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**EFFECT OF TRANSISTOR SELECTION ON FEEDBACK  
VCO PERFORMANCE  
' ANALYTICAL - NUMERICAL ' STUDY**

**BY  
RAVINDER KUMAR**

A Thesis submitted in partial fulfilment of the  
requirements for the degree Master of Science,  
Major in Electrical Engineering,  
South Dakota State University, 1987.

EFFECT OF TRANSISTOR SELECTION ON FEEDBACK  
VCO PERFORMANCE  
' ANALYTICAL - NUMERICAL ' STUDY

This thesis is approved as a creditable and independent investigation by a candidate for the degree, Master of Science, and is acceptable for meeting the thesis requirements for this degree. Acceptance of this thesis does not imply that the conclusions reached by the candidate are necessarily the conclusions of the major department.

Thesis Adviser

Date

Head, Electrical Engineering

Date

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KUMAR

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# CHAPTER 1

## INTRODUCTION

Voltage controlled oscillator circuits are one of the most important circuits for mobile radio service in the frequency range of 800-1000 MHz. In recent years, they are being extensively used as a part of PLLs in FM detectors for receivers and as modulators for FM transmitters used in these mobile radios. They must have wide tuning range, high output power, linearity, low phase noise and circuit compactness. Since this is a boundary region in which discrete components fail to operate as simple lumped single-parameter elements, a designer has to make a careful choice in mixing aspects of discrete components and microwave methods and elements[1].

The primary objective of this research was to find the effect of two important parameters,  $f_T$  and  $C_c$ , of the transistor on the feedback line length, stub line length, tuning capacitor and external  $Q$  of the circuit, in designing a VCO.

VCOs in this frequency range are normally designed using either an FET or a BJT, as an active device. In the following chapters, the model based preliminary designing process of a VCO is explained in detail, using a BJT.

The first step in the process was to develop a better high frequency model of a BJT, since the existing

models were inadequate in explaining the high frequency behaviour of the BJT, especially the transit time and the nature of RC distribution, which cause a delay from emitter to collector. An extensive literature search resulted in a model which accounts for the nature of the RC distribution.

The second step was to verify the validity of the model. One way of verifying was to calculate scattering or admittance parameters of the transistor numerically from the model and compare them with the measured scattering or admittance parameters of the transistor, which are normally listed in the manufacturers data sheets.

The third step was to find the parameters of the basic oscillator circuit. The basic oscillator circuit has a BJT, a feedback transmission line to give a zero phase shift and a stub line to give zero output susceptance. Both lines are microstrip elements. The free-space equivalent lengths of the lines are found and used.

The fourth step was to incorporate a varactor for the tuning of the oscillator circuit. The varactor was tried at both input and output of the circuit, and placed for the best tuning range.

The fifth step was to calculate the tuning range and external Q of the circuit.

To fulfill the primary objective, all the above steps were repeated for several transistors having different  $f_T$  and  $C_c$  values.

## CHAPTER 2

### HIGH - FREQUENCY MODEL OF A TRANSISTER

The high - frequency model of a transistor is normally a hybrid pi model. Most authors e.g.[2,3] have used this model to explain the high frequency behaviour of a transistor. The schematic diagram of the model is shown in fig.2.1.

In fig.2.1 -

$r_{ce}$  is the finite output resistance between collector C and emitter E

$r_{b,c}$  is the feedback resistor between collector and internal base node B'.

$r_{bb'}$  is the base spreading resistance between external base node B and internal base node.

$r_{b'e}$  is the input resistance between internal base node and the emitter.

$C_c$  is the output capacitance between the collector and the internal base node.

$C_e$  is the input capacitance between the internal base node and the emitter.

Some of the approximations made in analysing the behaviour of this model are as follows[4] -

1) The input resistance  $r_{b'e}$  : It is assumed that  $r_{b,c}$  is much greater than  $r_{b'e}$ . Hence the base current  $I_b$

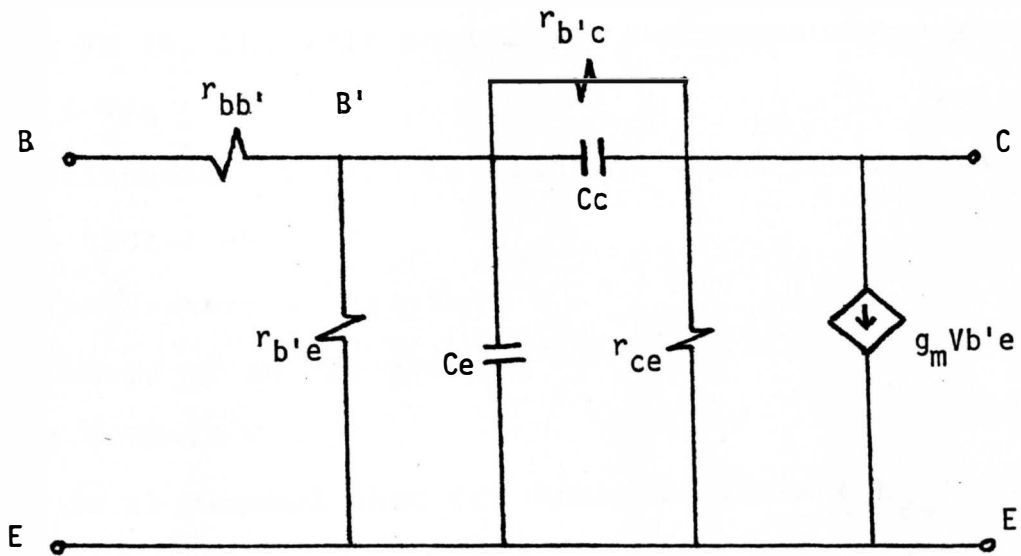


Fig.2.1 Hybrid - pi model

Since it is assumed that  $r_{b',c} \gg r_{b',e}$ , the input resistance is given by -

$$h_{ie} = r_{bb'} + r_{b',e}$$

or

$$\begin{aligned} r_{bb'} &= h_{ie} - r_{b',e} \\ &= h_{ie} - h_{fe} V_t / |I_c| \end{aligned}$$

4) The output resistor  $r_{ce}$ : This resistor is normally defined in terms of output conductance  $g_{ce}$ , with input open circuited. In terms of h - parameters it is given as  $h_{oe}$ .

Now for input open circuited, the collector current is given by -

$$I_c = V_{ce}/r_{ce} + V_{ce}/(r_{b',c} + r_{b',e}) + g_m V_{b'e} \quad \text{and}$$

$$h_{oe} = I_c/V_{ce} = (1/r_{ce}) + (1/(r_{b',c} + r_{b',e})) +$$

$$g_m h_{re}$$

Also if  $1/r_{ce} = g_{ce}$ ;  $1/r_{b',c} = g_{b',c}$ ;  $1/r_{b',e} = g_{b',e}$  and  $r_{b',c} \gg r_{b',e}$  then

$$h_{oe} = g_{ce} + g_{b',c} + g_m h_{re}$$

$$\text{But } g_m = g_{b',e} h_{fe} \quad \text{and} \quad h_{re} = g_{b',c} r_{b',e} = g_{b',c}/g_{b',e}$$

$$\text{Therefore } h_{oe} = g_{ce} + g_{b',c} + g_{b',e} h_{fe} g_{b',c}/g_{b',e}$$

$$\text{Hence } g_{ce} = h_{oe} - (1+h_{fe}) g_{b',c}$$

5) The output capacitance  $C_c$ : This is the measured common base output capacitance with the input open. Since in the active region, the collector junction is reverse biased,  $C_c$  is defined as the transition capacitance. The value of  $C_c$  is normally specified by the manufacturer.

6) The capacitance  $C_e$  : This represents the sum of the emitter diffusion capacitance  $C_{de}$  and the emitter junction capacitance  $C_{te}$ .  $C_{de}$  is usually larger than  $C_{te}$  for a forward - biased emitter junction. Hence

$$C_e = C_{de} + C_{te} \approx C_{de} = C_{ib}$$

where  $C_{ib}$  is the measured common base input capacitance with the output open.

With all these approximations it is very difficult to explain the high - frequency behaviour of a transistor accurately.

The two most important factors that are not explained by this model are the delay time from emitter to collector called the transit time and the nature and distribution of  $r_{bb'}$ , and  $C_c$ . These phenomena cause phase shift in excess of the (max.) 90 obtainable from  $r_{bb'}$ ,  $C_e$ . This is very important in an oscillator which depends on phase shift for its design.

One of the models that could explain the transit time as well as the distribution of  $r_{bb'}$ , and  $C_c$  is shown in fig.2.2 [5].

This model shows a discrete approximation to the distributed RC transmission line which the base current has to move through before reaching the active region of the transistor, which is the emitter periphery. However the effective values of  $r_{bb'}$ , and  $C_c$  are important to



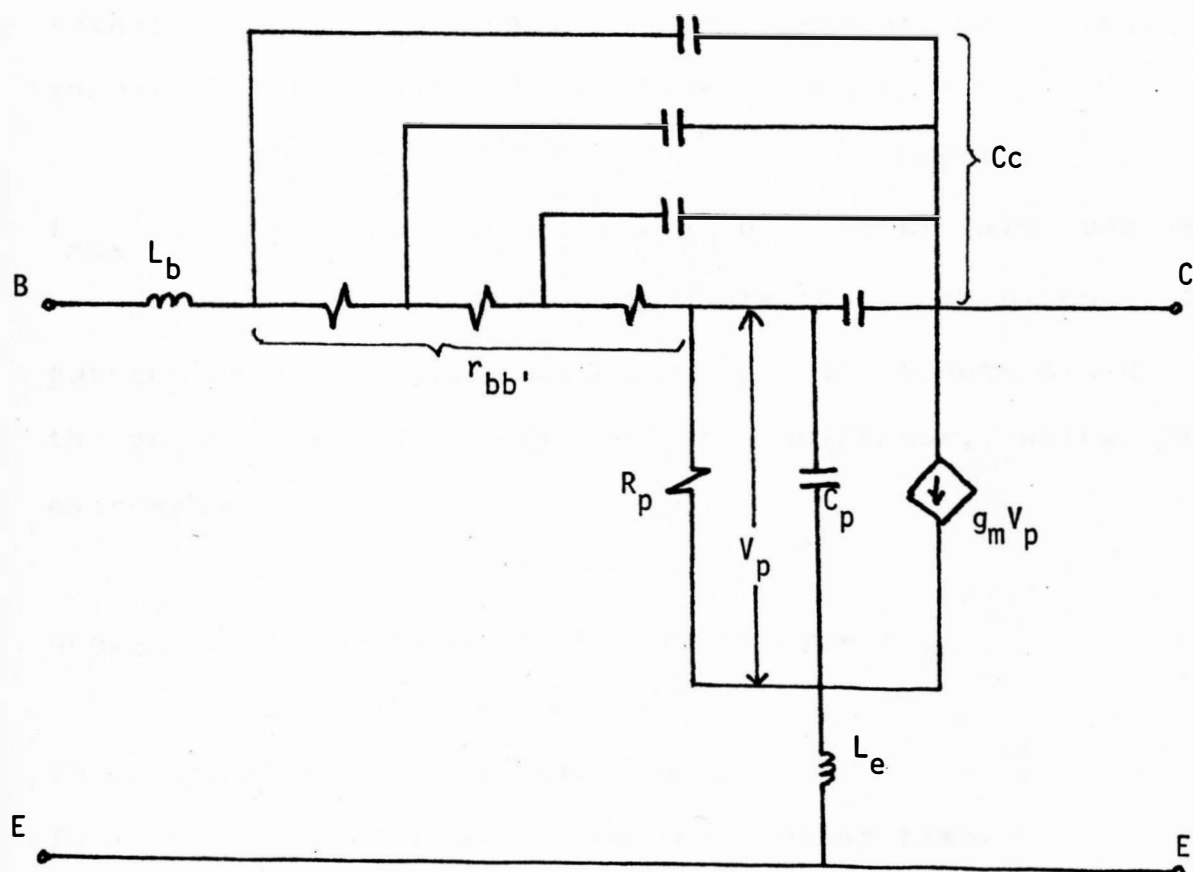


Fig.2.2 Model with base-collector delay line and lead inductances.

explain the performance of the transistor. This model also includes the bonding inductances to the base and emitter. These inductances become important at very high radio frequencies, which will be shown in the later chapters. There are two figures of merit for the bipolar transistor which are related in terms of the effective values of the RC distributed line, given by -

$$f_{\max}^2 = f_T / (8 \pi r_{bb} C_c) \quad \text{where}$$

$f_{\max}$  is the frequency at which unilateral gain becomes unity and  $f_T$  is the frequency where the short circuit CE current gain  $h_{fe}$  approximates unity. It is determined by the delay time from emitter to collector, using the expression

$$f_T = 1 / (2 \pi T_{ec})$$

where  $T_{ec}$  is the transit time given by -

$$T_{ec} = T_b + T_d \quad \text{where}$$

$T_b$  is the base transit time and

$T_d$  is the collector depletion layer delay time.

Though this model explains the performance of the transistor to a larger extent, it still does not account for the parasitic resistances associated with the leads. These resistances do play an important role in the performance of the transistor. A model which is developed based on the model shown in fig.2.2 and takes the parasitic resistances into account is shown in fig.2.3.

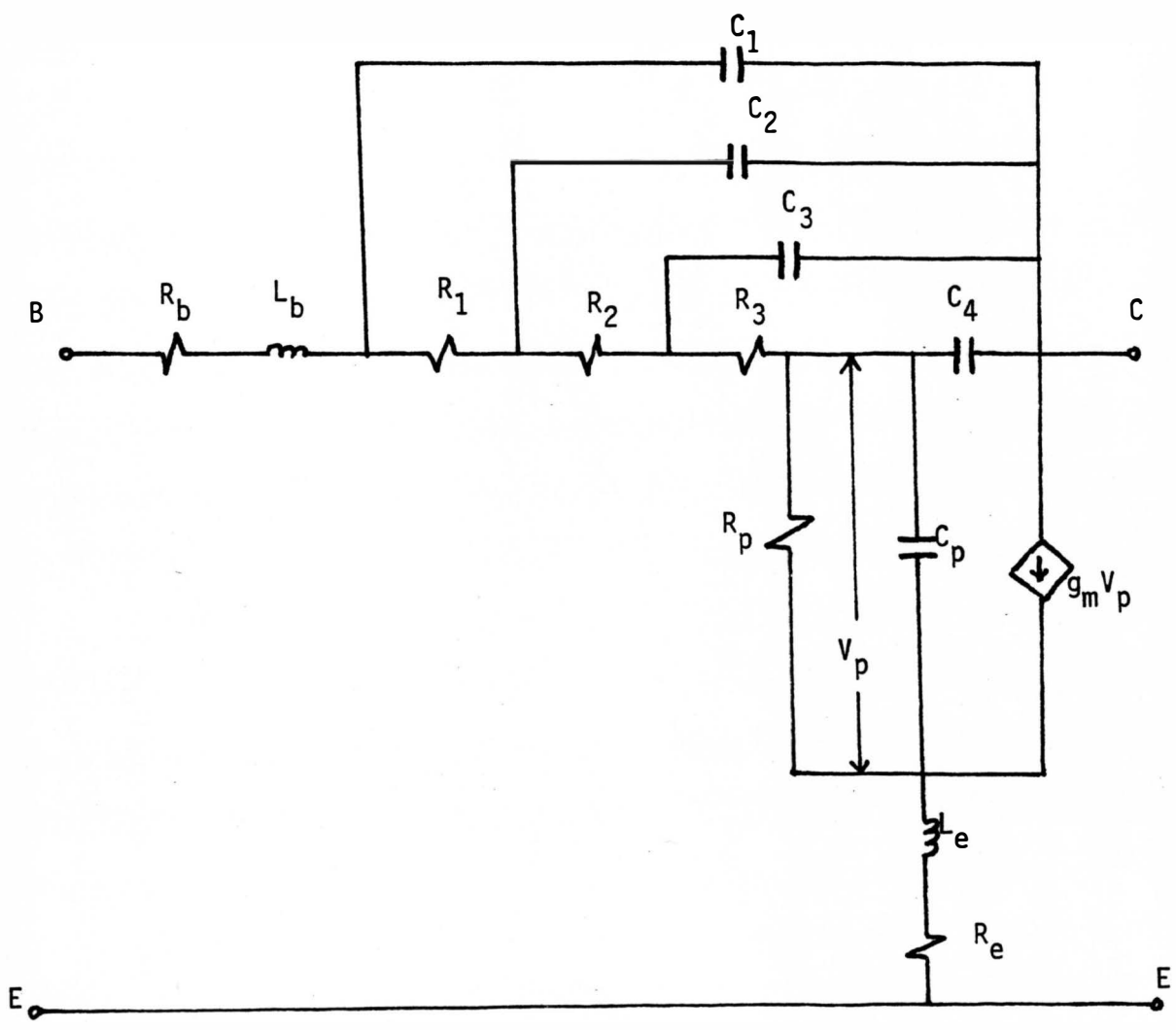


Fig.2.3 Model with base-collector delay line, lead inductances and parasitic resistances.

The importance of these resistances will be shown in later chapters.

