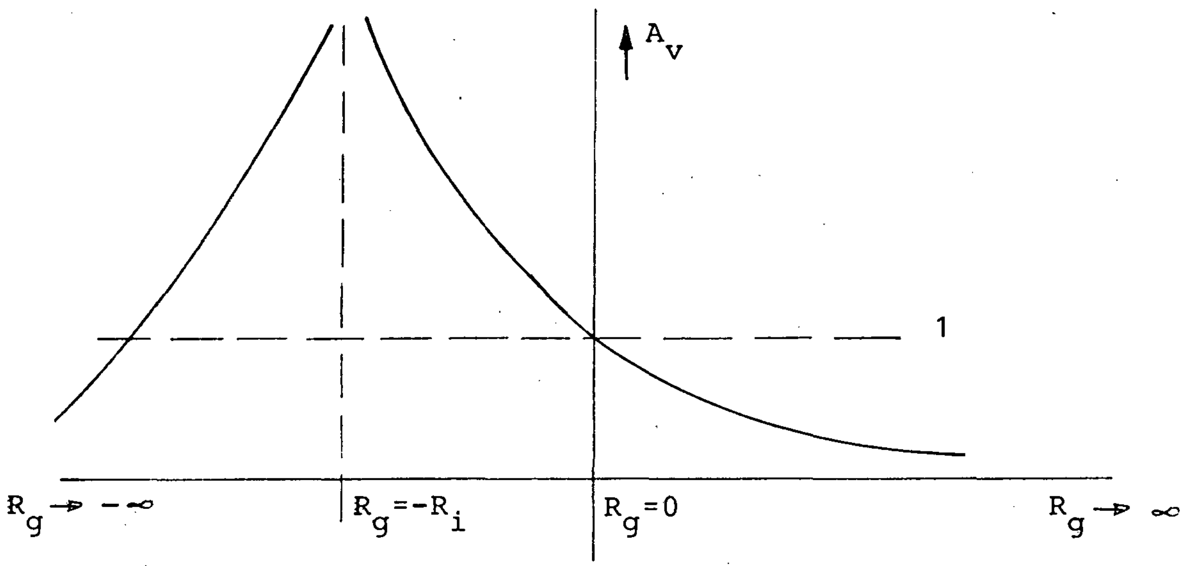


$+\beta$
$\frac{1}{2R_tC}$
$\sqrt{\frac{1}{LC} - \left(\frac{1}{2R_tC}\right)^2}$
$-\alpha$
$+\alpha$
$-\beta$
A
B
C
D

Fig 3.11

$$\frac{V_{if}}{V_i} = \frac{R_i}{R_g + R_i} = A_v$$

Now if $R_g = -R_i$ then A_v tends to infinity.



This is the simple gain curve, however oscillations will occur under certain conditions because of the presence of the LC network. From the locus diagram, oscillations will exist in region C. In terms of R_t the region is given as follows:-

If $-1/2R_tC$ is greater than 0, and $(1/2R_tC)^2$ is less than or equal to $1/LC$ then the poles will lie in region C.

This means that R_t should must be bounded as follows:-

$R_t < 0$ and $|R_t| > \sqrt{(L/C)}/2$.

Now R_i is constant, but R_g can become negative and hence the boundaries should be redrawn in terms of R_g.

In order to meet the first condition, ie. that R_t be negative, R_g must lie in the range:-

$$0 > R_g > -R_i \text{ - - - - - - - - - - - - } (a)$$

The second boundary is somewhat more difficult to set. The modulus of R_t tends to increase until $R_g = R_i$. At this point the poles shift to the left hand side of the Pole/Zero diagram and stability is reached. The breakpoint will thus reached as follows:-

$$\frac{R_i \, |R_g|}{R_i - |R_g|} = \frac{1}{2}\sqrt{\frac{L}{C}}$$

or

$$|R_g| = \frac{R_i}{\frac{2}{R_i}\sqrt{\frac{C}{L}} + 1} \text{ - - - - - - - - - - - - } (b)$$

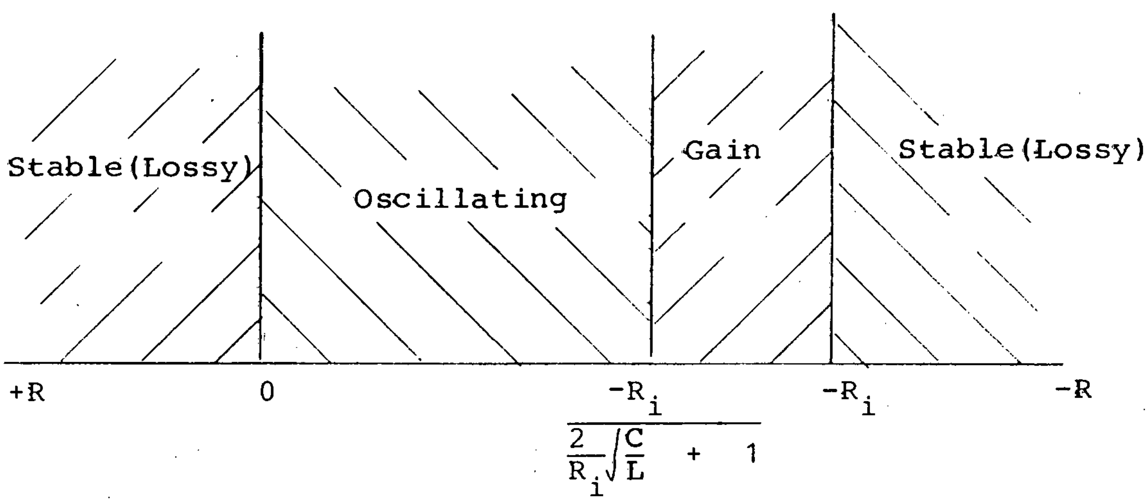
Now $R_g > -R_i$ to satisfy the first inequality (a). This means that a state of oscillation will exist while R_g ranges as follows:-

$$0 > R_g > -R_i \quad \text{and} \quad |R_g| < \frac{R_i}{\frac{2}{R_i}\sqrt{\frac{C}{L}} + 1}$$

Now the second term of the inequality has a factor greater than 1 in its denominator and hence it will always be less than R_i. A state of controlled amplification could thus concievably exist. The range of R_g for controlled amplification would thus be as follows:-

$$-R_i < R_g < \frac{R_i}{\frac{2}{R_i}\sqrt{\frac{C}{L}} + 1}$$

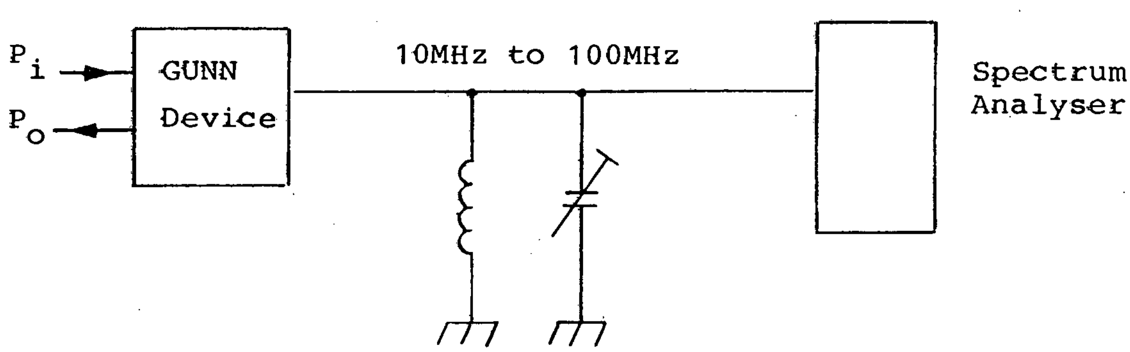
R_g can be plotted on a scale showing the various regions where stability will occur.



3.6.2 Measurement of Rg.

Measurement of R_g cannot be done by direct measurement on a vector impedance bridge because of the presence of energy sources other than the bridge supply. The alternative is to place the component in a passive circuit and to measure it's effect on the circuit.

In the actual experement a high Q tuned circuit similar to the circuit in the previous investigation was placed across the GUNN terminals and the change in Q was measured. A pseudo white noise source was applied at the Microwave Frequency and the output spectrum was observed on a spectrum analyser. The bandwidth and centre frequency could then be readily observed. The block diagram was as follows:-



If it is assumed that R_g can range from minus infinity to plus infinity then there are three possible results to be observed. (a)Reduction of observed Q. (b)Improvement of observed Q. (c)A state of oscillation. If the unloaded Q of the network is high then the network poles will be close to the vertical axis and reduced will indicate positive GUNN resistance. Increased Q will indicate the range of controlled amplification and oscillation will indicate the range of R_g to give instability.

R_g is dependant on the I/V characteristic of the device and hence will be dependant on the bias voltage applied.

If the equivalent circuit is now slightly redrawn to represent the GUNN as a current generator with equivalent parallel resistance then this resistance is given as follows:-

$$R_g = \frac{QQ'wL}{Q - Q'}$$

Where

Q = Unloaded Q of the Coil.

Q' = Loaded Q of the Coil.

L = Inductance of the coil.