## CHAPTER 3

### PARAMETERS OF A TWO PORT NETWORK

#### 1) SCATTERING PARAMETERS

A general block diagram of a two port network is shown in fig.3.1. In the figure port 1 is the input port and port 2 is the output port,  $a_1$  is the incident wave at port 1 and  $b_1$  is the reflected wave at port 1,  $a_2$  is the incident wave at port 2 and  $b_2$  is the reflected wave at port 2. In terms of these incident and reflected waves the Scattering parameters or S parameters of a two port network are given by :

$$b_1 = S_{11} a_1 + S_{12} a_2 \quad (3.1)$$

$$b_2 = S_{21} a_1 + S_{22} a_2 \quad (3.2)$$

From the above set of equations, the different S parameters can be defined as follows :

$S_{11} = b_1/a_1$  ;  $a_2 = 0$ , is the input reflection coefficient with output properly terminated.

$S_{21} = b_2/a_1$  ;  $a_2 = 0$ , is the forward transmission coefficient with output properly terminated.

$S_{22} = b_2/a_2$  ;  $a_1 = 0$ , is the output reflection coefficient with input properly terminated.

$S_{12} = b_1/a_2$  ;  $a_1 = 0$ , is the reverse transmission coefficient with input properly terminated.

It can be seen from the above definitions of the S parameters that these parameters are measured using a

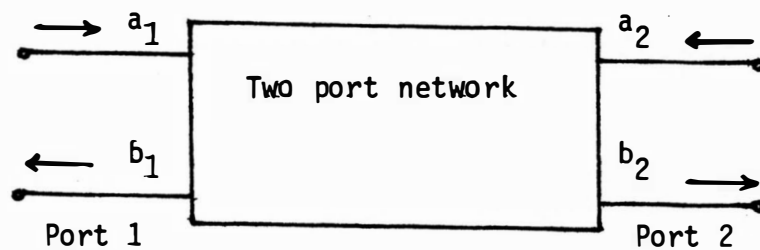


Fig.3.1 Block diagram of a two port network to calculate S parameters.

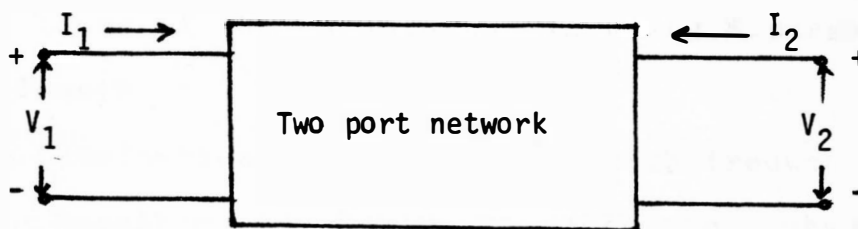


Fig.3.2 Block diagram to calculate Y parameters.

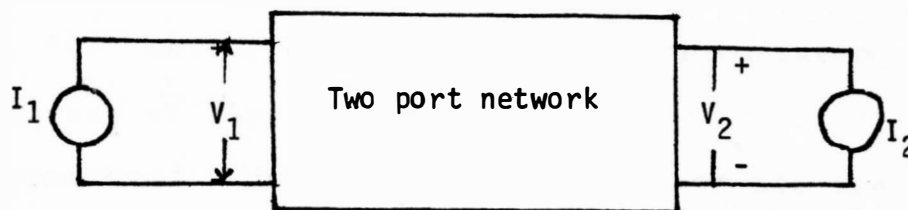


Fig.3.3 Block diagram to calculate Z parameters.

matched termination. For example, to measure  $S_{11}$ , the ratio  $b_1/a_1$  is measured at the input port with the output port properly terminated with an impedance equal to the characteristic impedance of the transmission line. This produces  $a_2 = 0$  since a travelling wave incident on the load will be totally absorbed and no energy will be returned to the output port. Similarly, to measure  $S_{22}$ , the input port is properly terminated with an impedance equal to the characteristic impedance of the transmission line which produces  $a_1 = 0$ .

Some of the advantages in using S parameters are as follows[6] :

- a) The termination is accurate at high frequencies since it is possible to build an accurate characteristic impedance load.
- b) No tuning is required to terminate a device in the characteristic impedance of the system because, if a characteristic impedance load is placed at the end of the line, the device will see the characteristic impedance regardless of line length.
- c) Broad band swept frequency measurements are possible because the device will remain terminated in the characteristic impedance as frequency changes.
- d) The termination enhances stability since it provides a power-absorbing resistive termination that stabilizes many negative resistance devices.

The S parameters are normally used to measure high frequency parameters of the transistor. For measuring the low frequency parameters, admittance, hybrid or impedance parameters are normally used. In the following sections admittance and impedance parameters are explained briefly. It is also explained why they are not suitable for measuring the high frequency parameters.

## 2) ADMITTANCE PARAMETERS

The admittance parameters or Y parameters of a two port network can be derived from the block diagram shown in fig.3.2. In deriving the Y parameters, the voltages are made independent variables. Hence the dependent currents are given by :

$$I_1 = Y_{11} V_1 + Y_{12} V_2 \quad (3.3)$$

$$I_2 = Y_{21} V_1 + Y_{22} V_2 \quad (3.4)$$

From this set of equations the Y parameters are defined as follows :

$Y_{11} = I_1/V_1$  ;  $V_2 = 0$ , is the short circuit input admittance.

$Y_{12} = I_1/V_2$  ;  $V_1 = 0$ , is the short circuit reverse transfer admittance.

$Y_{21} = I_2/V_1$  ;  $V_2 = 0$ , is the short circuit forward transfer admittance.

$Y_{22} = I_2/V_2$  ;  $V_1 = 0$ , is the short circuit output admittance.



Although these parameters are easy to use in theoretical calculations, they are not very easy to measure above 400 MHz. From the above definitions it is clear that, to measure any of the Y parameters, either the input port or the output port has to be short circuited, which is achieved by placing a large capacitance across the terminals of the port[7]. This method of obtaining a short circuit is reasonably accurate as long as the circuit impedance is high such as for a reverse biased collector base circuit, but it is not so good if the circuit impedance is low such as for a forward biased emitter base circuit. Also, getting a short that does not look inductive, and has a well defined location is more difficult as the test frequency is raised.

### 3) IMPEDANCE PARAMETERS

The impedance parameters or Z parameters of a two port network can be derived from the block diagram shown in fig.3.3. The currents are made independent variables. Hence the dependent voltages are given by :

$$V_1 = Z_{11} I_1 + Z_{12} I_2 \quad (3.5)$$

$$V_2 = Z_{21} I_1 + Z_{22} I_2 \quad (3.6)$$

From the above set of equations, the Z parameters can be defined as follows :

$Z_{11} = V_1/I_1$  ;  $I_2 = 0$ , is the open circuit input impedance.

$Z_{12} = V_1/I_2$  ;  $I_1 = 0$ , is the open circuit reverse transfer impedance.

$Z_{21} = V_2/I_1$  ;  $I_2 = 0$ , is the open circuit forward transfer impedance.

$Z_{22} = V_2/I_2$  ;  $I_1 = 0$ , is the open circuit output admittance.

These parameters can also be calculated accurately in theory but for practical measurements we have the same disadvantages as noted in the remarks on Y parameters. That is, for measurements of open circuit impedance parameters, an open circuit is required for the signal currents but not the dc currents since the device must be properly biased. Hence an open circuit for signal current is obtained by inserting a large inductance in series with the circuit to be opened[8]. This method is adequate if the circuit impedance is already small before the inductance is added as in the case of forward biased emitter base junction but this condition is hardly satisfied for a reverse biased collector base junction whose impedance is very high.

### S PARAMETERS OF A MICROWAVE TRANSISTOR

The performance of most microwave transistors is given in terms of S parameters. Hence in this section the validity of the high frequency model is tested by

calculating numerically the S parameters of the transistor and comparing them with the published data.

The model is the same as shown in fig.2.3 and is repeated for convenience in fig.3.4. In the figure R1, R2, R3, C1, C2, C3 and C4 describes the RC distribution. I1 and I2 are the independent current sources and gm Vp is the dependent current source.

Performing the nodal analysis the following equations are obtained for the currents entering or leaving the nodes, in the vector-matrix form :

$$\begin{bmatrix} I1 \\ I2 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} M \end{bmatrix} \begin{bmatrix} V1 \\ V2 \\ V3 \\ V4 \\ V5 \\ V6 \\ V7 \end{bmatrix}$$

where M is a 7 x 7 matrix given by :

$$\begin{bmatrix} 1/(Rb+jwLb) & 0 & -1/(Rb+jwLb) & 0 & 0 & 0 & 0 \\ 0 & (1/R4+jw(C1+C2+C3+C4)) & -jwC1 & -jwC2 & -jwC3 & gm-jwc4 & -gm \\ -1/(Rb+jwLb) & -jwC1 & (1/(Rb+jwLb)+1/R1+jwC1) & -1/R1 & 0 & 0 & 0 \\ 0 & -jwC2 & -1/R1 & (1/R1+1/R2+jwC2) & -1/R2 & 0 & 0 \\ 0 & -jwC3 & 0 & -1/R2 & (1/R1+1/R2+jwC3) & -1/R3 & 0 \\ 0 & -jwC4 & 0 & 0 & -1/R3 & (1/R3+1/Rp+jw(Cp+C4)) & (-1/Rp-jwCp) \\ 0 & 0 & 0 & 0 & (-1/Rp-jwCp-gm) & (1/Rp+gm+jwCp+1/(Re+jwLe)) \end{bmatrix}$$

In terms of independent currents, the dependent voltages can be defined as follows :

$$\begin{bmatrix} V1 \\ V2 \\ V3 \\ \cdot \\ \cdot \\ V7 \end{bmatrix} = \begin{bmatrix} M \end{bmatrix}^{-1} \begin{bmatrix} I1 \\ I2 \\ 0 \\ \cdot \\ \cdot \\ 0 \end{bmatrix}$$

From earlier definitions it can be seen that the first two rows of the above equation are equivalent to the Z parameter definitions and  $[ M ]^{-1}$  gives the Z parameter values in its first two rows and columns.

If  $[ MI ] = [ M ]^{-1}$  then

$$MI[ 1, 1 ] = Z11$$

$$MI[ 1, 2 ] = Z12$$

$$MI[ 2, 1 ] = Z21$$

$$MI[ 2, 2 ] = Z22$$

It can also be seen that the Z parameters are independent of the current sources. However they are a function of frequency since capacitors and inductors are involved. Hence the Z parameters are complex quantities and should be defined in complex form.

To evaluate any parameter of a transistor the following values have to be known :

1) Output capacitance  $C_c$ .

- 2) The base spreading resistance  $r_{bb'}$ .
- 3) The collector bias current  $I_c$ .
- 4) Small signal current gain  $h_{fe}$ .
- 5) The frequency at which common emitter current gain reduces to unity  $f_T$ .

All these values for a given transistor are normally specified in the manufacturers' data sheet.

Two of the above mentioned values  $C_c$  and  $r_{bb'}$ , are the effective values which can not be used directly. It can be seen in fig.3.4 that  $C_c$  and  $r_{bb'}$ , are distributed. The present work uses an empirical rule, based on Vendelin's model [9], to account for the distributive nature of  $C_c$  and  $r_{bb'}$ . It is given as follows :

$$C1 = 0.294643 C_c$$

$$C2 = 0.3125 C_c$$

$$C3 = 0.25 C_c$$

$$C4 = 0.142857 C_c$$

$$R1 = 0.02469 r_{bb'}$$

$$R2 = 0.4321 r_{bb'}$$

$$R3 = 0.54321 r_{bb'}$$

For a given collector current in mA, the transconductance of the transistor  $g_m$  can be calculated from  $g_m = |I_c| / 26$  mhos

From the value of  $h_{fe}$ , the value of  $R_p$  can be calculated as  $R_p = h_{fe} \times r_e$ , where  $r_e = 1/g_m$ .

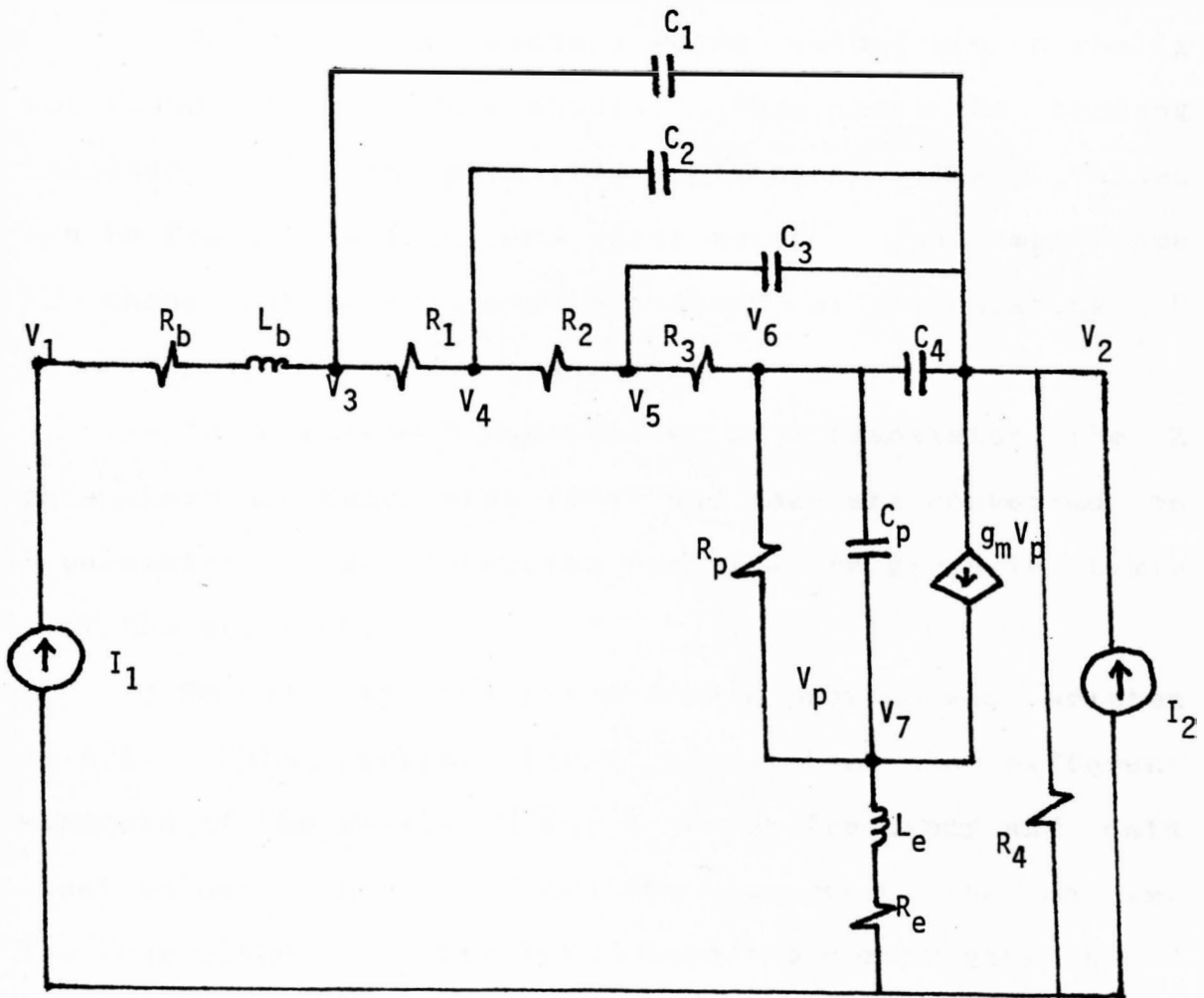


Fig.3.4 A complete model of a transistor showing lead inductances, parasitic resistances and RC distribution.

From the value of  $f_T$ , the value of  $C_p$  can be calculated as  $C_p = 1/r_e$ ,  $w_T = 1/(2\pi f_T r_e)$ .

Hence it can be seen that from the data sheet values, most of the values can be calculated.

There are two factors whose values are normally not found in the data sheets. They are the bonding inductances and the parasitic resistances. These values can be found by a trial and error method. The importance of these parameters can be shown by calculating S parameters of a typical transistor.

To calculate S parameters of a transistor, the Z parameters are calculated first and then are converted to S parameters. The conversion formulas are given in table 1 of the appendix.

To calculate the parameters a program was written in APL. This program first calculates the different elements of the matrix M for a given frequency and data sheet values. Then it finds the inverse of the matrix. The four elements in the left hand top corner give the Z parameters. The program also converts the Z parameters to S parameters both in rectangular and polar forms.