3.6.3 Observed results.

In the first place it was observed that only two conditions could be reached with the circuit components as shown. i.e. A stable condition and a state of oscillation. Now if R_i is large (Say greater than 10K) then the inequality becomes as follows:-

For Stability:

$R_g > 0$ (Where R_g is the series component)

For Oscillation:

$$0 > R_g > -\frac{1}{2}\sqrt{\frac{L}{C}}$$

For Amplification. (Increased Q)

$$-R_i < R_g < -\frac{1}{2}\sqrt{\frac{L}{C}}$$

In the actual experement the breakpoints were calculated as follows:-

R_i = 10K

C = 22pF

L = 148nH

This gives values as follows:-

$(L/C)/2$ = 41ohms

This means that a state of amplification could occur. If the bias voltage were sloely reduced the small signal resistance R_g would become more negative. R_g would reach the -41ohms point first. ie $R_g > -41$ and would also be $> -R_i$ and an area of damped oscillation would appear. As soon as R_g becomes less than -10K a state of lossy stability would occur. In practice a state of amplification never occured and hence it could assumed that the negative resistance region could never reach -41 ohms. A change of the ratios of L and C was made to the following values to observe any change in the conditions.

L = 36nH

C = 86pF

The break points would hence be at 10ohms and not at 41ohms. This did not assist the stability situation and a state of amplification was still not reached.

3.5.4 Conclusion.

It seemed that selection of the (L/C)/2 breakpoint did not affect the interface stability to any great extent as reduction of the bias voltage did not result in a region of stable gain.

This suggests that the L and C selection criteria should only depend on the required Q and centre frequency and not on any gain conciderations for the IF interface.

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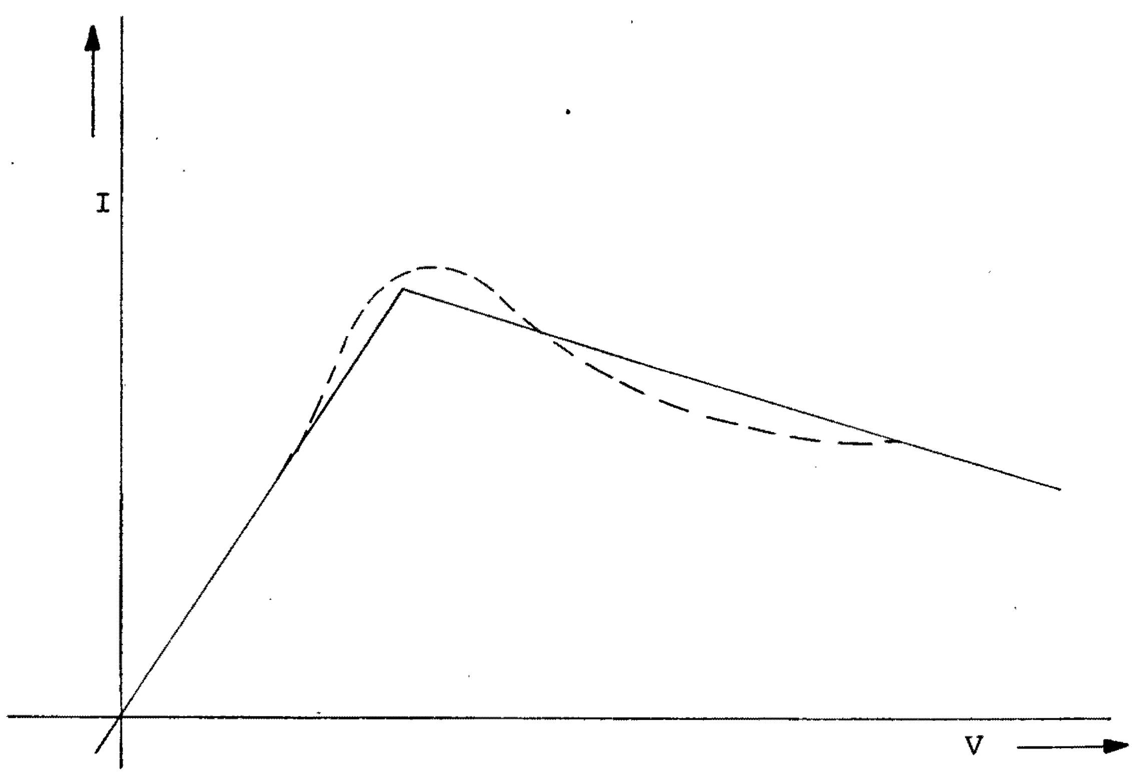
CHAPTER 4

IMPROVEMENT OF CORRELATION BETWEEN THEORY AND EXPERIMENT

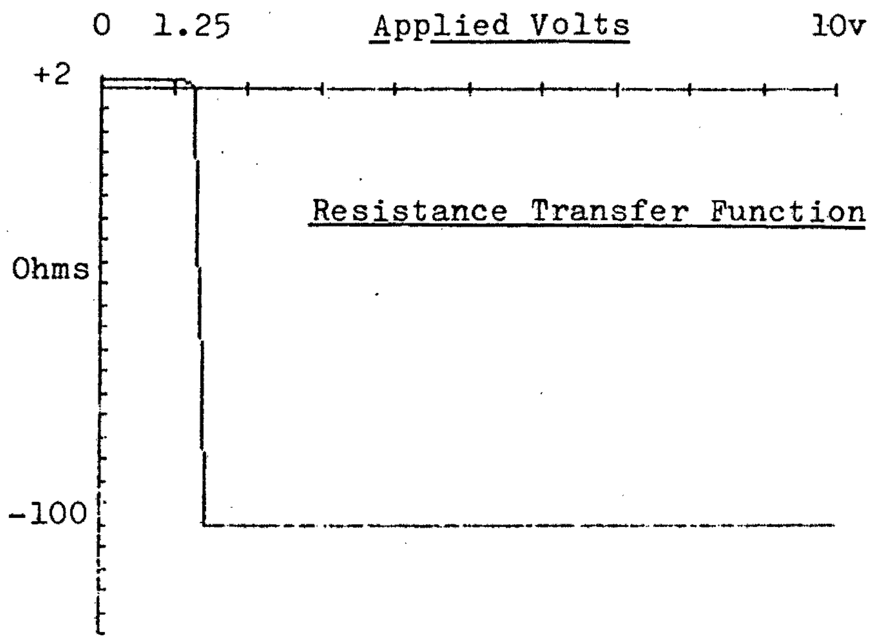
4.1 Computer Simulation of Conversion Loss

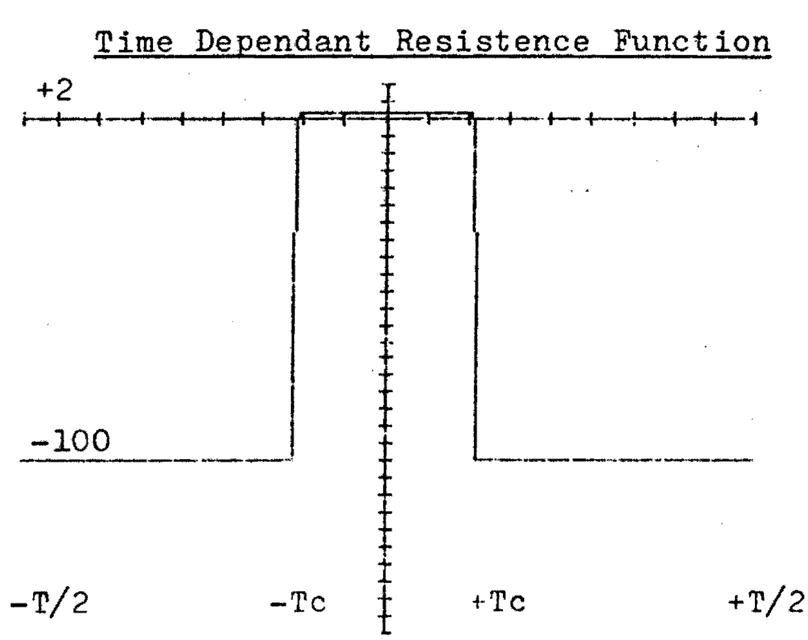
In this section we will try to demonstrate a technique for using a computer simulation to derive a more accurate figure for Conversion Loss.

Our original assumption as regards the transfer charactaristic of the GUNN and the resultant Time Dependent Reflection Coefficient was a straight line function and a rectangular wave for the Time Dependent Reflection Coefficient. In practice this is not the case and there results a small error in the Conversion Loss.



Straight Line Transfer Characteristic against actual Transfer Characteristic.


0
1.25
Applied Volts
10v
+2
Resistance Transfer Function
Ohms
-100


Time Dependant Resistence Function
+2
-100
-T/2
-Tc
+Tc
+T/2

The computer simulation was done on a Hewlett Packard desk top computer, type HP85.

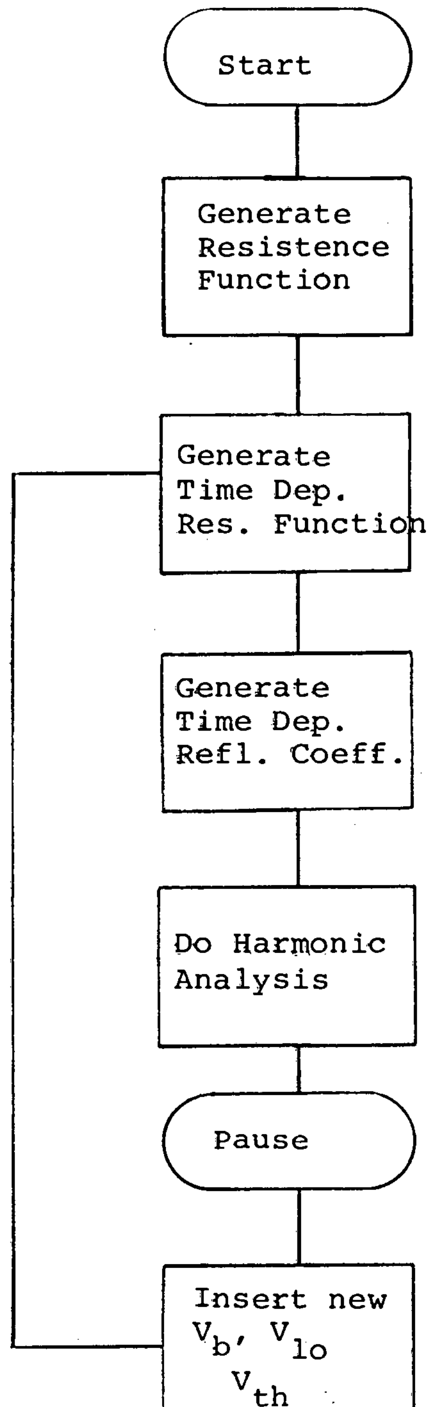
4.1.1 Computer Simulation

The computer simulation was done in two phases and requires two different programs.