To verify the validity of the model, four transistors were picked whose S parameters or Y parameters were given by the manufacturers. These parameter values were measured values. The measured and calculated values of each transistor are given as follows :

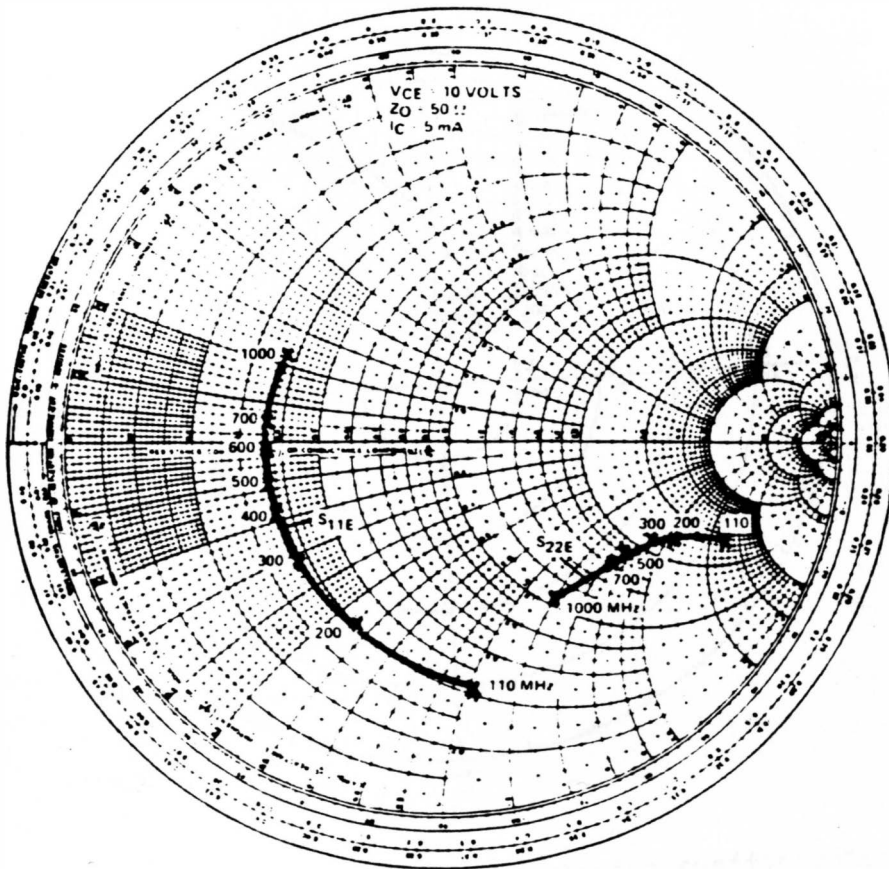


Fig.3.5 Input and output reflection coefficients of transistor A as a function of frequency.  
 ( measured values )



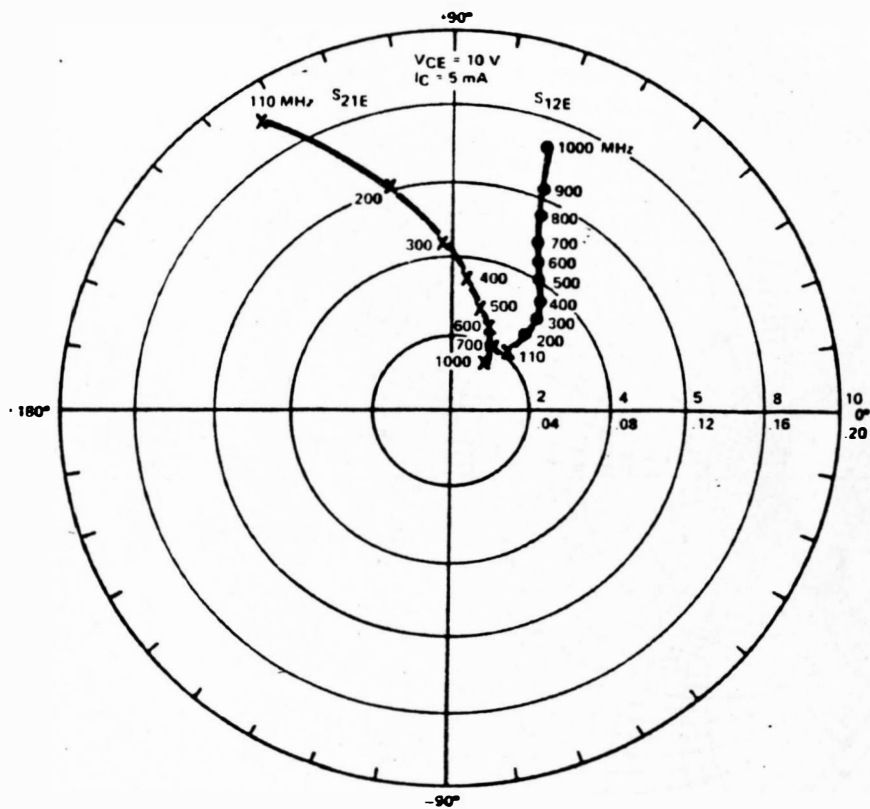


Fig.3.6 Forward and reverse transmission coefficients of transistor A as a function of frequency.  
 ( measured values )

NAME	TITLE	DWG. NO.
SMITH CHART FORM 82B3PR (2-49)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.	DATE

Supersedes G.R. Form 5301 7560 N

IMPEDANCE OR ADMITTANCE COORDINATES

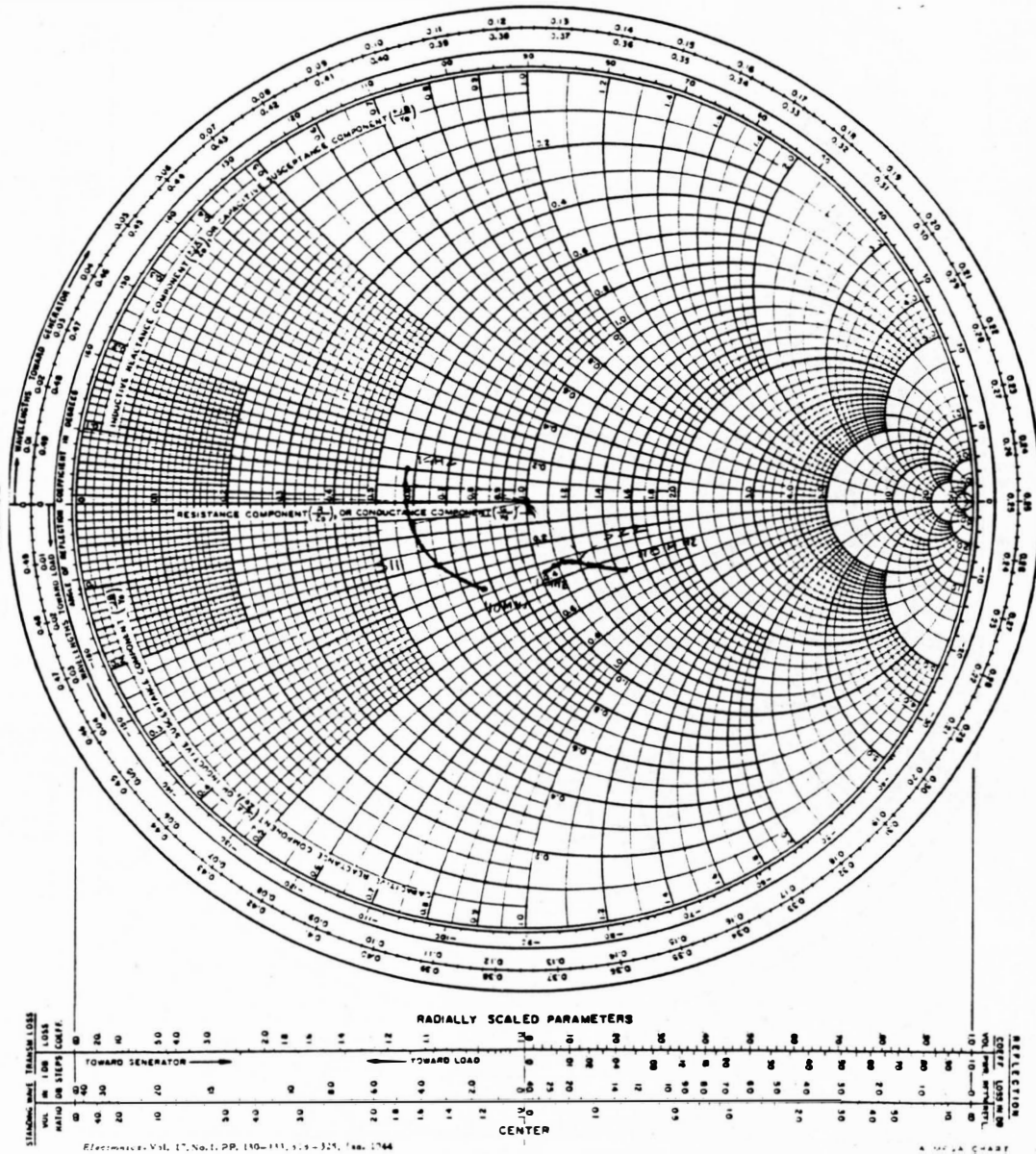


Fig.3.7 Input and output reflection coefficients of transistor A as a function of frequency. (calculated values)

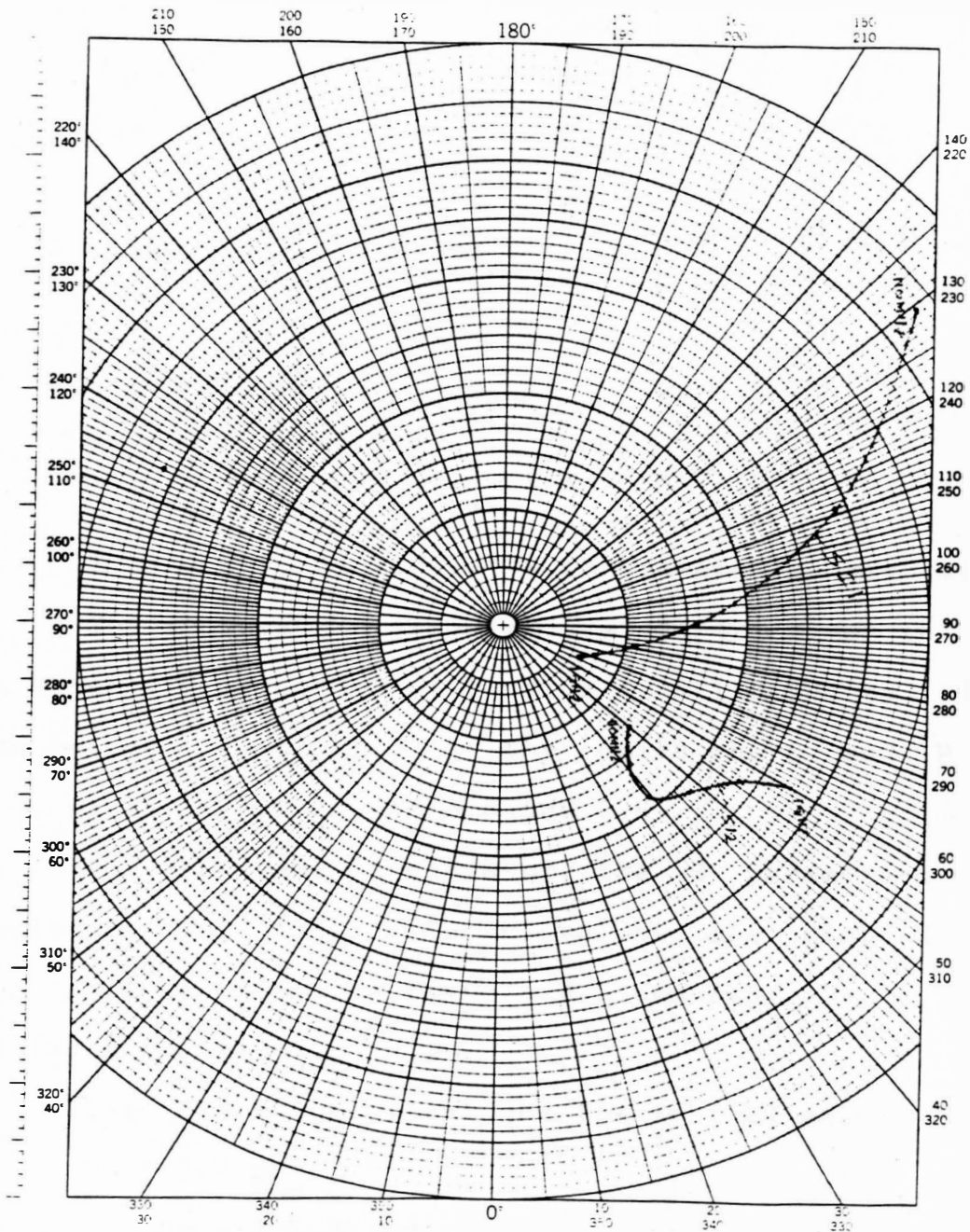


Fig.3.8 Forward and reverse transmission coefficients of transistor A as a function of frequency.  
( calculated values )

**TRANSISTOR A [10]**

Data sheet specifications for  $I_c = 5 \text{ mA}$ .

$$C_c = 1.7 \text{ pF}$$

$$r_{bb}, C_c = 5.7473 \text{ ps}$$

$$h_{fe} = 20$$

$$f_T = 1.3 \text{ GHz.}$$

$$L_e = 1.5 \text{ nH}$$

$$L_b = 1.5 \text{ nH}$$

$$R_e = 3.5 \text{ ohms}$$

$$R_b = 3.5 \text{ ohms}$$

Maximum power = 450 mW

Type - NPN, MA 42001-509 (Microwave Associates)

The measured S parameter values are given in the data sheet, for frequencies ranging from 110 MHz to 1 GHz. These values are plotted on the Smith chart and Polar chart, as shown in figures 3.5 and 3.6.

For the same transistor, the S parameters are calculated numerically. These values are plotted for comparison, as shown in figures 3.7 and 3.8.

**TRANSISTOR B [11]**

Data sheet specifications for  $I_c = 20 \text{ mA}$ .

$$C_c = 1.7 \text{ pF}$$

$$r_{bb}, C_c = 1.7507 \text{ ps}$$

$$h_{fe} = 70$$

$$f_T = 4.4 \text{ GHz.}$$

$$L_e = 2.8 \text{ nH}$$

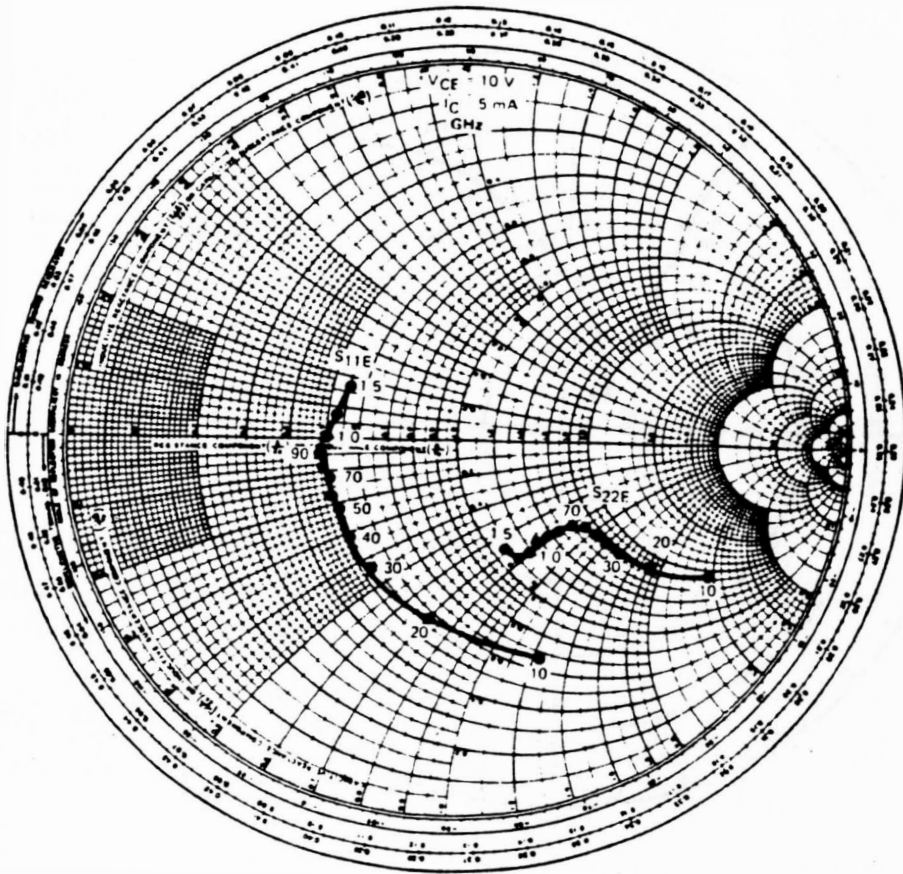


Fig.3.9 Input and output reflection coefficients of transistor B as a function of frequency.  
( measured values )