The starting point was the two input variables. These were fed into the first program which then outputs the Time Dependent Reflection Coefficient. It generates two intermediate stages i.e. The differential of the current/voltage Transfer Function which when inverted gives a Resistance Transfer Function. Then the second intermediate stage is the generation of the Time Dependant Resistance Function. This stage requires the insertion of the Bias Voltage and Local Oscillator Amplitude variables.

The Time Dependent Reflection Coefficient was then inserted into a program which did a Harmonic Analysis of the waveform assuming it to be periodic.

The process can be shown graphically as follows:



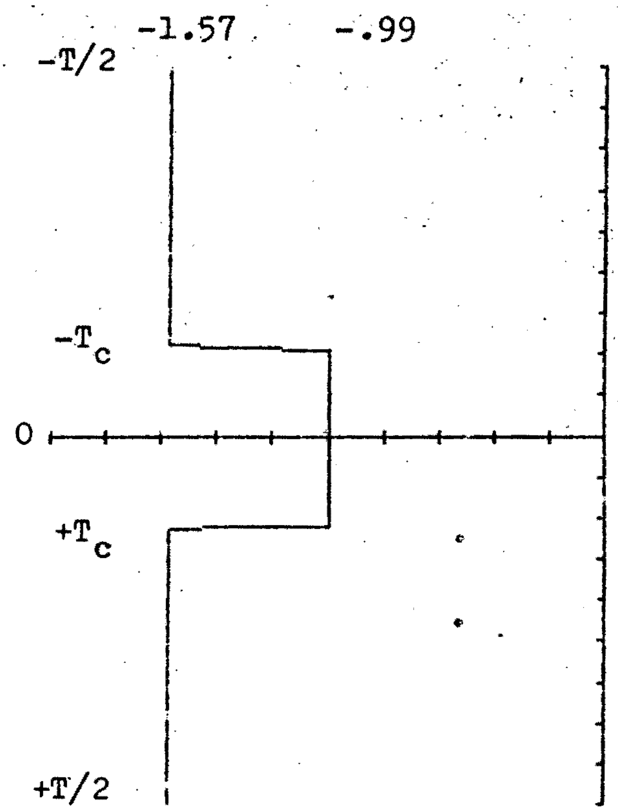
<u>Flow Chart Representation of the Process.</u>

The listings of the programs in Basic are given in the appendices.

4.1.2 Results

These results were obtained for a range of bias voltages and local oscillator amplitudes. The results were then plotted for the two variables seperately.

TIME DEPENDANT REFLECTION COIEFF.



V_b=5v V_{th}=1.25

V_l=5v

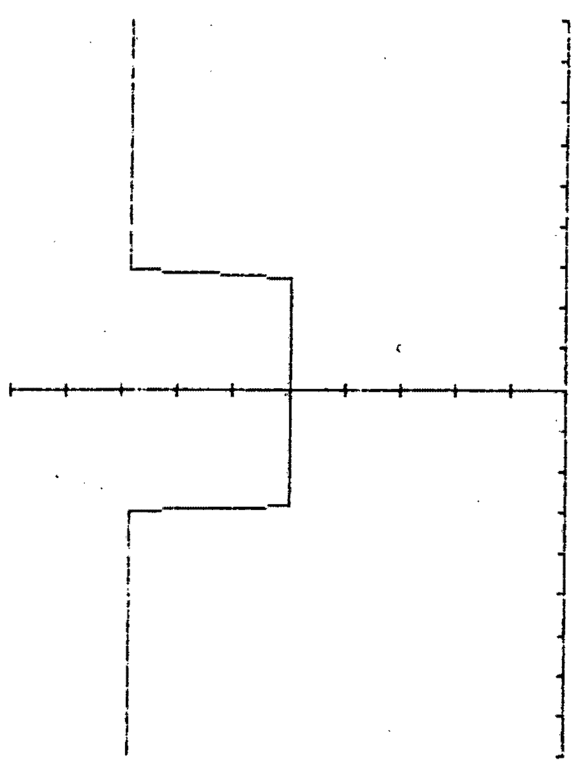
FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.43E+000	
1	1	-2.56E-001	-5.79E-011
2	2	18.45E-002	60.93E-012
3	3	-9 20E-002	-7.58E-011
4	4	73.26E-004	-1.36E-011

.1.2.1 Fixed Local Oscillator Amplitude with varying Bias Voltage.

$V_b=3.5v$ $V_{lo}=3v$

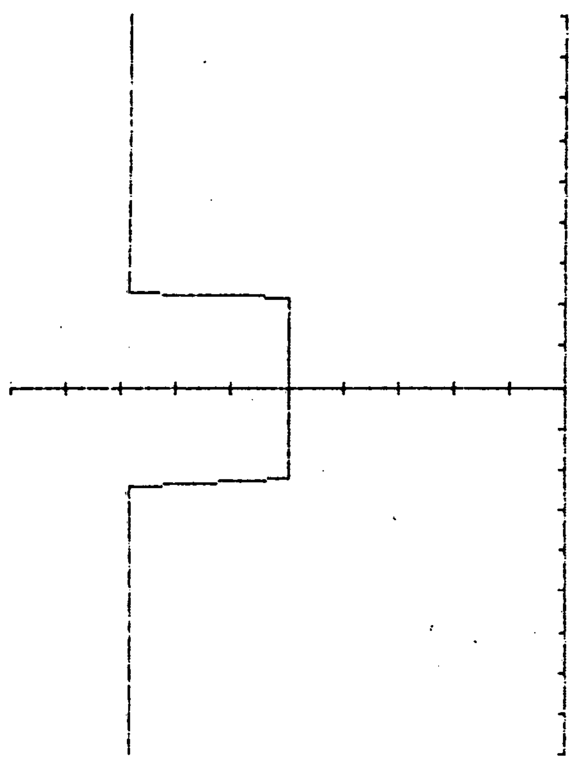
$V_{th}=1.25v$



```
FOURIER COMPONENTS (PEAK VALUES)
HARM  FREQ        A(N)        B(N)
0     0    -1.38E+000
1     1    -3.11E-001   -6.59E-011
2     2    16.71E-002   62.93E-012
3     3    -1.55E-002   -2.78E-011
4     4    -7.09E-002   -7.56E-011
```

$V_b=3.5v$ $V_{lo}=3v$

$V_{th}=1.25v$

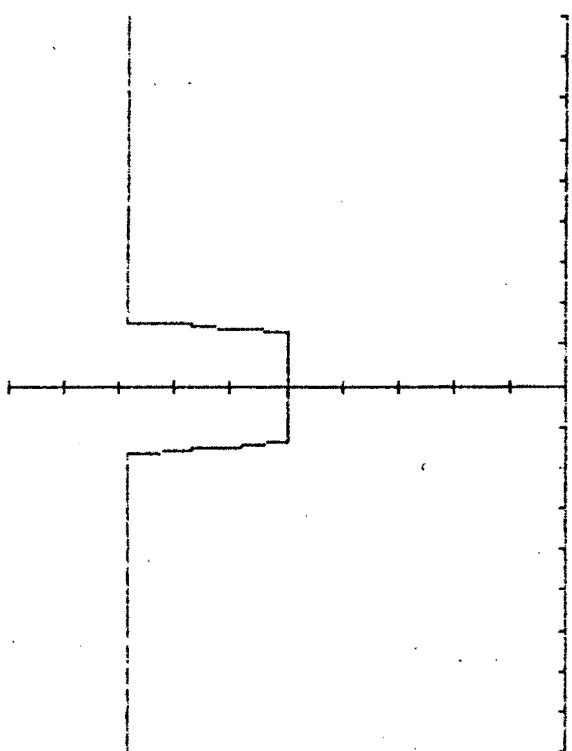


FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.42E+000	
1	1	-2.63E-001	-6.19E-011
2	2	18.46E-002	64.93E-012
3	3	-8.44E-002	-6.98E-011
4	4	-3.56E-003	-2.56E-011

V_b=4v V_{lo}=3v

V_{th}=1.25

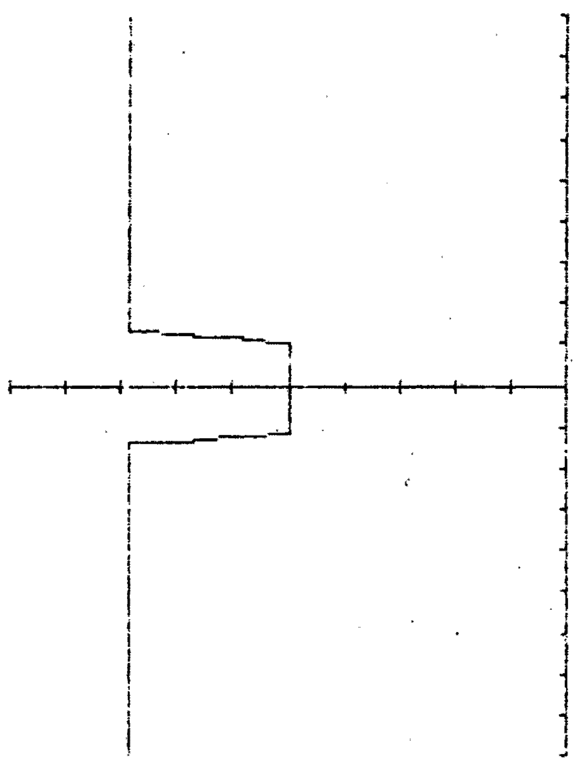


FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1 47E+000	
1	1	-1.82E-001	-5.79E-011
2	2	15 85E-002	50.93E-012
3	3	-1.22E-001	-9 38E-011
4	4	89.46E-003	46 38E-012

V_b=4.125 V_{lo}=3v

V_{th}=1.125v

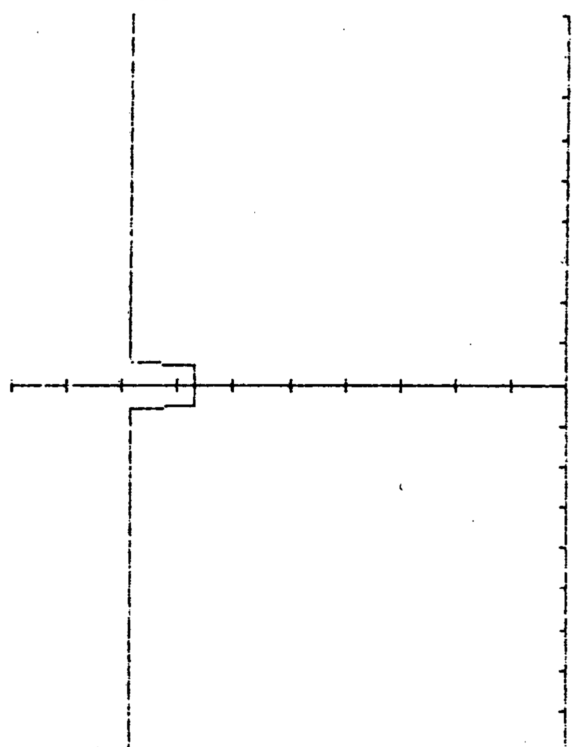


FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.49E+000	
1	1	-1.53E-001	-4.99E-011
2	2	13.96E-002	42.93E-012
3	3	-1 17E-001	-8.58E-011
4	4	90.52E-003	54.38E-012

V_b=4.4v V_{lo}=3v

V_{th}=1.25v



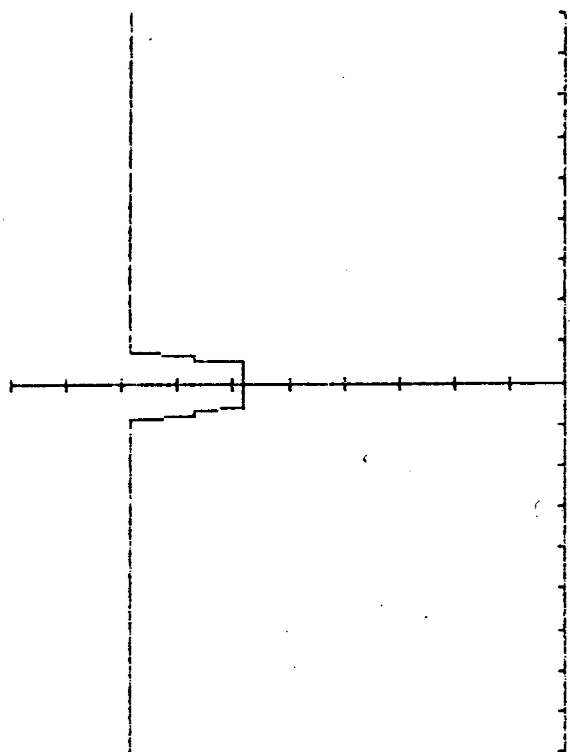
FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.55E+000	
1	1	-2.64E-002	-2.99E-011
2	2	25.97E-003	-7.07E-012
3	3	-2.52E-002	-3.98E-011
4	4	24.25E-003	38.40E-014

4.1.2.2 <u>Fixed Bias Voltage with Local Oscillator amplitude changing</u>

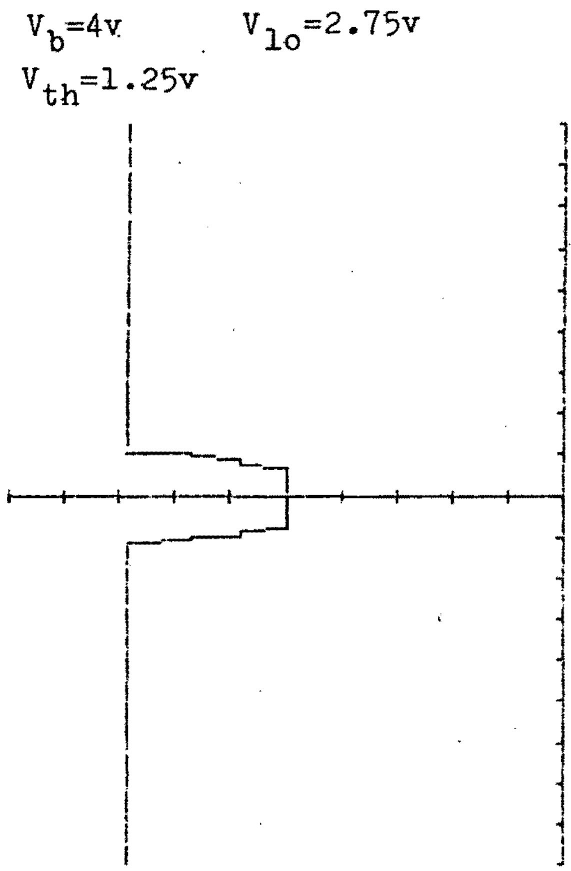
V_b=4v V_{lo}=2.65v

V_{th}=1.25



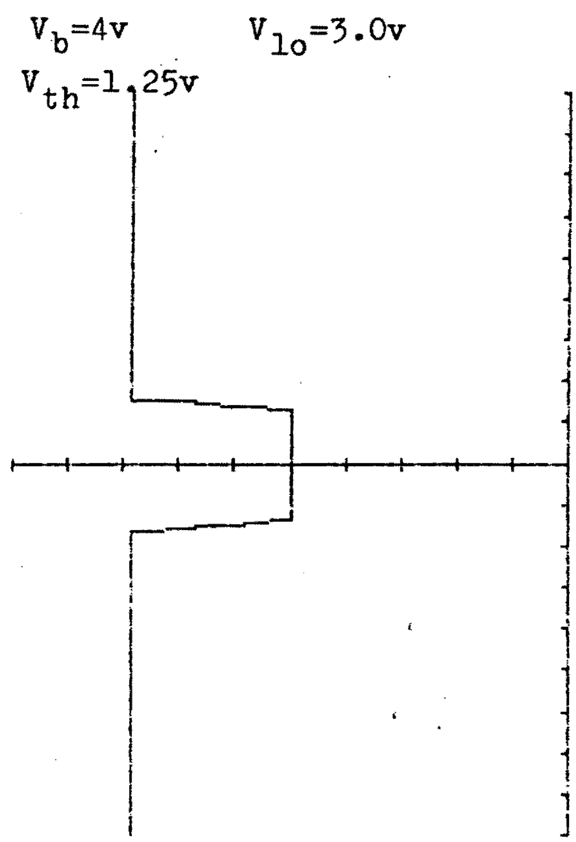
FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.54E+000	
1	1	-6.22E-002	-4.39E-011
2	2	60.40E-003	89.29E-013
3	3	-5.73E-002	-5.38E-011
4	4	53.24E-003	22.38E-012



FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.51E+000	
1	1	-1.19E-001	-4.99E-011
2	2	11.29E-002	30.93E-012
3	3	-1.01E-001	-8.38E-011
4	4	87 59E-003	52.38E-012

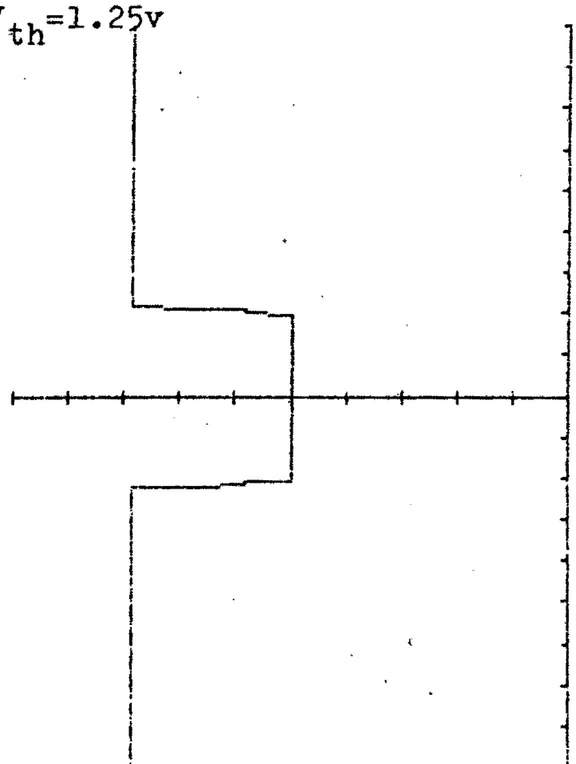


FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.47E+000	
1	1	-1.82E-001	-5.79E-011
2	2	15.85E-002	50.93E-012
3	3	-1.22E-001	-9.38E-011
4	4	80.46E-003	46.38E-012