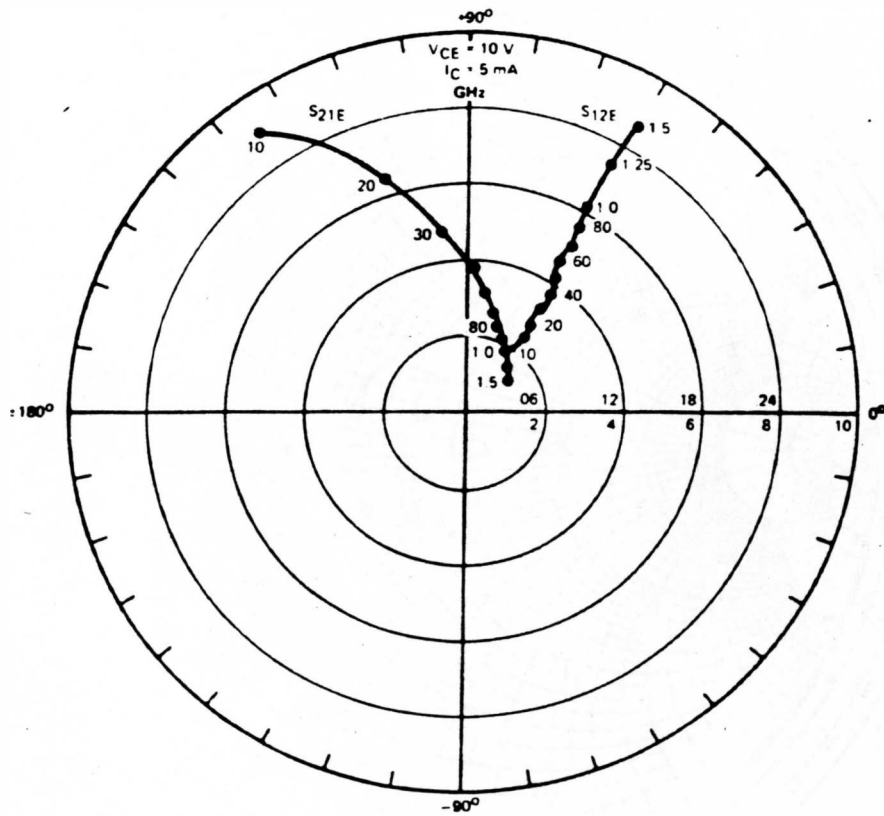


Fig.3.10 Forward and reverse transmission coefficients of transistor B as a function of frequency.  
 ( measured values )

NAME	TITLE	DWG. NO.
SMITH CHART FORM 8285PR (2-49)	RAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.	DATE

Supersedes G. R. Form 5301 7560 N

IMPEDANCE OR ADMITTANCE COORDINATES

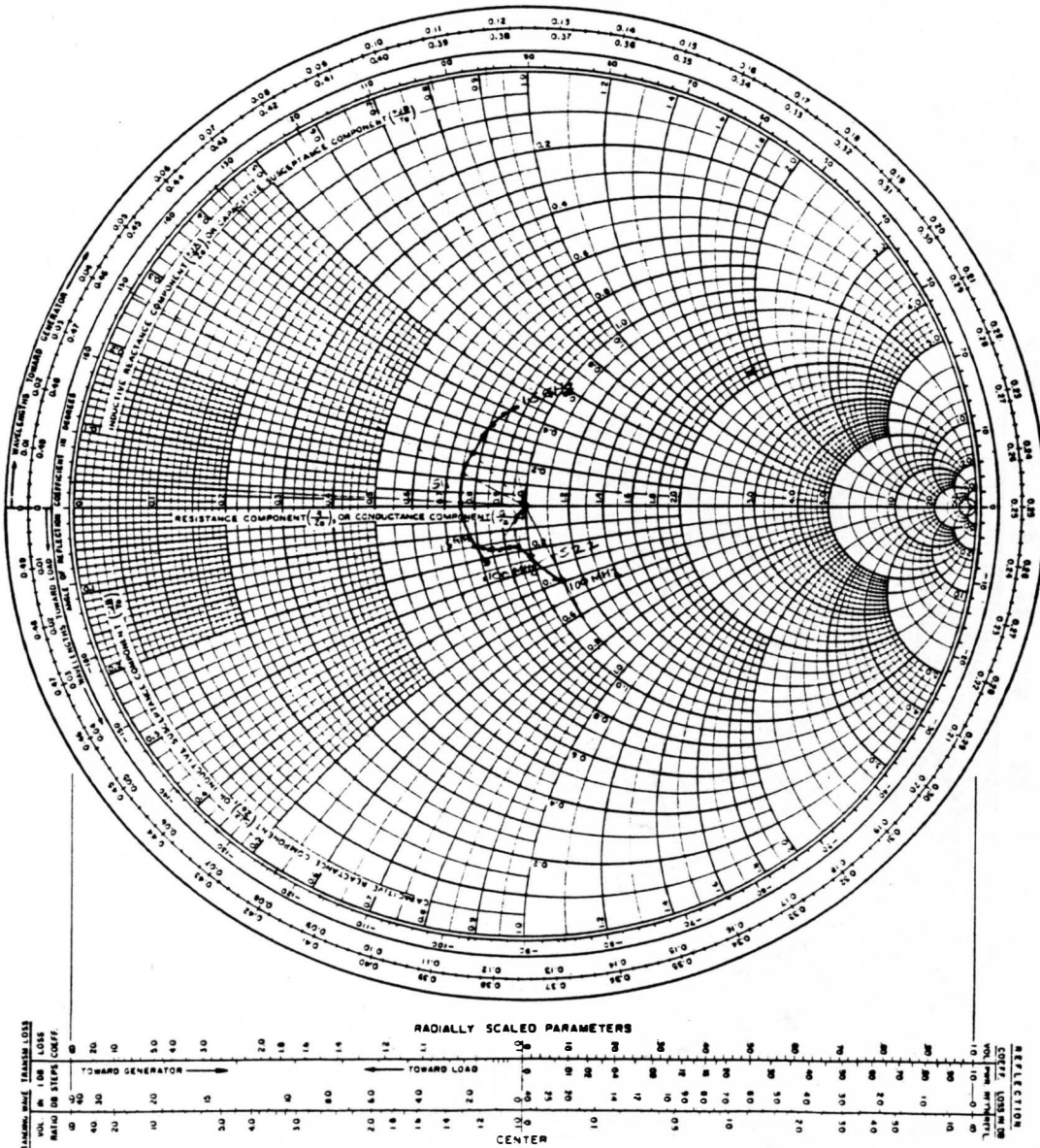


Fig.3.11 Input and output reflection coefficients of transistor B as a function of frequency. (calculated values)

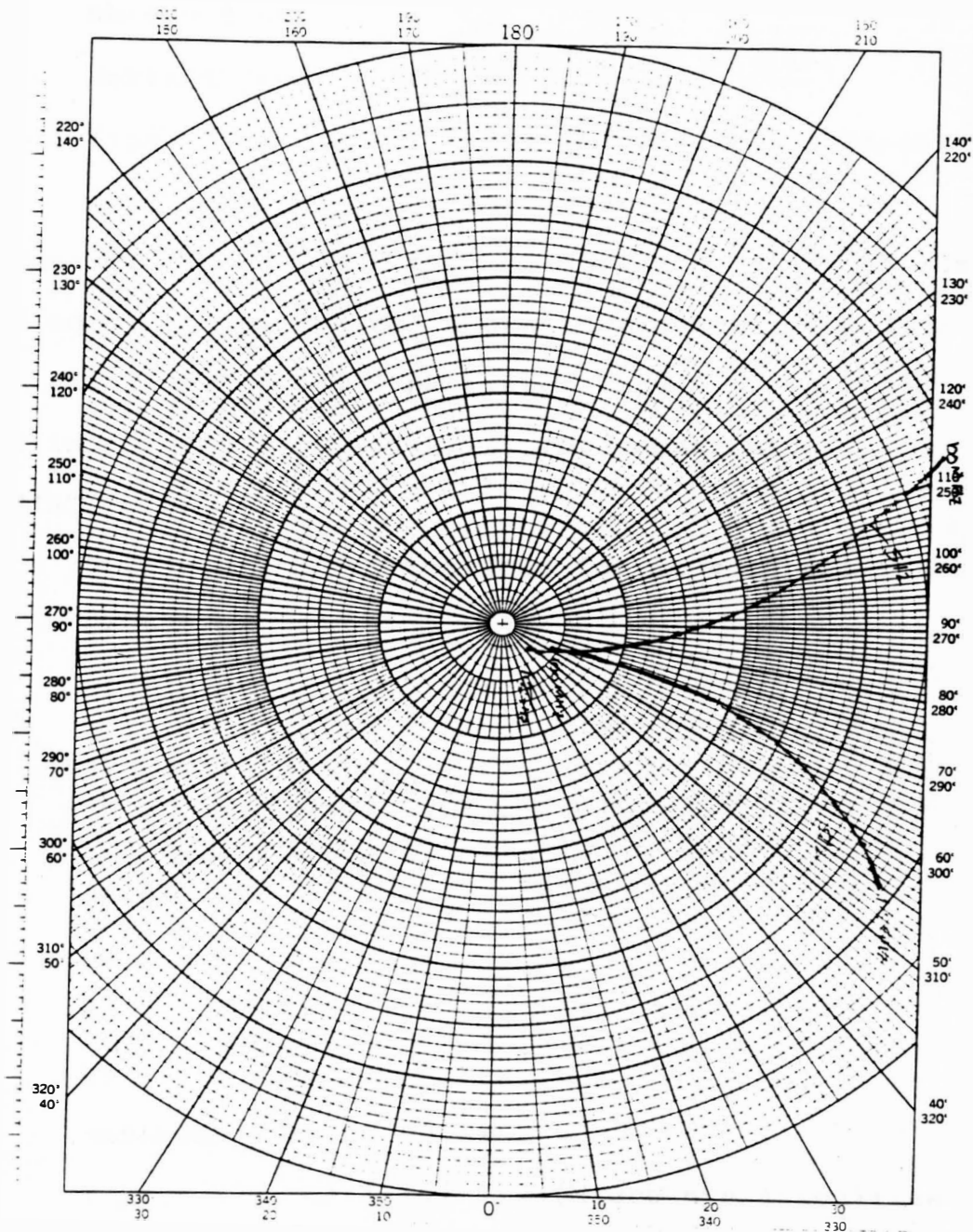


Fig.3.12 Forward and reverse transmission coefficients of transistor B as a function of frequency. ( calculated values )



$$L_b = 2.8 \text{ nH}$$

$$R_e = 0.8 \text{ ohms}$$

$$R_b = 0.8 \text{ ohms}$$

$$\text{Maximum power} = 450 \text{ mW}$$

Type - NPN, MA 42111-509 (Microwave Associates)

The measured S parameter values are for frequencies from 110 MHz to 1.5 GHz. The plots on Smith chart and Polar chart are shown in figures 3.9 and 3.10. Numerically calculated S parameter values are plotted and shown in figures 3.11 and 3.12, for comparison.

#### TRANSISTOR C [12]

Data sheet specifications for  $I_c = 15 \text{ mA}$ .

$$C_c = 1 \text{ pF}$$

$$r_{bb'} C_c = 1.9649 \text{ ps}$$

$$h_{fe} = 20$$

$$f_T = 4 \text{ GHz.}$$

$$L_e = 1 \text{ nH}$$

$$L_b = 1 \text{ nH}$$

$$R_e = 1.5 \text{ ohms}$$

$$R_b = 1.5 \text{ ohms}$$

$$\text{Maximum power} = 400 \text{ mW}$$

Type - NPN, MA 42141-510 (Microwave Associates)

The plots of measured values of S parameters for frequencies from 400 MHz to 5000 MHz are shown in figures 3.13 and 3.14. The plots of numerically calculated S

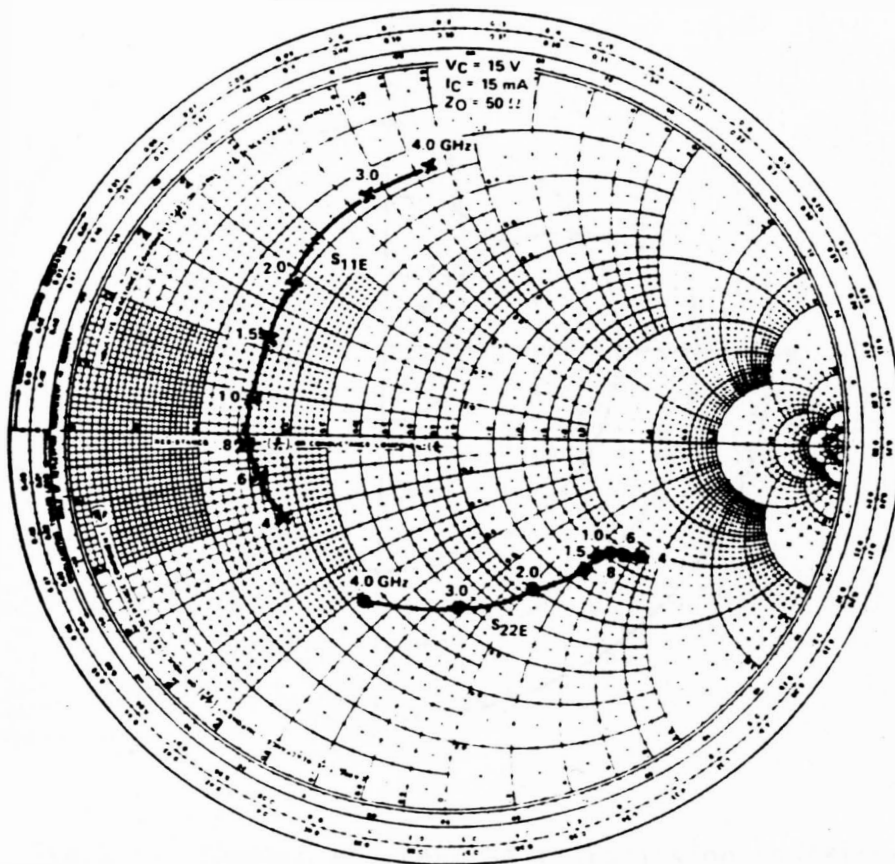


Fig.3.13 Input and output reflection coefficients of transistor C as a function of frequency.  
( measured values )

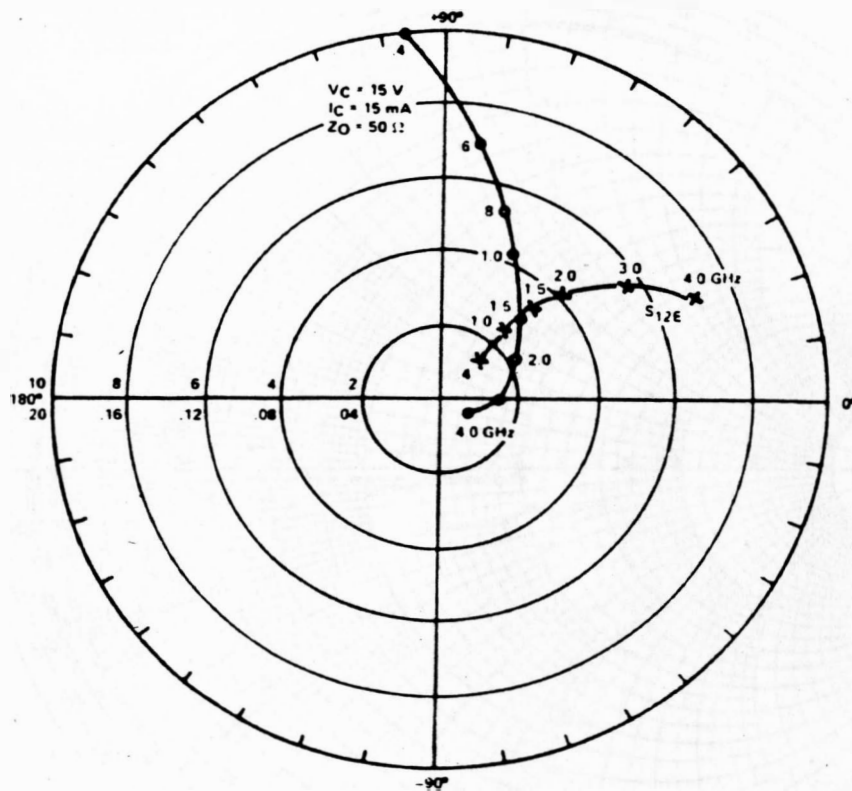


Fig.3.14 Forward and reverse transmission coefficients of transistor C as a function of frequency. ( measured values )

NAME	TITLE	DWG. NO.
SMITH CHART FORM 8253PR (2-49)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1949 PRINTED IN U.S.A.	DATE