V_b=4v V_{lo}=3.5v

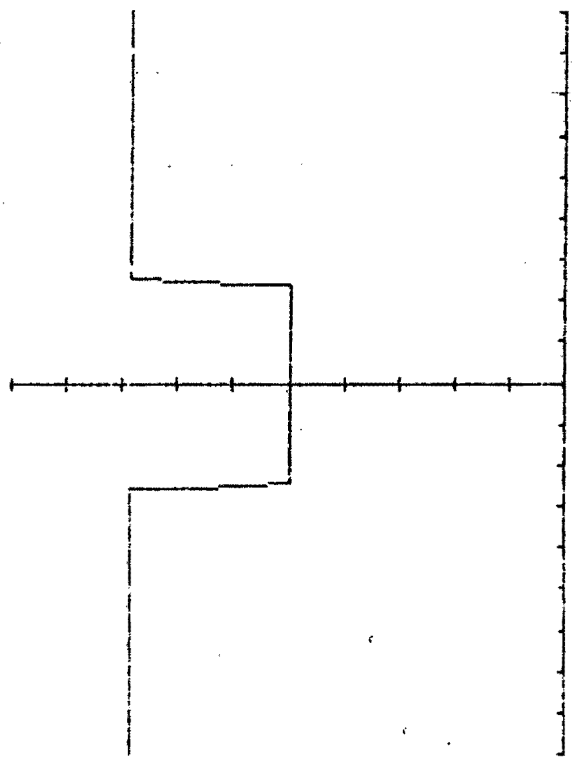
V_{th}=1.25v



FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.43E+000	
1	1	-2.48E-001	-5.79E-011
2	2	18.37E-002	60.93E-012
3	3	-9.83E-002	-7.98E-011
4	4	17.35E-003	-7.61E-012

V_b=4v V_{lo}=4v
V_{th}=1.25v



FOURIER COMPONENTS (PEAK VALUES)

HARM	FREQ	A(N)	B(N)
0	0	-1.41E+000	
1	1	-2.81E-001	-6.19E-011
2	2	!8.23E-002	66 93E-012
3	3	-6.38E-002	-5.58E-011
4	4	-2.89E-002	-4.16E-011

4.1.3 Conclusion

A typical set of computer results against calculated results was taken as follows:

V_b	V_{lo}	V_{th}	L_c(calc.)	L_c(computer)
5	5	1.25	-16.3dB	-18.0db
4	3	1.25	-20.7dB	-20.8dB
4.125	3	1.25	-23.0dB	-22.0dB
3.5	3	1.25	-16.3dB	-17.6dB

Notice that the results were taken with a negative resistance of -100 and that the Conversion Losses are rather large.

As can be seen the calculated results vary only marginaly from the computed results. This suggests that the original Straight Line model was quite adequate.

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4.2 Harmonic Mixing

In this section we consider the use of the receiving device at harmonics of its fundamental frequency. The author was unable to verify any of the precepts given in this section as equipment was not available for work at multiples of 35 GHz. (This being the author's frequency of interest).

4.2.1 Operation

The device would operate by mixing the received signal with one of the harmonics of the Time Dependent Reflection Coefficient. The Time Dependent Reflection Coefficient appears to have predominent odd harmonics, however even harmonics would exist. These harmonics would be considerably lower in amplitude than the fundamental and hence their noise skirts would have an equivalent lower amplitude. This holds out the promise that mixing with the harmonic could lead to an improved Noise Figure than with fundamental mixing.

The improved Noise Figure would lead to range improvement if the same transmit power could be maintained. The other aspect to consider would be the effect of atmospheric absorbtion at higher frequencies.

All of this leads to an interesting speculation. The fact is that the received frequency at a harmonic of the fundamental is in fact mixing with a harmonic of the Time Dependent Reflection Coefficient and not with an actual harmonic of the fundamental. We may now assume that this fundamental energy of the local oscillator is spectraly distributed according to a Gausian distribution around its fundamental as well as being amplitude modulated according to the thermal noise on the GUNN terminal. - each giving rise to its particular noise skirt. (See section 3.4).

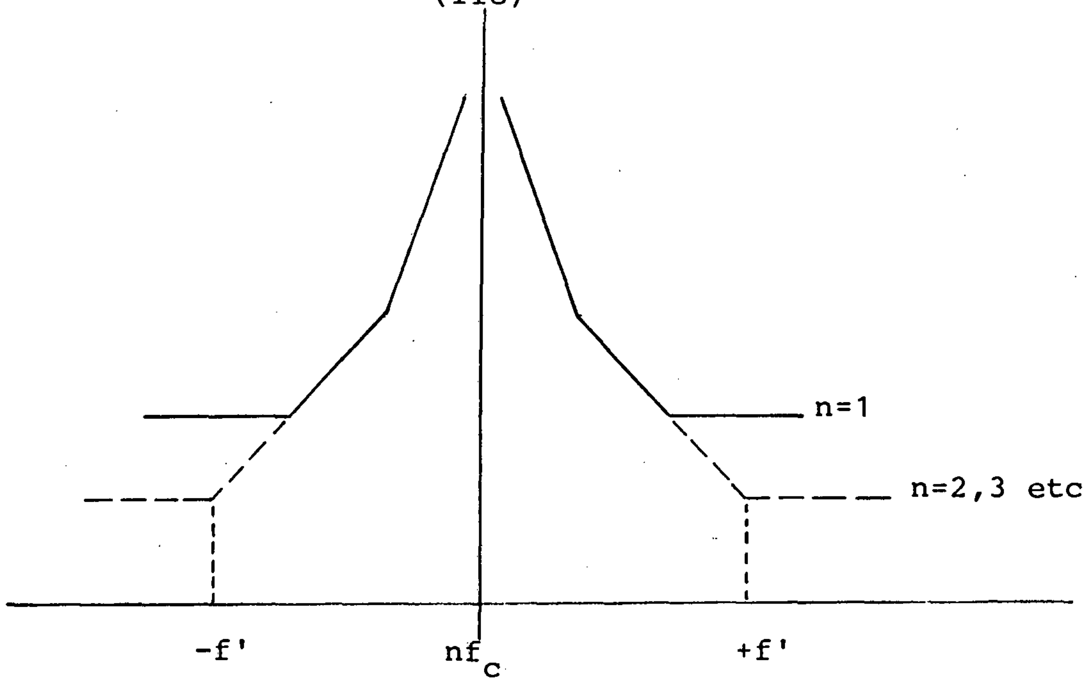
If we assume that each term of the Time Dependent Reflection Coefficient is given by H_n where this can be represented as follows:

$$H_n = K.k_n.Cos[nwt]$$

Now K is a statistical function in time and gives rise to the A.M. noise skirt of the harmonic. w is also a statistical function in time and gives rise to the F.M. noise skirt of the harmonic.

In general K_n will be a factor rather less than 1 and hence the A.M. noise skirt will be reduced. On the other hand, w is multiplied by n and hence the F.M. noise skirt will be increased.

This means that the point at which the F.M. noise skirt becomes less significant than the A.M. noise skirt is further removed in frequency from the "carrier" by both the mechanisms described in the previous paragraph.



Where n = 1,2,3 etc

This suggests that if one could use a high I.F. frequency then one could still be in the region of the A.M. noise and outside the region of F.M. noise and hence still obtain an improved Noise Figure.

4.2.2 Power of the Harmonic of the T.D.R.C.

n	Effective F'	Effective kn
1	0dB	0.263
2	+7dB	0.184
3	+10dB	0.0844
4	+24dB	0.00356

4.2.3 Possible practical implementation

Practical problems would lie in the area of the electronic tuning range. If one required to use the device as a transceiver then one would have to select devices to have substantially different centre frequencies to be able to cope with the overly wide electronic tuning range required.

On the other hand if one required a Data link in one direction only, then one could set up a system as follows.

Example of a possible system.

If it was required to use say Ka band (35Ghz) as a carrier then one could use a GUNN diode at 35GHz as the transmitter but use a cheap GUNN at about 12GHz as the receiver, but mix with its third harmonic. The IF frequency could then be guite freely chosen to place it ouside the new resultant F.M. noise skirt.

4.2.4 Conclusion

Harmonic mixing seems to hold various outstanding advantages for communications links but only for links that nead to operate in one direction only. A great deal of work would still need to be done in order to establish the feasibility of such a system.

The other advantage of harmonic mixing could be that it allows operation at higher frequencies without having to provide apparatus at these higher frequencies. This could mean for example that the beamwidth of an antenna of a certain aperture could be reduced by a factor of three if the third harmonic were used instead of the fundamental.

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4.3 Effects of dis-similar powers in the two GUNN Transceivers.

In this section we try to show that if the GUNN transceivers have dis-similar power outputs then the resultant range performance will be degraded. The powers referred to are the intrinsic powers of the diodes and not the output powers of the sources. By this we mean that attenuation of the output power to match the power output of the other device will only degrade performance further.

4.3.1 Basic Range Performance.

If we take the criterion discussed in Chapter 2 section 2 then we arrive at a figure for maximum path loss and hence for maximum range given as follows.

$$L_p(\mathrm{dB}) = 10\mathrm{Log}\left(\frac{P_o}{N_i}\right) - 10\mathrm{Log}\left(F\right) + 10\mathrm{Log}\left(G_r G_t\right)$$

Now in a Self Oscillating Mixer the Noise Figure is substantially dependent on the receiving transceiver's power output . See Section 3.4.3.

$$F = 1 + N_i/kT$$

Where $N_i = P_s$ x Noise to Carrier Ratio.

Say that Noise to Carrier ratio is constant and is given by M, then the overall expression becomes as follows:-

$$L_p(\mathrm{dB}) = 10\mathrm{Log}\,\frac{P_o}{(N_i)'} - 10\mathrm{Log}\left(1 + \frac{P_sM}{kT}\right) + 10\mathrm{Log}\,G_rG_t$$

But $(N_i)' = kT$ from the definition for Noise Figure.

P_o = Transmit Power

P_s = Source Power

$L_p(\mathrm{dB})$ = Maximum possible path loss.

Removing the logs gives:-

$$L_p = \frac{P_o}{kT} \times \frac{1}{(1 + P_sM/kT)} \times G_rG_t$$

$$L_p = \frac{P_oG_rG_t}{kT + P_sM}$$

L_p increases as Po, Gr or Gt increase which is intuitively correct. If Ps or M should increase then L_p will decrease. The anomaly is that for the reverse path the two powers are precisely reversed in their positions in the equation.

$$L_p = \frac{P_sG_rG_t}{kT + P_oM}$$

Assuming that M is the same ratio for each source.

4.3.2 Effect on System Performance

If we remember that Conversion Loss plays very little part in the System Performance and hence that changing Conversion Loss due to changing Receiver Power also plays very little part in System Performance, then we can set up a number of criteria for establishing the respective source powers.

1) Where both sides transmit and receive over the same channel.

   In this case the Source Powers should be equal.

2) One way transmission

   In this case the power of the transmitter GUNN should be maximum and the power of the receive GUNN should be minimum.

4.3.3. Evaluation of System Performance for Unequal Powers

For equal forward and reverse path losses we can set the ratio of the two expression to be as close to one as possible.

$$\frac{L_p}{L_p(\text{reverse})} = 1 = \frac{P_o}{P_s}\,\frac{(\,kT + P_oM\,)}{(\,kT + P_sM\,)}$$

If $P_o = P_s$ then this condition is true. If on the other hand the two powers are not equal (say $P_o = 2P_s$) then the range performance would be impaired in one direction but not the other.

Say $M = -160dB$ $kT = 4.14E-21$

Therefore $P_oM = P_o\ 10E-16$

We can thus ignore kT (This is a particular case but can generally be applied)

$$\frac{L_p}{L_p(\text{reverse})} = \frac{P_o^2}{P_s^2}$$

But for $P_o = 2P_s$

$$\frac{L_p}{L_p(\text{reverse})} = 4$$

Thus performance of one or the other path is degraded or enhanced by 4.

It is useful to consider this in terms of decibells.

$$\frac{L_p}{L_p(\text{reverse})} = 6dB$$

In practical terms the effect would be as follows:

Say 35GHz channel and L_p was equal to 155dB. This would indicate a range of 40km in free space plus the oxygen attenuation. If L_p(reverse) were now 6dB less than this would approximate to 20km or half the range in the other direction. This means that the overall System Performance will be degraded to 20km.

It would be better to specify Source Powers in terms of a tolerance of an absolute power, than to specify these powers in terms of the power of the other source.

eg. Say Source Power was given as $P_s = P_a x1.414$ to $P_s = P_a/1.414$. Then the worst possible case would be when $P_s = P_a x1.414$ and $P_o = P_a/1.414$.

$$\frac{L_p}{L_p(\text{reverse})} = \left(\frac{1}{1.414} \times \frac{1}{1.414}\right)^2 = 4$$

This is still equal to 4 but loss in one direction is increased by 3dB while that in the other direction is degraded by 3dB. This would degrade range by about 10km only.

Rewriting the expression to give:-

$$\frac{L_p}{L_p(\text{reverse})} = \frac{P_o^{\,2}}{P_s^{\,2}} \frac{(kT/P_o + M)}{(kT/P_s + M)}$$

If M = -180dB or greater, then we would find that M becomes less significant than kT/P_o or kT/P_s. The expression then simplifies to:-

$$\frac{L_p}{L_p(\text{reverse})} = \frac{P_o}{P_s}$$

This is a much improved situation and would allow more latitude in specifying Source Powers. This would be more like the situation for a local oscillator with a separate mixer. However in practice this situation does not arise when using the Self Mixing Oscillator.

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4.4 Mounting in Image Waveguide

The thrust of this research has been to show that the GUNN diode can be used as the receiver and the transmitter of a system. It is thus fully consistent with this approach to show how the device can be mounted in cheap waveguide structures to obtain full benefit from the technique.

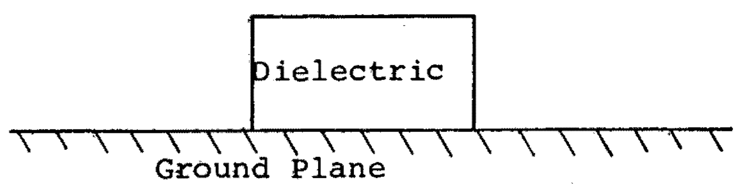
4.4.1 Image Waveguide

In recent years there has been considerable research into the use of Dielectrics as wave guiding structures. These were first studied in Germany before World War 2 as the radiating element in phased arrays for Radar applications.

In essence a waveguide is a structure that guides electromagnetic waves along a certain route by means of these waves interacting with at least two boundaries. The conventional rectangular waveguide accomplishes this by setting conductive boundaries to an air dielectric space. The Dielectric Waveguide does this by establishing boundaries between air and some form of plastic or mylar materials with a permitivity quite different to that of air.

The Image Waveguide is a sort of hybrid between the conductive waveguide and the Delectric Waveguide in that it has both dielectric boundaries and conductive boundaries.

The structure of the waveguide is as follows:

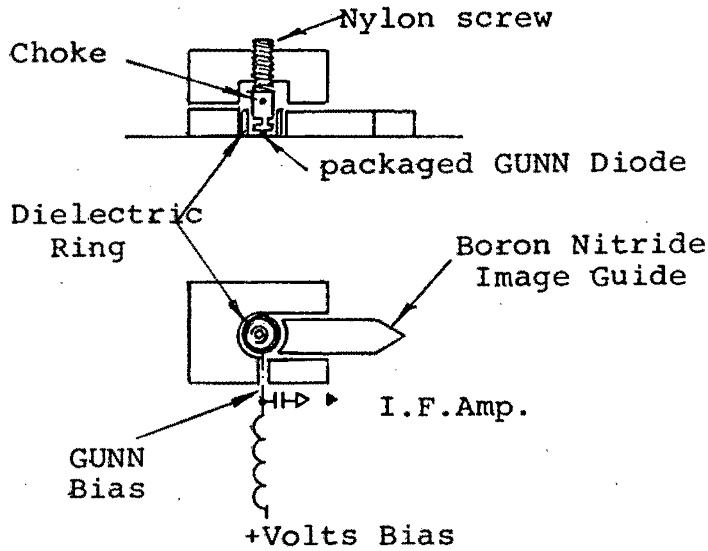


The particular properties of the waveguide are that it theoretically does not have a lower cut-off frequency under which it becomes evernescent. It also does not confine the wave to the Dielectric but the Dielectric guides the wave.

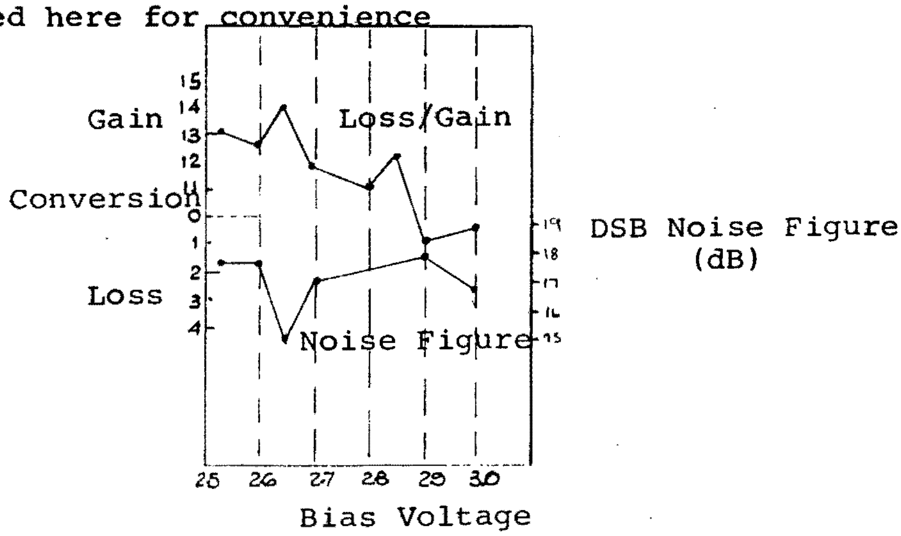
In all other aspects the guide can be used as a conventional transmission line ie. terminated sections of a fraction of a wavelength exhibit reactance and not pure resistance. Half wavelength sections of the guide act as resonators. Circulators and directional couplers can also be constructed. See references (42,43) for further details.