Supersedes G. R. Form 5301-7560 N

IMPEDANCE OR ADMITTANCE COORDINATES

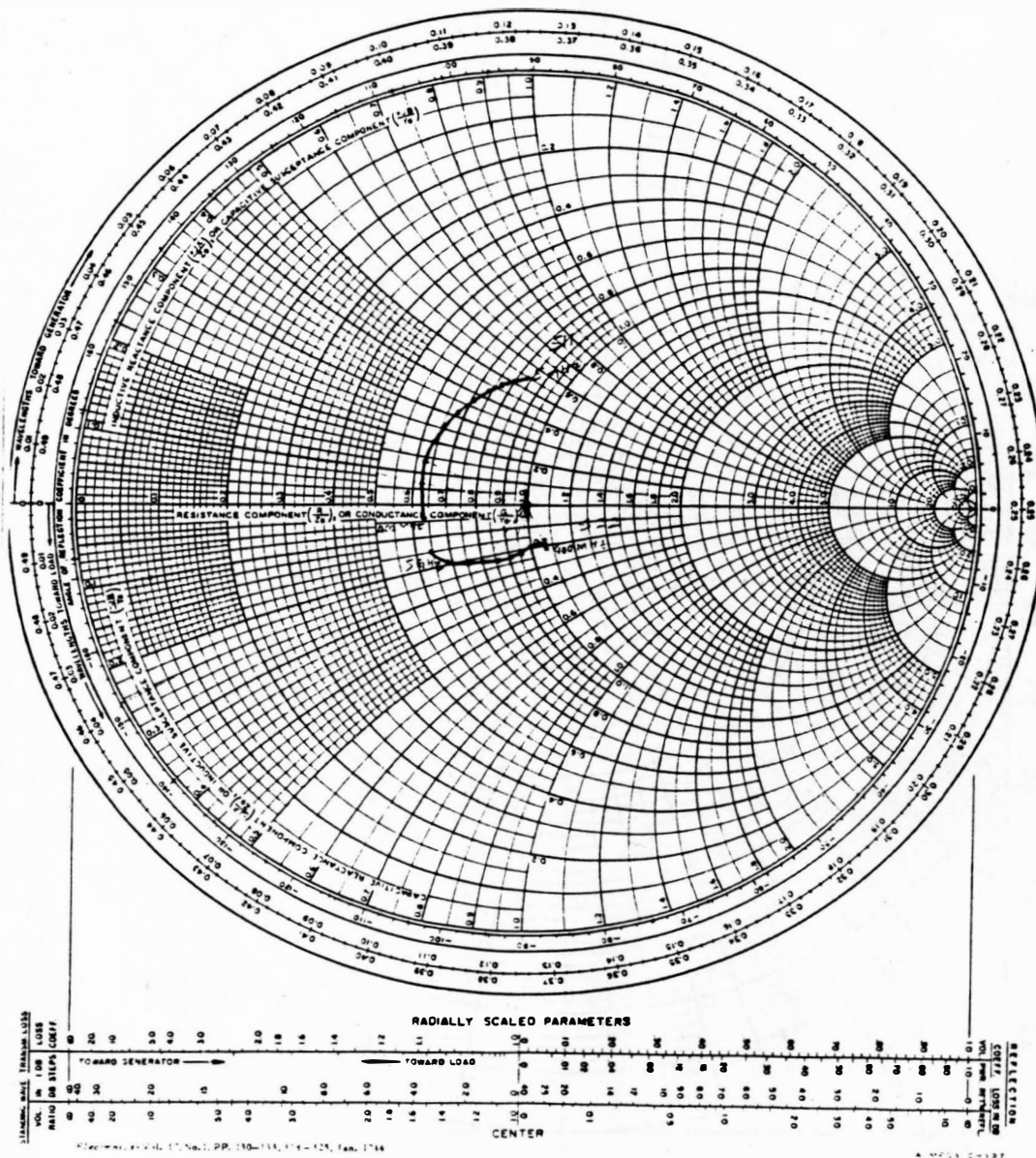


Fig.3.15 Input and output reflection coefficients of transistor C as a function of frequency.  
( calculated values )

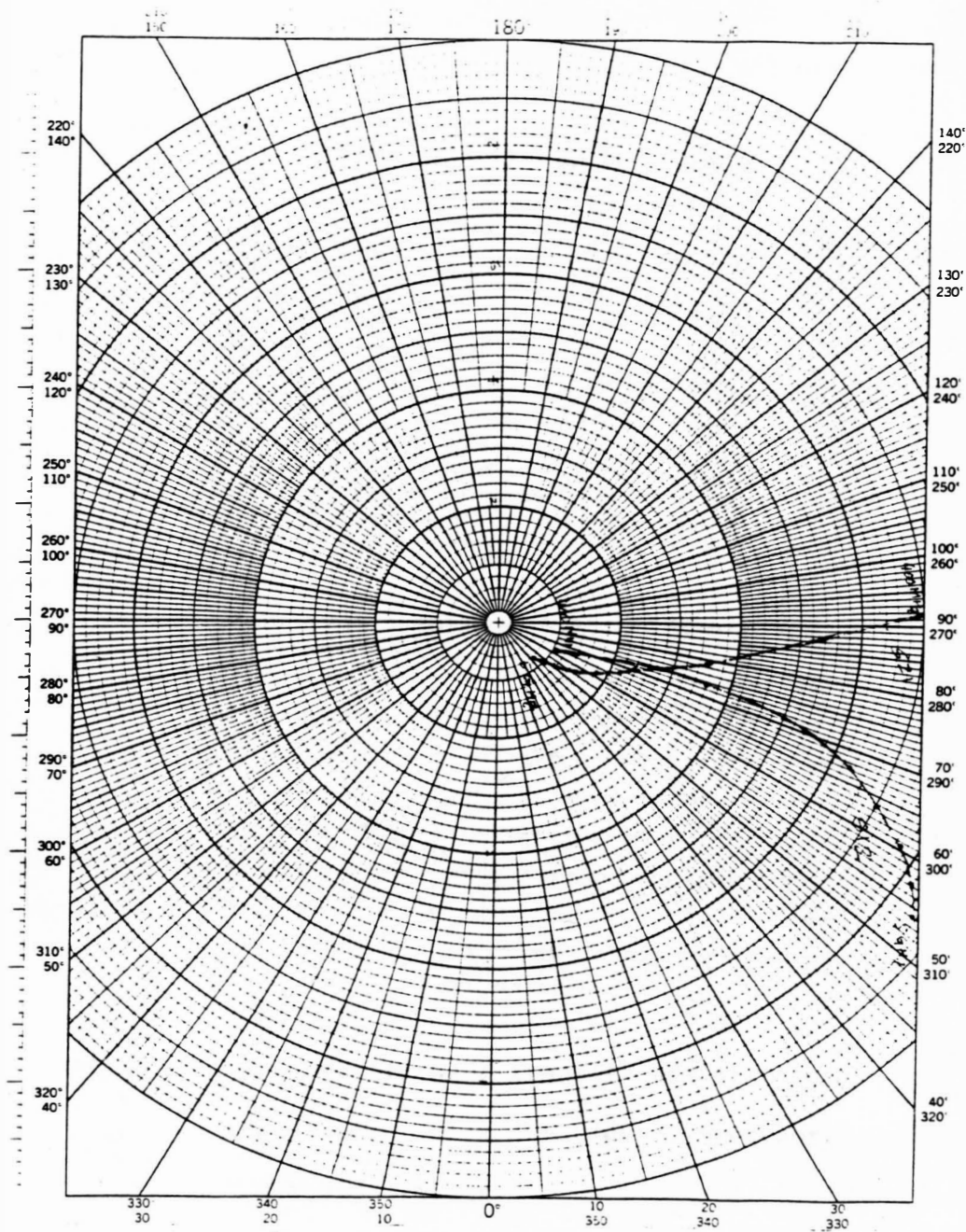


Fig.3.16 Forward and reverse transmission coefficients of transistor C as a function of frequency.  
( calculated values )

### $y$ PARAMETERS versus FREQUENCY

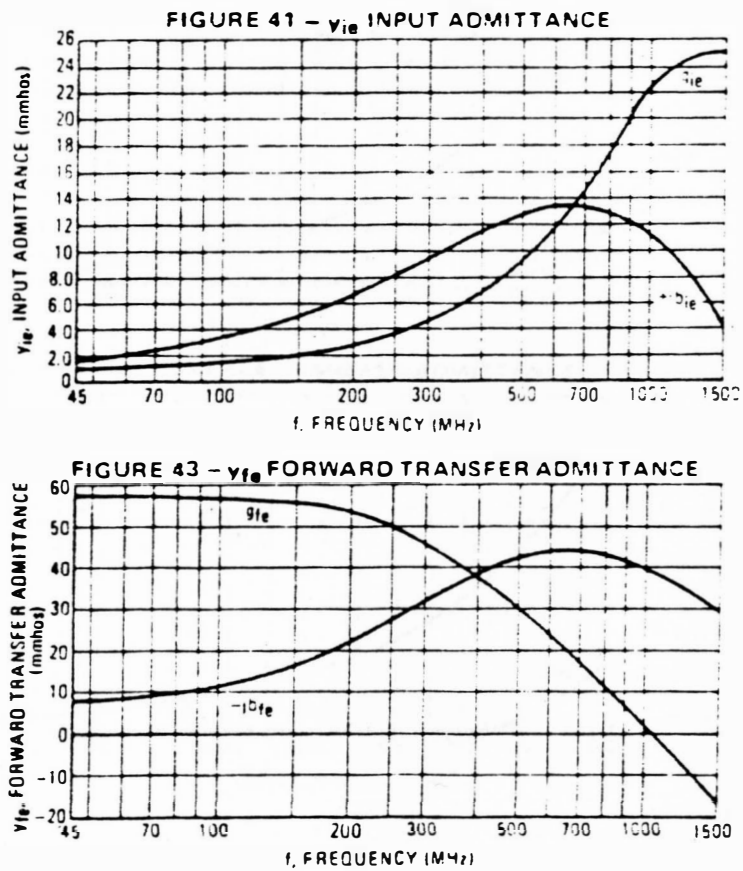


Fig.3.17 Input and forward transfer admittance of transistor D as a function of frequency. ( measured values )

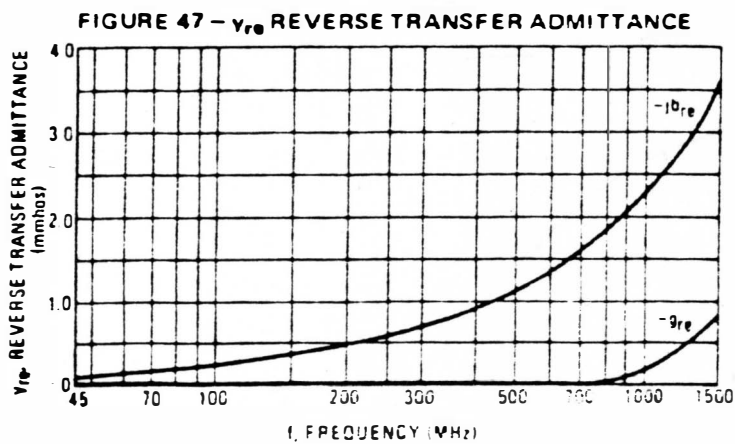
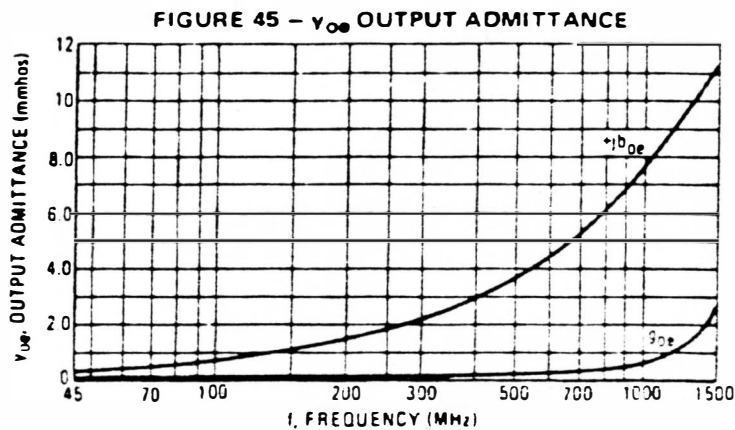


Fig.3.18 Output and reverse transfer admittance of transistor D as a function of frequency.  
( measured values )

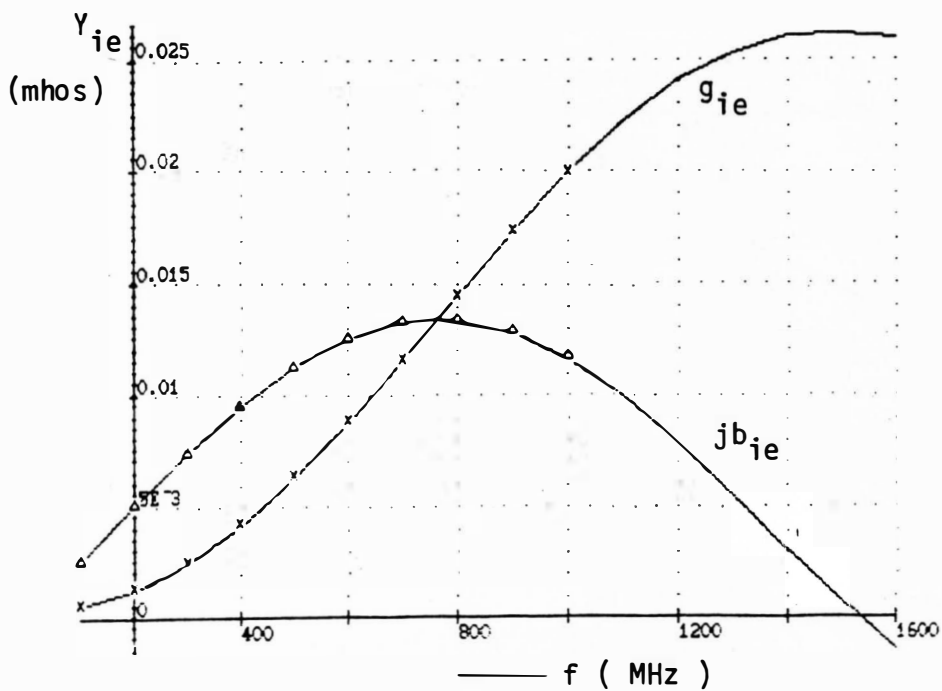
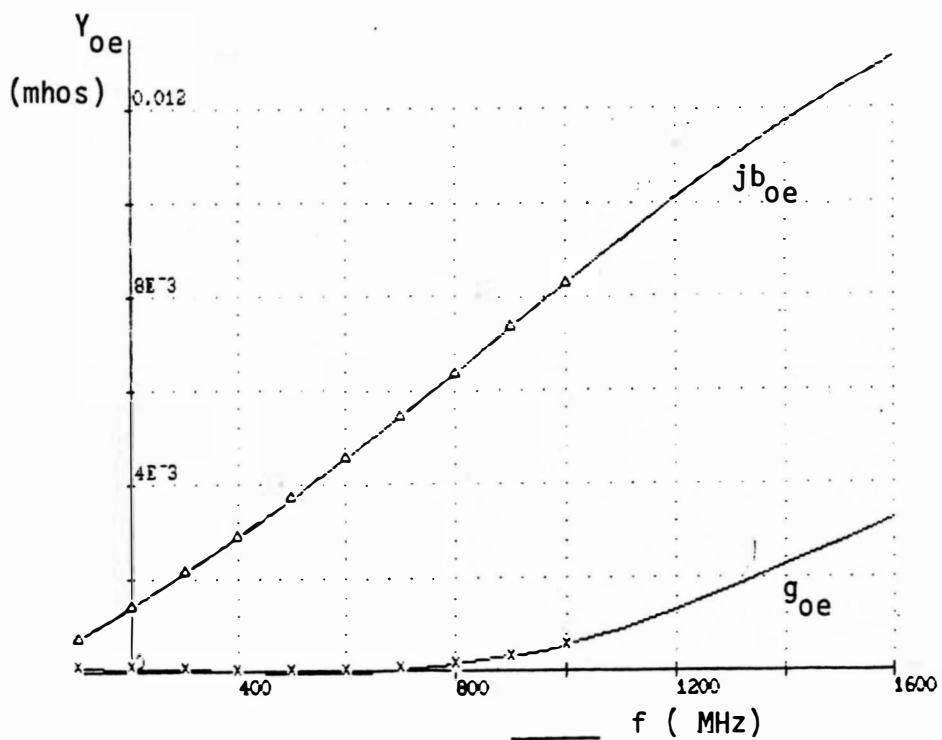


Fig.3.19 Output and input admittances of transistor D as a function of frequency. ( calculated values )



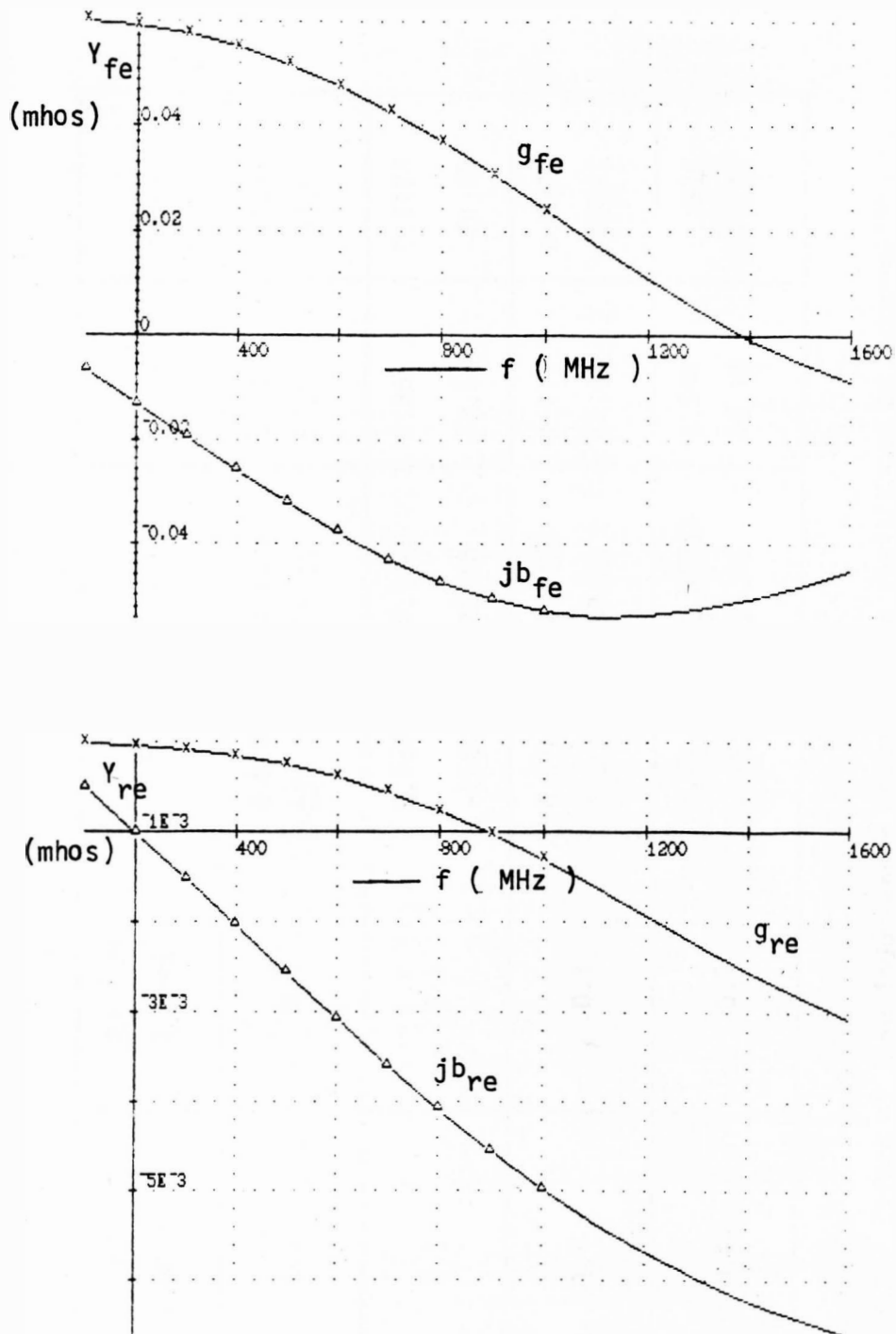


Fig.3.20 Forward and reverse transfer admittance of transistor D as a function of frequency. ( calculated values )

	Le = Lb ( nH )	Re = Rb ( ohms )	S11	S12	S21	S22
1	1.5	3.5	0.5264 159.7	0.1506 55.35	1.315 67.04	0.4037 -69.6
2	0.1	3.5	0.6663 -166.3	0.1225 22.18	1.36 78.54	0.4123 -74.84
3	1.5	0.1	0.6763 155	0.1328 67.85	1.423 65.43	0.3891 -68.91
4	0.1	0.1	0.85 -172.4	0.07877 16.46	1.504 78.14	0.3912 -76.52

TABLE 1 Effect of lead inductances and parasitic resistances on S parameters of transistor A, at 900 MHz.

parameter values for same frequencies are shown in figures 3.15 and 3.16, for comparison.

#### TRANSISTOR D [13]

Data sheet specifications for  $I_c = 2 \text{ mA}$ .

$C_c = 0.8 \text{ pF}$

$r_{bb}, C_c = 8 \text{ ps}$

$h_{fe} = 200$

$f_T = 2.5 \text{ GHz}$ .

$L_e = 1.5 \text{ nH}$

$L_b = 1.5 \text{ nH}$

$R_e = 3.5 \text{ ohms}$

$R_b = 3.5 \text{ ohms}$

Maximum power = 200 mW

Type - PNP, 2N4957 ( Motorola )

The plots of measured values of common emitter Y parameter values for frequencies from 100 MHz to 1000 MHz are shown in figures 3.17 and 3.18. The plots of numerically calculated Y parameter values for same frequencies are shown in figures 3.19 and 3.20 for comparison.

The importance of  $R_e$ ,  $R_b$ ,  $L_e$  and  $L_b$  can be seen from an example. For transistor A, at  $f = 900 \text{ MHz}$ , the S parameters were calculated numerically, for different values of  $R_e$ ,  $R_b$ ,  $L_e$  and  $L_b$ . The results are shown in table 1.

## CHAPTER 4

### OSCILLATOR CIRCUIT

In the previous chapter, the validity of the high frequency model of the transistor was verified. Now the model can be used in designing an oscillator circuit. The ac design is done using Y parameters.

The following steps are followed in the ac design of a VCO:

**STEP 1** - Finding out the net phase shift produced by the transistor itself. This is found by calculating the forward transmission coefficient S21 of the transistor at the desired frequency.

**STEP 2** - Estimating the electric length needed to obtain zero phase shift with feedback, since a circuit will oscillate when a feedback path is present providing at least unity loop gain with zero phase shift. This can be calculated from the phase angle of S21.

$$\text{That is } \text{Bd} = \text{S21} \text{ mod } 90^\circ$$

$$(2\pi/\lambda) d = \text{S21}$$

$$(2\pi/(f/c)) d = \text{S21}$$

$$\text{Therefore } d = (f/c) \times (1/2\pi) \times \text{S21}$$

where  $f$  is the desired frequency in MHz

$c$  is the speed of light in free space in Mm/s

$d$  is the length of the feedback transmission line in meters.

**STEP 3** - Calculating the Y parameters of the feedback transmission line. The transmission line can be considered as a two port network as shown in fig.4.1. In the figure,  $d$  is the length of the transmission line,  $Z_0$  is the characteristic impedance,  $Z_L$  is the load impedance in which the transmission line is terminated.

The input admittance of the transmission line can be given by [14]:

$$Y_{11} = 1/(j Z_0 \tan \beta d) = -j Y_0 \cot \beta d$$

and the forward transfer admittance is given by [14]:

$$Y_{21} = 1/(-j Z_0 \cot \beta d) = j Y_0 \tan \beta d.$$

By symmetry -

$$Y_{22} = Y_{11} = -j Y_0 \cot \beta d$$

$$Y_{12} = Y_{21} = j Y_0 \tan \beta d$$

**STEP 4** - Calculating the composite Y parameters of the transistor and feedback line. The block diagram in fig.4.2 shows the feedback connections.

Since the feedback transmission line is connected in parallel with the transistor, each composite Y parameter of the entire block is the sum of the respective Y parameters of the transmission line and transistor.

If  $Y_{11}$  is the input admittance of the transistor and  $Y_{11L}$  is the input admittance of the transmission line, then the input admittance of the entire block is  $Y_{11C}$  given by  $Y_{11C} = Y_{11} + Y_{11L}$ . (4.1)

In general, the above result can be written as -

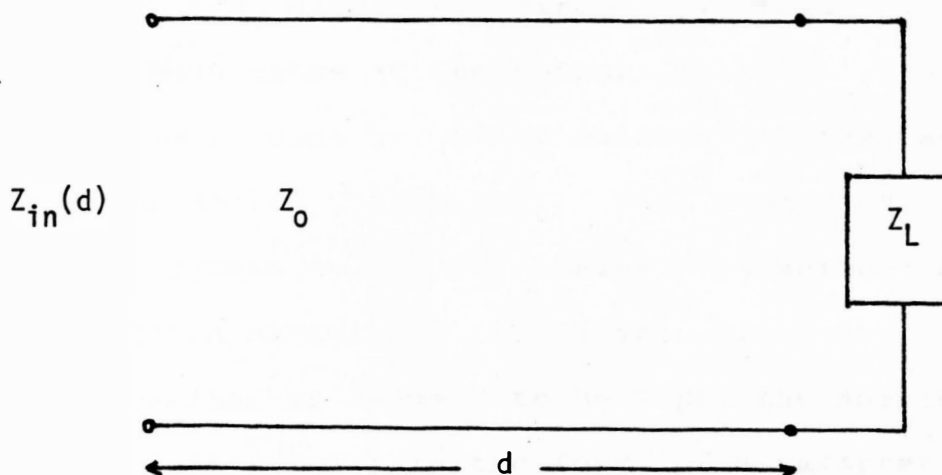


Fig.4.1 Transmission line as a two port network.

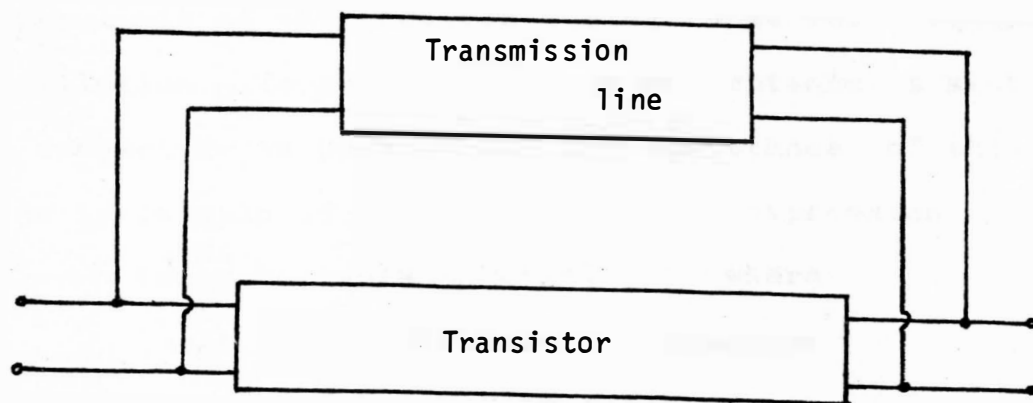


Fig.4.2 Block diagram showing feedback connections.

$$Y_{ijC} = Y_{ij} + Y_{ijL} \quad (4.2)$$

STEP 5 - Calculating the stub line length so as to cancel the susceptance due to output capacitance of the transistor and connected circuitry, thus giving net negative conductance at the output.

This is done by first calculating the admittance of the connected circuitry. The connected circuitry basically consists of a tuning capacitor and the transformation capacitor. To start with, the value of tuning capacitor is assumed to be 2 pF, the smallest value available commercially in the form of a varacter and the value of transformation capacitor is calculated to be 3.3 pF to match a 500 ohms load. The output susceptance is calculated due to these capacitors along with the output capacitance of the transistor, at 1000 MHz, which is the upper limit of the required tuning range for frequency of oscillation. To neutralize this susceptance, a stub line is connected in parallel. The admittance of this stub line is calculated from the following expression:

$$Y_{st} = Y_o / \tan(w \times D_{st} / c) \quad \text{where}$$

$Y_o$  is the characteristic impedance

$w$  is the frequency in radians/s

$D_{st}$  is the stub line length.

The block diagram incorporating the stub line and impedance transformation network is shown in fig.4.3. The equivalent circuit results in the stub being connected in

parallel to the output elements. Hence the resultant output admittance will be the sum of  $Y_{22C}$ ,  $Y_{inN}$  and  $Y_{st}$ . That is  $Y_{22CM} = Y_{22C} + Y_{inN} + Y_{st}$ .

STEP 6 - So far in the above steps, ac design is discussed for a single frequency of oscillation. To make the circuit oscillate at different frequencies, a varacter diode is added at the base or collector port. The basic expression which gives the capacitance of the diode is:

$$C_T = \epsilon A/W \quad \text{where}$$

$\epsilon$  = Permittivity of the material at the junction.

A = area of the junction.

W = Width of the space charge.

The width of the space charge is a function of the reverse voltage. As reverse voltage is increased, the width increases resulting in smaller capacitance. Hence the oscillator can be tuned to the desired frequency by varying the bias voltage which results in variation in capacitor value.

Initially the tuning capacitor was added at the base, but did not prove very useful as the tuning range obtained was very small. Then the tuning capacitor was added at the collector. This proved very useful since a larger tuning range was possible. If this capacitor has capacitance  $C_T$  then the admittance of the capacitor is given by:

$$Y_{C_T} = j\omega C_T \quad (4.3)$$



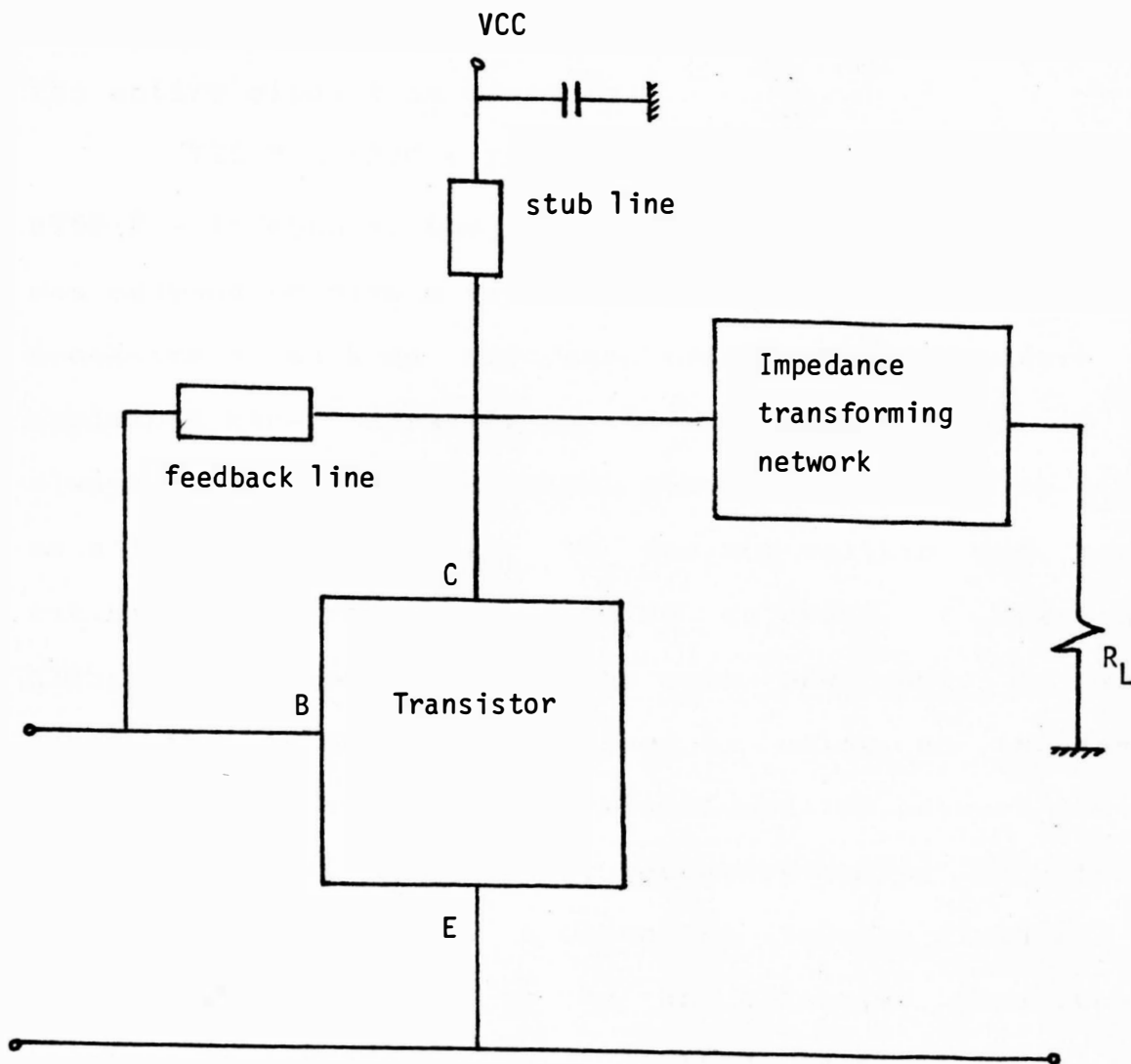


Fig.4.3 Block diagram of an oscillator.

Output admittance  $Y_{22CM} = Y_{22C} + Y_{st} + Y_{CT} + Y_{CTR}$

It was discussed earlier that for oscillation to take place, the output admittance should have negative real part which is the conductance and a zero imaginary part which is the susceptance. The expression to calculate output admittance is:

$$Y_{out} = Y_{22CM} - (Y_{21CM} \times Y_{12CM})/Y_{11CM}.$$

AC design steps are explained in detail with examples in the following chapter.

## CHAPTER 5

### NUMERICAL RESULTS

The ac design was carried out for four different transistors. Emphasis was given to  $f_T$  and  $C_c$  in selecting the transistors. The purpose of repeating the design procedure for different transistors was to find the effects of  $f_T$  and  $C_c$  on feedback line length, stub line length, tuning capacitor range and the external  $Q$  of the circuit. The transistors that were used in this study were the same discussed earlier.

#### TRANSISTOR A

**STEP 1 - Finding the best feedback line length for the tuning range of 800 MHz to 1000 MHz.**

First, an approximate feedback line length was calculated for 800 MHz and 1000 MHz, the two extreme frequencies, from the  $S$  parameters of the transistor at these frequencies.

It was explained earlier that the electrical length that was necessary to get zero phase shift can be obtained from the phase angle of  $S_{21}$ . For transistor A the phase angle of  $S_{21}$  at 800 MHz was  $61.63^\circ$  which was the electrical length necessary. That is -

$$B d = 61.63^\circ = 0.3424 \text{ rad.}$$

Therefore  $d = 0.0642$  meters.

With this approximate length, a length response was done for frequency of 800 MHz. The function used is called FBLENGTH, which is given in the appendix. This function calculates the real part of  $Y_{out}$  as a function of feedback line length. To run the function, tuning capacitor, transformation capacitor and stub admittance were set to zero. The optimum length was the one that gave maximum real part. The best length obtained was 0.068 meters, for which the real part was -23.1 mmhos.