4.4.2 The GUNN Diode in Image Waveguide

A GUNN source can be constructed using an Image Waveguide resonator. The heat is conducted away by the groundplane and the groundplane also forms the second contact for the diode. Chrepta and Jacobs in their papers (6,7) showed that a binocular radio at 90GHz could be constructed using Image Guide. Their construction was as follows:



The results as published in their paper are repeated here for convenience



Note that they seemed to apply the theory discussed in section 4.5 in analysing their system performance.

4.4.3 Conclusion

There seems to be a bright future for the S.M.O. scheme where cheapness is of special importance. This is particularly the case where cheap mounting structures such as Image Waveguides are available.

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5 Device as a Negative Resistance Amplifier followed by a non linear transfer function.

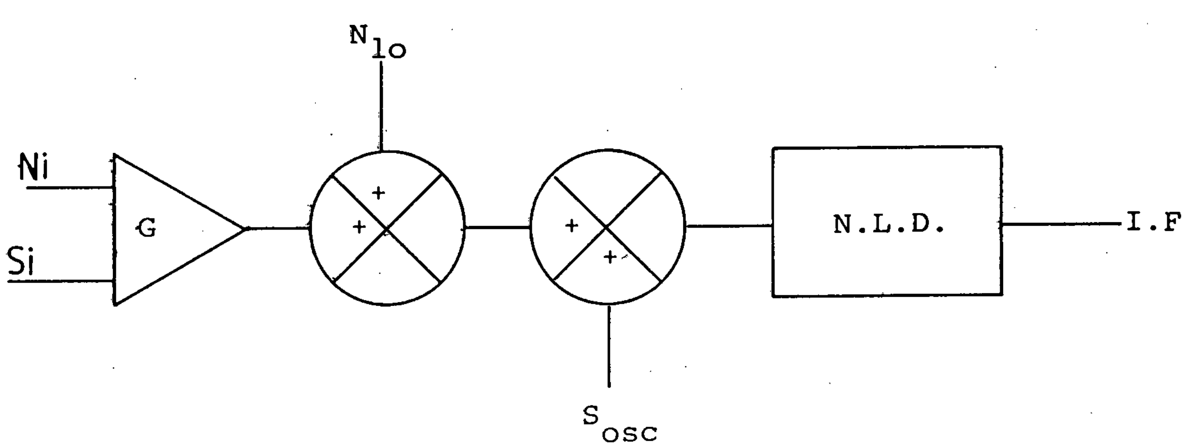
In the initial stages of this research the author took the view that the device consisted of an amplifier/attenuator followed by a non linear device. This seemed to comply to some extent with the view taken by the researchers at Lancaster [2,3,5]. However the author was unable to repeat the results given in the papers referred to above.

There seemed to be considerable emphasis on the noise presented to the IF by Shot Noise, Flicker Noise and other Noise Sources in the GUNN current itself. Dr F. Myers of the Plessey Research Centre at Casswell in the United Kingdom expressed the view that GUNN current fluctuations would be the dominant Noise Source in the GUNN/IF Interface. In the author's opinion most of the research done at Casswell and at Lancaster concentrated on the use of the S.M.O as a Doppler Radar for intruder alarms etc. This implies an IF frequency equal to the Doppler frequency and hence well within the F.M. Noise skirts of the device. The author's research has concentrated on the use of the device as a receiver with a fixed IF frequency greater than 10MHz.

The Noise Figures quoted in the papers from Lancaster seemed to be very optimistic and would be consistant with the Negative Resistance Amplifier/Non Linear Device approach. The author thus felt the theory to be deficient to some extent and hence the alternative approach presented earlier in this thesis was developed. As has been shown this alternative approach seems to be far more consistent with practical results obtained.

The first approach of Negative Resistance Amplifier etc. is presented in this section only as an attempt at completeness. It does have some merit as a technique for analysis and with some thought it could be corrected to present more accurate results.

4.5.1 System Layout



Where:-

S_i = Signal Input

N_i = kTB = Noise from Input Termination

NRA = Negative Resistance Amplifier

G = Gain of Negative Resistance Amplifier
(can be less than one)

N_{lo} = [F - 1]kTB = Local Oscillator Noise at frequency of interest.

S_{osc} = Signal from the Local Oscillator

N.L.D. = Non Linear Device

Notes

It must be assumed that the noise introduced into the system can all be lumped into the N_{lo} term. It must also be assumed that the Negative Resistance Amplifier has a Noise Figure of 1. The Non Linear Device can be assumed to have the same effect on the Noise, Input Signal and Local Oscillator Signal and can hence be regarded as an amplifying factor to the sum, difference and all other components.

4.5.2 Gain of the Negative Resistance Amplifier

The GUNN device is used quite successfully as a negative resistance reflection amplifier (20,21). The device is, however, very prone to instability (hence its use as an oscillator) but it may present a nett negative resistance under certain conditions whilst still oscillating. If it now acts as the termination of a waveguide then the reflected power is a function of the negative resistance. If the resistance is positive then the device acts as an attenuator.

If the waveguide impedance is given by Z_o then the VSWR is given by:

$$VSWR = \frac{R - Z_o}{R + Z_o}$$

Where R is GUNN resistance. If V_i is the incident voltage then the reflected voltage is given by:

$$V = V_i \times VSWR$$

and absorbed voltage is given by:

$$V_i - V = V_a$$

therefore $V_a = V_i\ (1 - VSWR)$

Incident power is given by

absorbed power is given by

$$P_a = \frac{V_i^2(1 - |VSWR|)^2}{R}$$

gain of the device is given by

$$\frac{P_a}{P_i} = \frac{(1 - |VSWR|)^2}{R} \frac{Z_o}{1}$$

Therefore:

$$\frac{P_a}{P_i} = \frac{Z_o}{R} (1 - 2|VSWR| + |VSWR|^2)$$

Therefore

$$\frac{P_a}{P_i} = \frac{Z_o}{R} - \frac{Z_o}{R}\left(\frac{R - Z_o}{R + Z_o}\right) + \frac{Z_o}{R}\left(\frac{R - Z_o}{R + Z_o}\right)^2$$

If $R = Z_o$ then $P_a/P_i = 1$

If $R = -Z_o/2$ then $P_a/P_i = +26$

then $P_a/P_i = 14dB$

If $R = Z_o/2$ then $P_a/P_i = 0.888$

then $P_a/P_i = -0.5dB$

4.5.3 System Noise Performance

By definition the Noise Figure is given by the signal to noise ratio on the output to the signal to noise ratio on the input of the system.

S/N ratio on the input is given by:

$$\frac{S_i}{N_i} = \frac{S_i}{ktB}$$

S/N ratio on the output is given by:

$$\frac{S_o}{N_o} = \frac{GS_i}{GkTB + (F_{osc} - 1)kTB}$$

But Noise Figure of the system is defined as:

$$F = \frac{S_i N_o}{S_o N_i}$$

Therfore:

$$F = \frac{S_i(GkTB + (F_{osc} - 1)kTB)}{(S_i G + S_{osc})kTB}$$

Therefore:

$$F = \frac{S_i(G + F_{osc} - 1)}{S_i G + S_{osc}}$$

But S_{osc} should be ignored as it disappears in the difference signal.

$$F = 1 + \frac{(F_{osc} - 1)}{G}$$

This would imply that the overall noise figure of the system could be improved to 1 if the gain of the front end amplifier is increased to infinity. In reality this is not possible.

The noise figure of the oscillator at a particular frequency can be found as follows:

$$N_{osc} = M.S_{osc}$$

Where M is the noise to carrier ratio

$$M.S_{osc} = (F - 1)kT$$

$$F - 1 = \frac{MS_{osc}}{kB}$$

Therefore:

$$F = 1 + \frac{MS_{osc}}{kT} \qquad \text{(Where B = 1Hz)}$$

If M = -160dBC

$$S_{osc} = 25 \text{ mW}$$

KT = 4.14 x 10E - 21

<u>Therefore F = 605</u>

Say R = -100 and Z_o = 450

<u>Therefore G = 22</u>

This suggests that system noise figure is:

F = 27.7

Therefore NF = 14.4 dB

4 CONCLUSION

The author found that with carefully controlled experiments, those sorts of noise figures could not be realized. If, however, the noise injection was moved to the input of the negative resistance amplifier, then the G factor would disappear from the expression. The author was then unable to obtain a realistic figure for conversion loss.

The work done at Lancaster gives such noise figures but the author was unable to establish these. One further note is that the GUNN/IF interface will generally be a tuned circuit of some form and under negative resistance conditions, the very small coherent oscillations produced would be injection locked by the received signal and would appear to enhance the input signal. However, they do not actually contain the input information and hence cannot be regarded as amplified input signal. On the other hand they would appear to be such to a power meter measuring output power and would thus tend to enhance the apparent signal to noise ratio on the output.

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CHAPTER 5

CONCLUSION

5 Conclusion

The author has attempted to show the theoretical and practical aspects of using a GUNN diode millimeter wave source as both the transmitter and the receiver of a system. His findings can be summarized as follows:

5.1 What theory holds good for mixer operation

The author found that a theory based on the use of a Time Dependent Reflection Coefficient successfully demonstrates the conversion loss of the device. In addition the view that the device noise figure is due to added noise prior to the mixing process seems to correspond to the results measured in practice. The author found that he was able to measure noise figures of as low as 22dB but that the lower noise figures predicted by the researchers at Lancaster (2, 3, 4, 5) seemed to be optimistic. Noise figure seemed to be affected by among other things, source power, cavity Q and IF frequency.