The same procedure was repeated for  $f = 1000$  MHz. The phase angle of  $S_{21}$  for  $f = 1000$  MHz was  $56.84^\circ$ . Hence the feedback line length calculated was 0.0474 meters. After running the function FBLENGTH the optimum length obtained was 0.053 meters, for which the real part was -14.4 mmhos.

Two important points were observed from the above. First, as the frequency increased, the length of the feedback line decreased. Second, with the increase in frequency, the absolute magnitude of the real part decreased. With the shorter feedback line length the following advantages can be realized:

- 1) Reduced board space.
- 2) Easier fabrication of feedback line resulting in lower cost.

However it was necessary to have a negative real part for this feedback line length at 800 MHz. The function

FBLENGTH was run again to find out the value of real part at 800 MHz. It was found to be -12 mmhos, which was sufficient to meet the requirements.

STEP 2 - Finding the best stub line length to neutralize all the capacitance in the circuit to result in zero susceptance and negative conductance. The best value of the stub line length was calculated by setting the feedback line length to 0.053 meters, the transformation capacitor to 3.3 pF, the tuning capacitor to 2 pF and the frequency to 1000 MHz. The tuning capacitor value was chosen to be 2 pF because this is the smallest value that is available from a suitable varactor for tuning purposes. With these values set, a function called STLENGTH was run. This function calculated the output admittance  $Y_{out}$  as a function of stub line length. The optimum value of the stub line length was the one for which the imaginary part of  $Y_{out}$  is zero. The optimum value was found to be 0.0207 meters.

STEP 3 - Verifying the tuning range. The oscillator circuit was designed to oscillate from 800 MHz to 1000 MHz depending on the tuning capacitor value. This was verified by running the function called COVER. This function calculated the value of  $Y_{out}$  as a function of frequency for a fixed value of feedback line length, stub line length, transformation capacitor and tuning

capacitor. The oscillation frequency was the frequency at which the imaginary part of  $Y_{out}$  went to zero.

This function COVER was run for different values of tuning capacitor and the following frequencies of oscillation were found:

For  $C_T = 10$  pF, the frequency of oscillation was found to be 800 MHz and the real part of  $Y_{out}$  was found to be -12 mmhos.

For  $C_T = 5.75$  pF,  $f_o = 900$  MHz and  $Y_{out}(\text{real}) = -15.34$  mmhos.

For  $C_T = 2$  pF,  $f_o = 1000$  MHz and  $Y_{out}(\text{real}) = -14.2$  mmhos.

The output admittance  $Y_{out}$  as a function of frequency for different values of tuning capacitor was plotted as shown in figures 5.1 through 5.4. From these plots both oscillation frequency and the external  $Q$  of the circuit were calculated. As explained earlier, the oscillation frequency was the frequency at which the imaginary part of  $Y_{out}$  went to zero, and the value of external  $Q$  was calculated in the following way:

From the plot, the change in the imaginary part of  $Y_{out}$  was found corresponding to change in the frequency. Hence the external  $Q$  was calculated from the following expression -

$$Q_{ext.} = f_o \times (dB_{out}/df \times RL) \quad \text{where}$$

$f_o$  is the frequency of oscillation

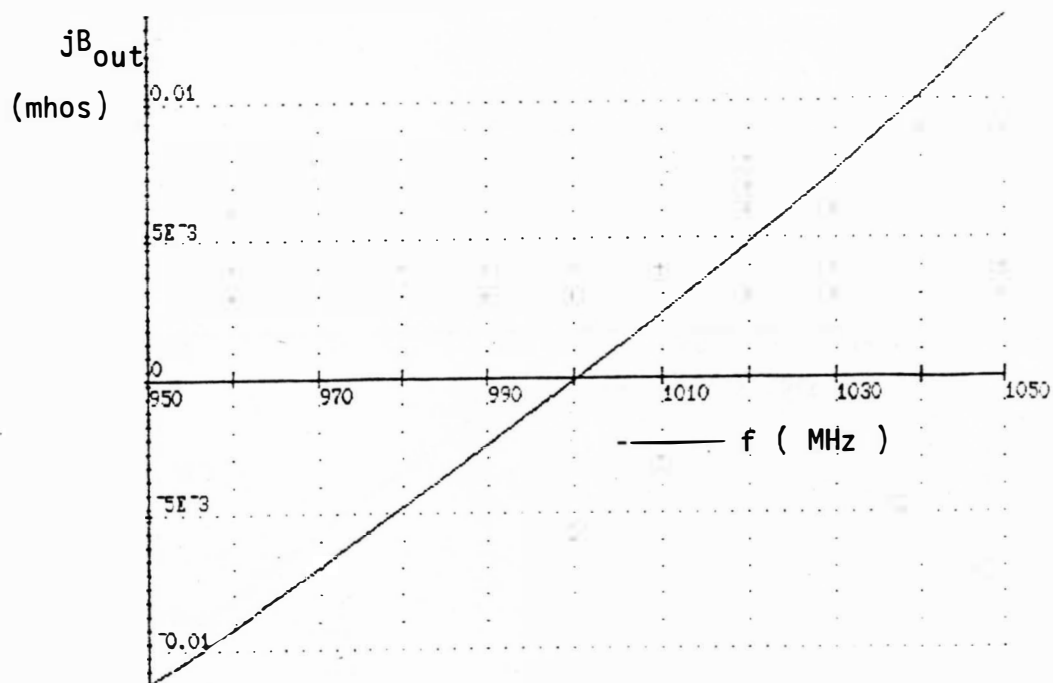


Fig.5.1 Output susceptance as a function of frequency for tuning capacitor of 2 pF, for transistor A.

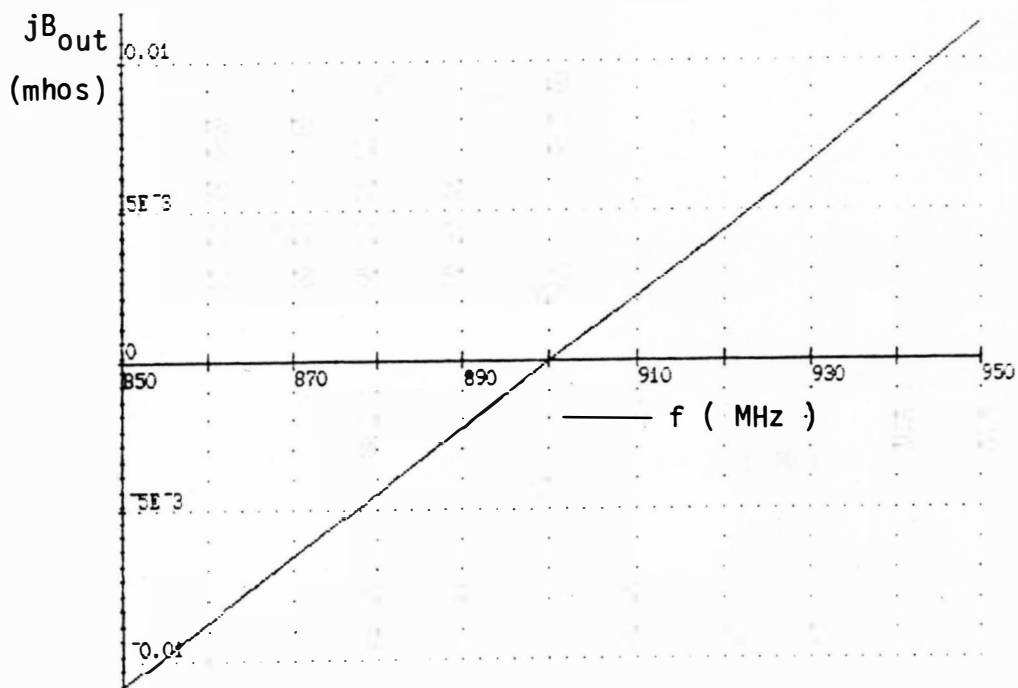


Fig.5.2 Output susceptance as a function of frequency for tuning capacitor of 5.75 pF, for transistor A.



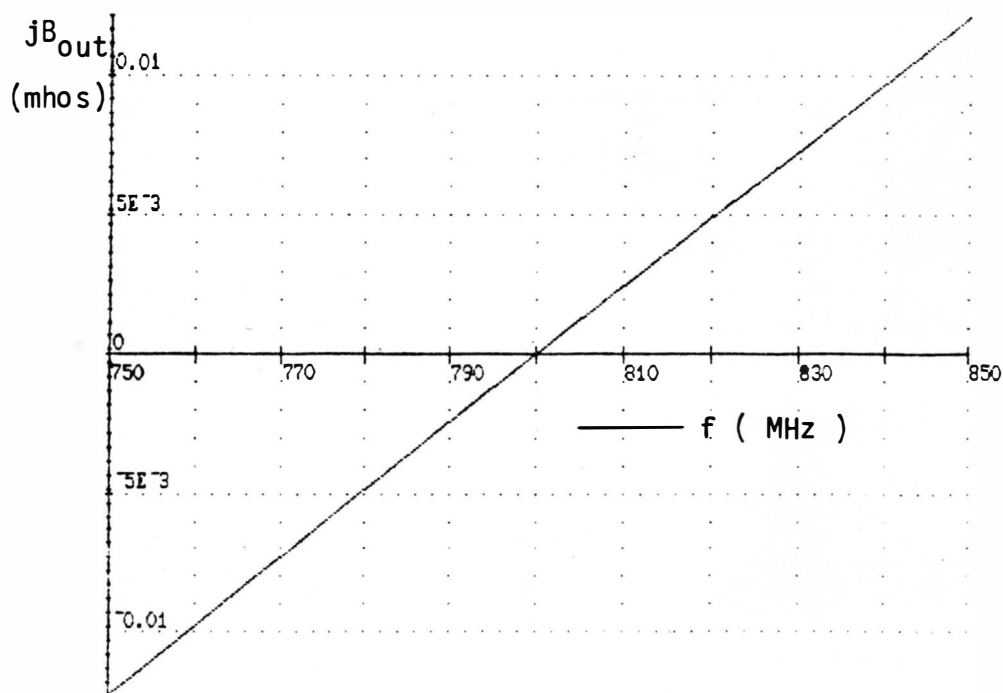


Fig.5.3 Output susceptance as a function of frequency for tuning capacitor of 10.05 pF, for transistor A.

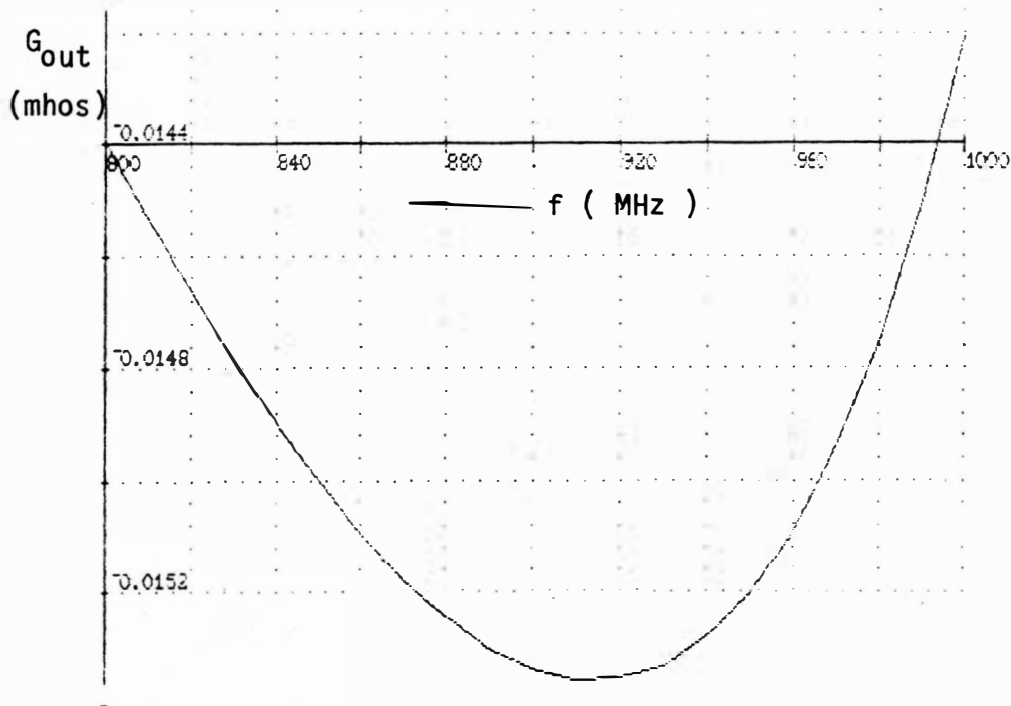


Fig.5.4 Output conductance as a function of frequency for  $D_L$  of 0.053 mts, for transistor A.

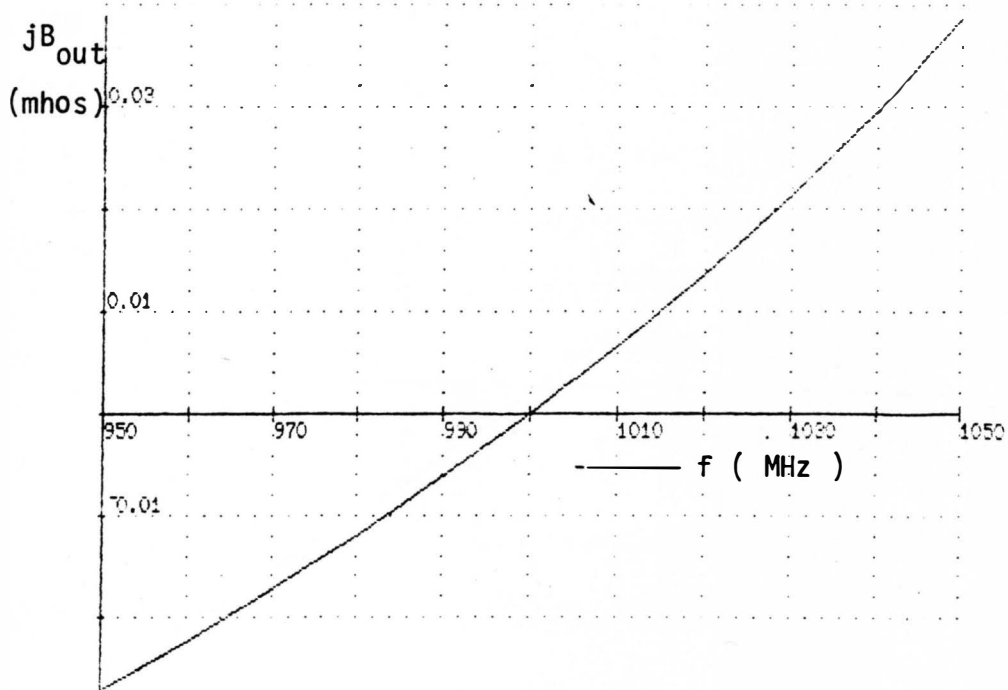


Fig.5.5 Output susceptance as a function of frequency for tuning capacitor of 2 pF, for transistor B.

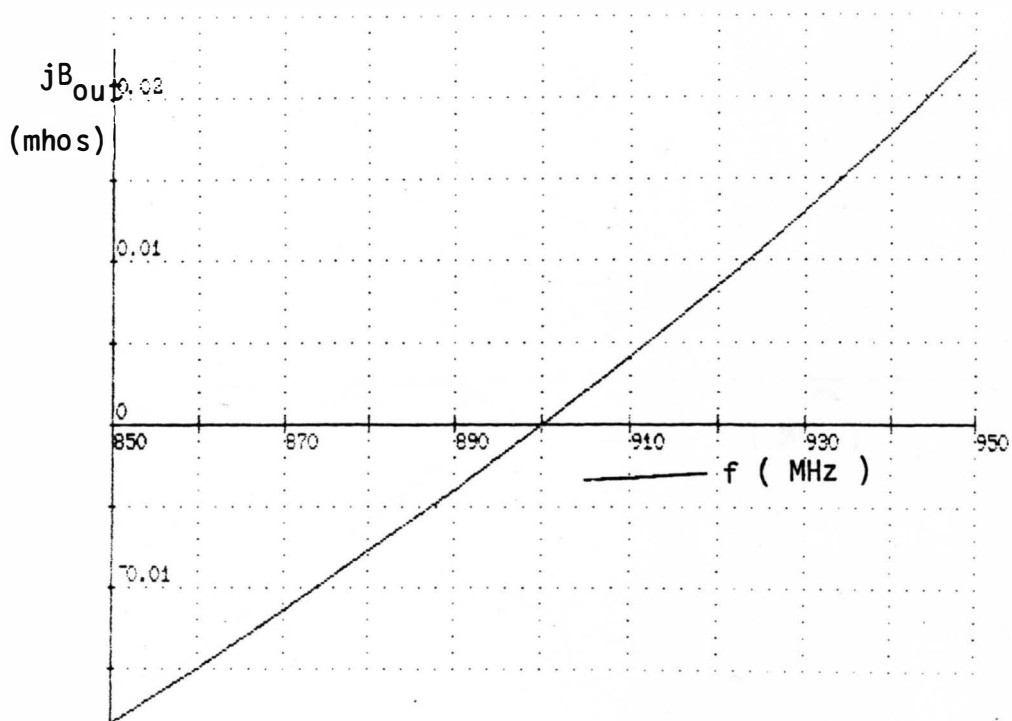


Fig.5.6 Output susceptance as a function of frequency for tuning capacitor of 10.35 pF, for transistor B.