Other factors affecting noise figure but to a much lower extent are IF matching Impedance, Bias Voltage and actual values of R_n and R_p. Conversion loss also affected system noise figure but only to a very small extent. It can be noted here that very repeatable noise figures were obtained and these noise figures readily suggest the use of the device for commercial applications.

5.2. As regards future applications

The device could readily be mounted in cheap waveguide structures such as Image Waveguide. This could readily lead to cheap Dopplar Radar or other forms of radar for such applications as vehicle safety or burglar alarms.

Harmonic mixing also seems to hold out promise of commercial use with definite advantages in certain areas such as the Electronic Distance Measuring Instrument where substantially reduced beamwidth could be obtained by changing the microwave path from the fundamental to say the third harmonic.

5.3 General conclusion

The S.M.O. concept appears to have very good commercial possibilities if its drawbacks in terms of range performance are taken into account. In fact the S.M.O. is now in commercial use by Plessey S.A. in their new Microwave Distance Measuring Instrument.

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CHAPTER 6

APPENDIX

6.1 <u>Computer program for storing the transfer function D.A.T.A. and generating the Time Dependant Reflection Coifficient.</u>

```
10 DEG
20 DIM T(201),R(361)
30 COM P(361)
40 T(0)=2
50 FOR J=1 TO 200
60 V=(J-1)/20
70 IF V<1.15 THEN T(J)=2
80 IF V>1.35 THEN T(J)=-100
90 IF V=1.15 THEN T(J)=1.32
100 IF V=1.2 THEN T(J)=.66
110 IF V=1.25 THEN T(J)=0
120 IF V=1.3 THEN T(J)=-33
130 IF V=1.35 THEN T(J)=-66
140 NEXT J
150 FOR J=0 TO 360
160 V=5+5*COS(J)
170 M=IP(20*V)
180 M=M/20
190 R(J)=T(M*20)
200 NEXT J
210 GCLEAR
220 SCALE 0,360,-150,10
230 XAXIS 0,20 @ YAXIS 180,5
240 PENUP
250 MOVE 0,R(0)
260 FOR J=1 TO 360
270 DRAW J,R(J)
280 NEXT J
290 FOR J=0 TO 360
300 P(J)=(R(J)-450)/(R(J)+450)
310 NEXT J
320 GCLEAR
330 SCALE 0,360,-2,0
340 PENUP
350 XAXIS 0,20 @ YAXIS 180,.2
360 PENUP
370 MOVE 0,P(0)
380 FOR J=1 TO 360
390 DRAW J,P(J)
400 NEXT J
410 PAUSE
420 CHAIN "HAR2"
430 END
```

6.2. <u>Computer program for doing the harmonic analysis.(Shortened version)</u>

```
 10 ! PROGRAM CALCULATES FOURIER
 20 ! COEFFICIENTS OF A PERIODIC
 30 ! WAVEFORM. WAVEFORM IS
 40 ! AS A LIST OF EVENLY SPACED
 50 ! POINTS.
 60 DIM S(400),C(400),Z(2,10)
 70 COM P(361)
 80 RAD
 90 N=361
100 T=1
110 D=T/(N-1)
120 S=0
130 PRINT ""
140 PRINT "FOURIER COMPONENTS (P
    EAK VALUES)"
150 PRINT "HARM  FREQ      A(N)
      B(N)    "
160 IMAGE D,2X,DD,2X,DD.DDE,2X,D
    D.DDE
170 F=1/T
180 W=F*2*PI
190 GOSUB 450
200 PRINT USING 160 ; 0,0,S
210 F1=0
220 FOR I=1 TO 4
230 T=0
240 F1=F1+F
250 W=2*PI*F1
260 C(1)=P(1)
270 S(1)=0
280 C=0
290 S=0
300 FOR R=2 TO N
310 T=T+D
320 C(R)=P(R)*COS(W*T)
330 C=C+(C(R)+C(R-1))/2*D
340 S(R)=P(R)*SIN(W*T)
350 S=S+(S(R)+S(R-1))/2*D
360 Z(1,I)=C @ Z(2,I)=S
370 NEXT R
380 C=C*2/T
390 S=S*2/T
400 H=(C^2+S^2)^.5
410 PRINT USING 160 ; I,F1,C,S
420 NEXT I
430 CHAIN "G1"
440 END
450 ! SUBROUTINE TO INTIGRATE FU
    NTION. TRAPIZOIDAL INTIGRATI
    ON
460 S=0
470 FOR J=2 TO N
480 S=S+(P(J-1)+P(J))/2
490 NEXT J
500 S=S/(N-1)
510 RETURN
```

6.3. Computer program for doing the harmonic analysis.(Full version)

```
 10 ! PROGRAM CALCULATES FOURIER
 20 ! COEFFICIENTS OF A PERIODIC
 30 ! WAVEFORM. WAVEFORM IS
 40 ! AS A LIST OF EVENLY SPACED
 50 ! POINTS.
 60 DIM R(100),S(100),C(100),Z(2
    ,10)
 70 DISP "INPUT WAVEFORM DATA "
 80 DISP "POINT BY POINT AS FOLL
    OWS"
 90 DISP "HOW MANY POINTS"
100 INPUT N
110 DISP "WHAT IS PERIOD OF WAVE
    FORM"
120 INPUT T
130 M=0
140 FOR J=1 TO N
145 DISP " ENTER POINT # ";J;
150 INPUT R(J)
160 A=R(J)
170 IF R(J)<0 THEN A=-R(J)
180 IF A>M THEN M=A
190 NEXT J
200 D=T/(N-1)
210 S=0
220 PRINT ""
230 PRINT "  N     TIME(S)     M
    AGNITUDE"
240 FOR J=1 TO N
250 PRINT USING 260 ; J,S,R(J)
260 IMAGE DDD,5X, D.DDE, 5X,DDDD
270 S=S+D
280 NEXT J
290 CLEAR
300 DISP "DATA CORRECT? Y/N"
310 INPUT F$
320 IF F$="Y" THEN 410
330 DISP "WHAT ITEM"
340 INPUT J
350 DISP "NEW VALUE"
360 INPUT R(J)
370 A=R(J)
380 IF R(J)<0 THEN A=-R(J)
390 IF A>M THEN M=A
400 GOTO 210
410 PRINT ""
420 GCLEAR
430 PENUP
440 SCALE 0,T,-M,M
450 XAXIS 0,T/10 @ YAXIS 0,M/10
460 PENUP
470 MOVE 0,0
480 S=0
490 FOR J=1 TO N
500 DRAW S,R(J)
510 S=S+D
520 NEXT J
530 COPY
540 GCLEAR
550 PRINT ""
```

```
560 PRINT "FOURIER COMPONENTS (P
    EAK VALUES)"
570 PRINT "HARM  FREQ   A(N)   B
    (N)   C(N)"
580 IMAGE DD,2X,D.DE,2X,D.DE,2X,
    D.DE,2X,D.DE
590 F=1/T
600 W=F*2*PI
610 GOSUB 840
620 PRINT USING 580 ; 0,0,S
635 F1=0
640 FOR I=1 TO 10
645 T=0
650 F1=F1+F
660 W=2*PI*F1
670 C(1)=R(1)
680 S(1)=0
690 C=0
700 S=0
710 FOR R=2 TO N
720 T=T+D
730 C(R)=R(R)*COS(W*T)
740 C=C+(C(R)+C(R-1))/2*D
750 S(R)=R(R)*SIN(W*T)
760 S=S+(S(R)+S(R-1))/2*D
765 Z(1,I)=C @ Z(2,I)=S
770 NEXT R
780 C=C*2/T
790 S=S*2/T
800 H=(C^2+S^2)^.5
810 PRINT USING 580 ; I,F1,C,S
820 NEXT I
830 END
840 ! SUBROUTINE TO INTIGRATE FU
    NTION. TRAPIZOIDAL INTIGRATI
    ON
850 S=0
860 FOR J=2 TO N
870 S=S+(R(J-1)+R(J))/2
880 NEXT J
885 S=S/(N-1)
890 RETURN
```

```
560 PRINT "FOURIER COMPONENTS (P
    EAK VALUES)"
570 PRINT "HARM  FREQ   A(N)   B
    (N)   C(N)"
580 IMAGE DD,2X,D.DE,2X,D.DE,2X,
    D.DE,2X,D.DE
590 F=1/T
600 W=F*2*PI
610 GOSUB 840
620 PRINT USING 580 ; 0,0,S
635 F1=0
640 FOR I=1 TO 10
645 T=0
650 F1=F1+F
660 W=2*PI*F1
670 C(1)=R(1)
680 S(1)=0
690 C=0
700 S=0
710 FOR R=2 TO N
720 T=T+D
730 C(R)=R(R)*COS(W*T)
740 C=C+(C(R)+C(R-1))/2*D
750 S(R)=R(R)*SIN(W*T)
760 S=S+(S(R)+S(R-1))/2*D
765 Z(1,I)=C @ Z(2,I)=S
770 NEXT R
780 C=C*2/T
790 S=S*2/T
800 H=(C^2+S^2)^.5
810 PRINT USING 580 ; I,F1,C,S
820 NEXT I
830 END
840 ! SUBROUTINE TO INTIGRATE FU
    NTION. TRAPIZOIDAL INTIGRATI
    ON
850 S=0
860 FOR J=2 TO N
870 S=S+(R(J-1)+R(J))/2
880 NEXT J
885 S=S/(N-1)
890 RETURN
```

CHAPTER 7

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