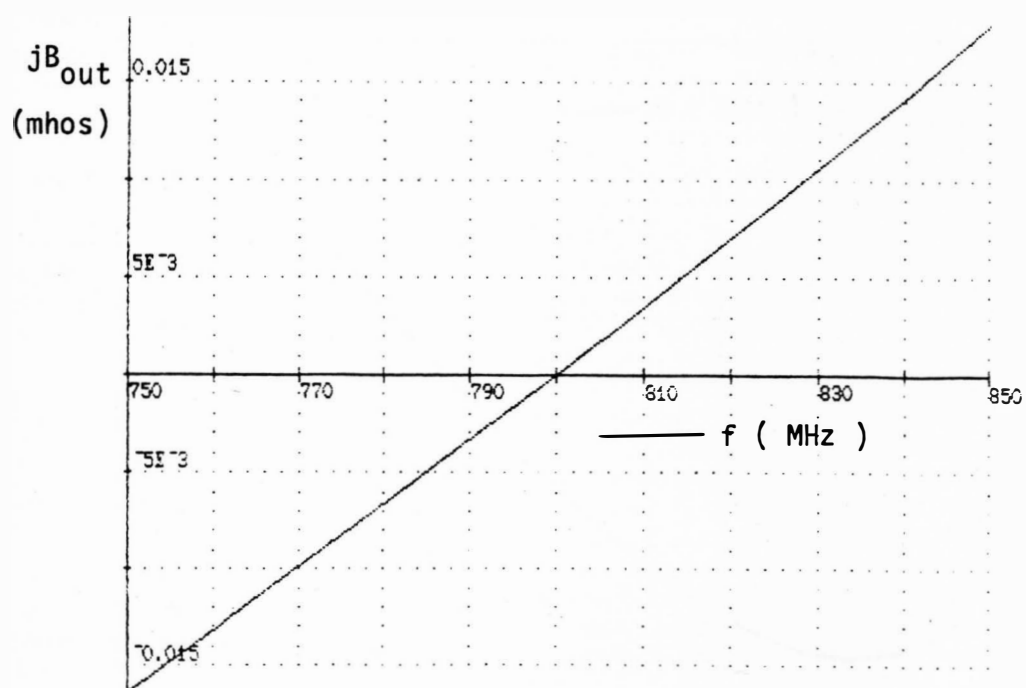


Fig.5.7 Output susceptance as a function of frequency for tuning capacitor of 17.15 pF, for transistor B.

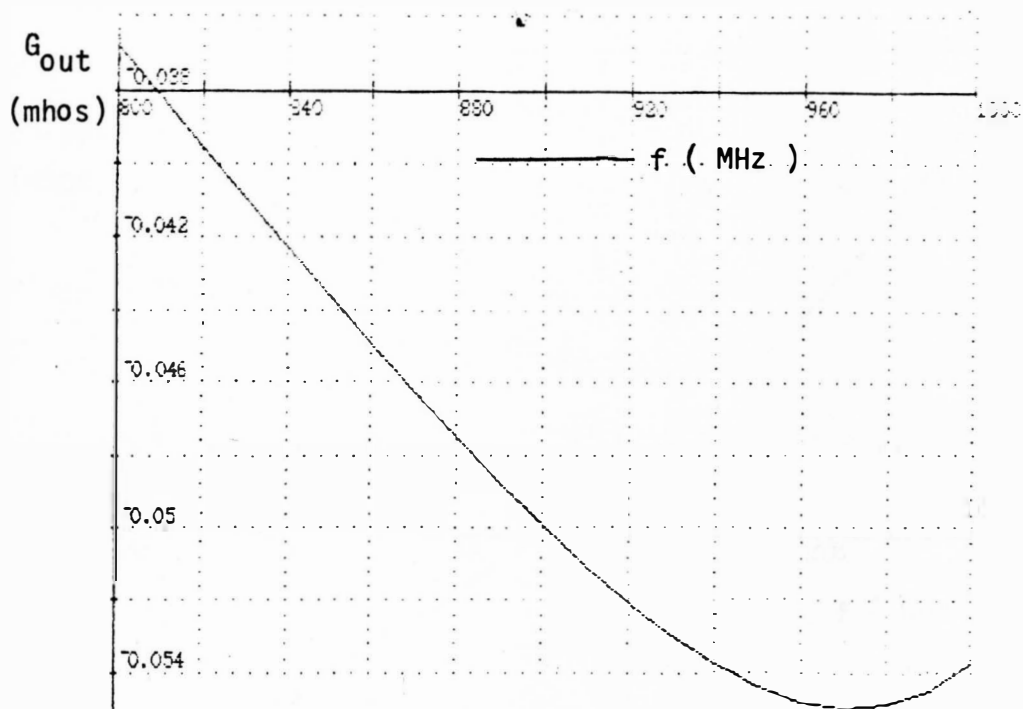


Fig.5.8 Output conductance as a function of frequency for  $D_L$  of 0.0558 mts. for transistor B.

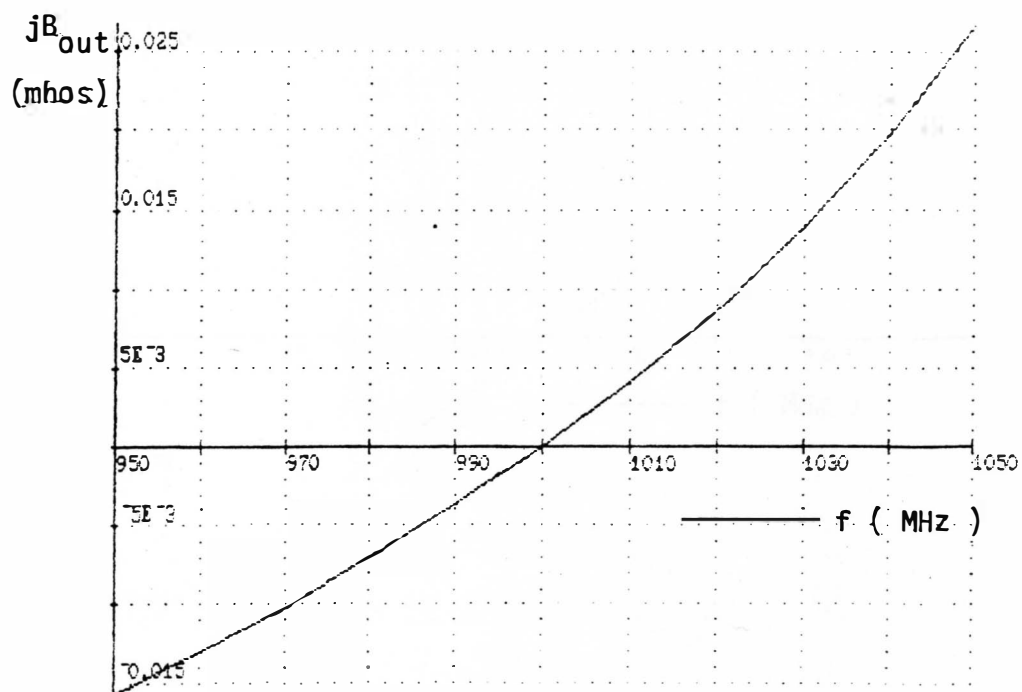


Fig.5.9 Output susceptance as a function of frequency for tuning capacitor of 2 pF, for transistor C.

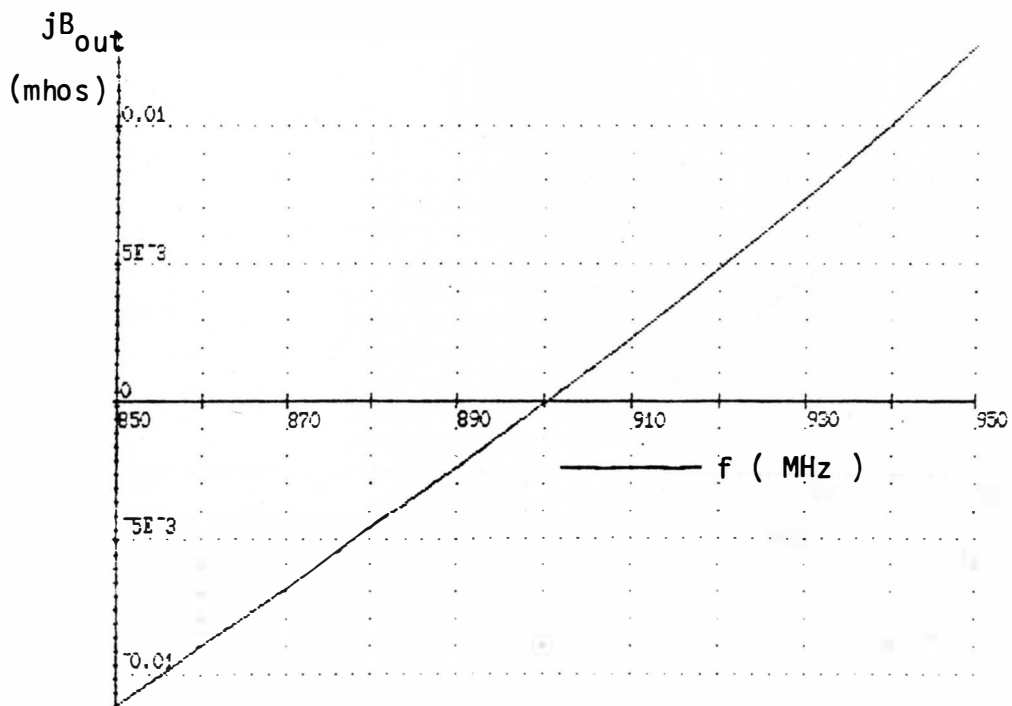


Fig.5.10 Output susceptance as a function of frequency for tuning capacitor of 6.8 pF, for transistor C.



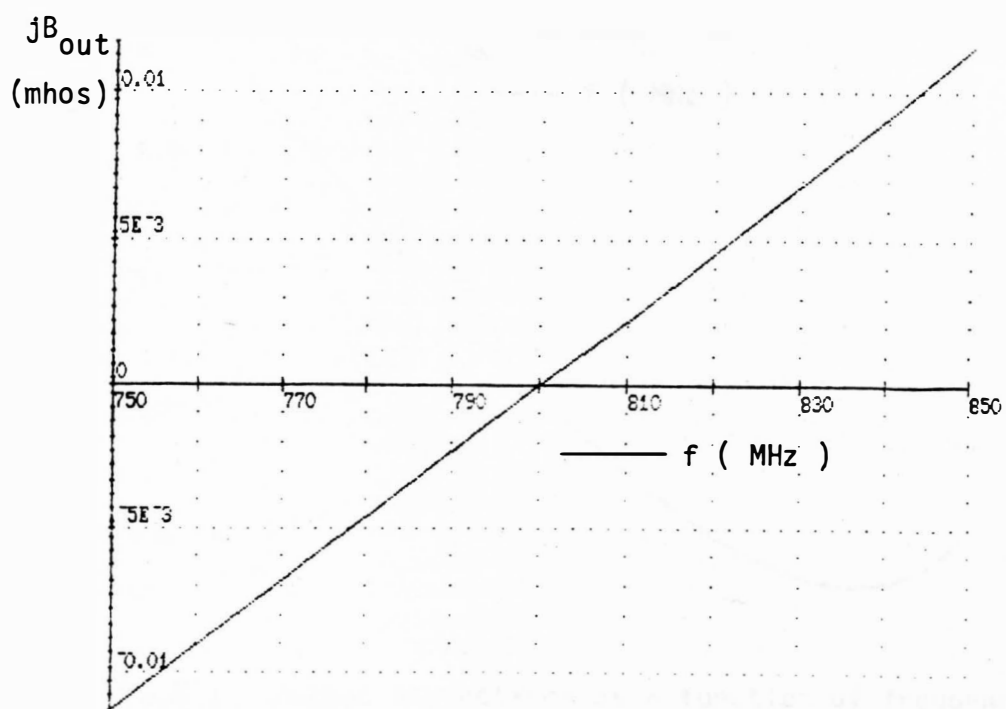


Fig.5.11 Output susceptance as a function of frequency for tuning capacitor of 11.05 pF, for transistor C.

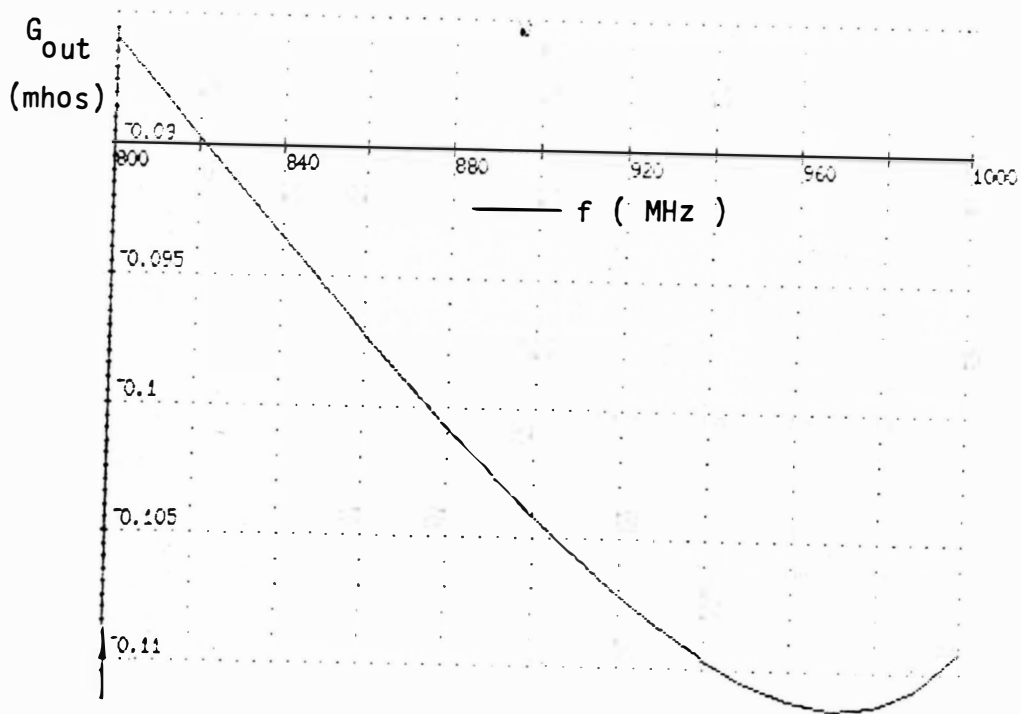


Fig.5.12 Output conductance as a function of frequency for  $D_L$  of 0.06175mts., for transistor C.

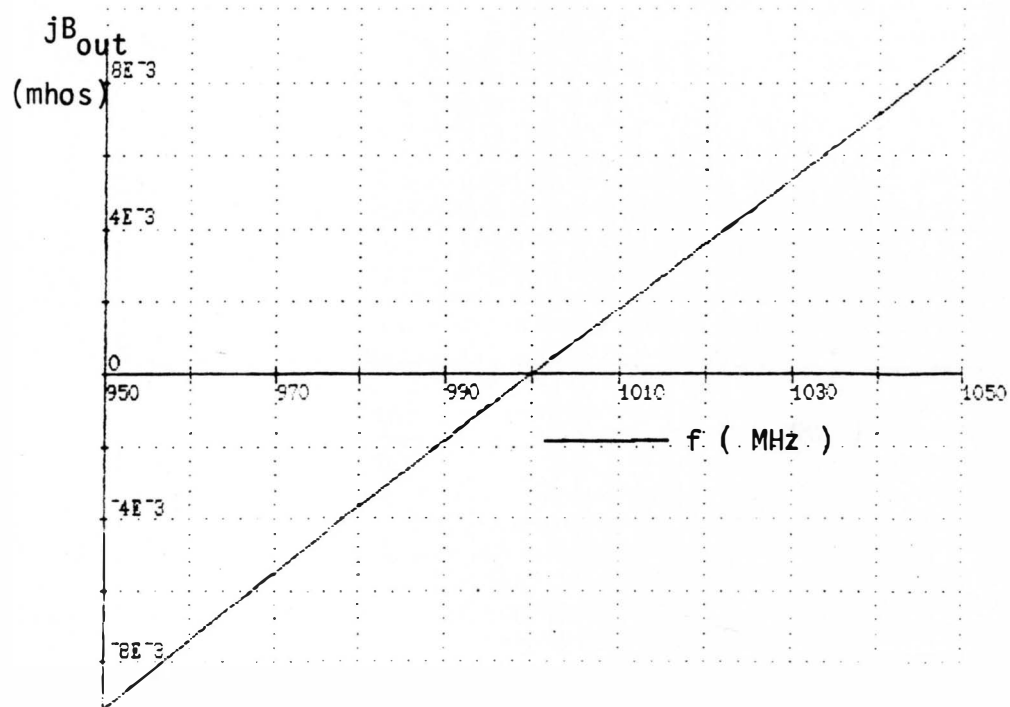


Fig.5.13 Output susceptance as a function of frequency for tuning capacitor of 2 pF, for transistor D.

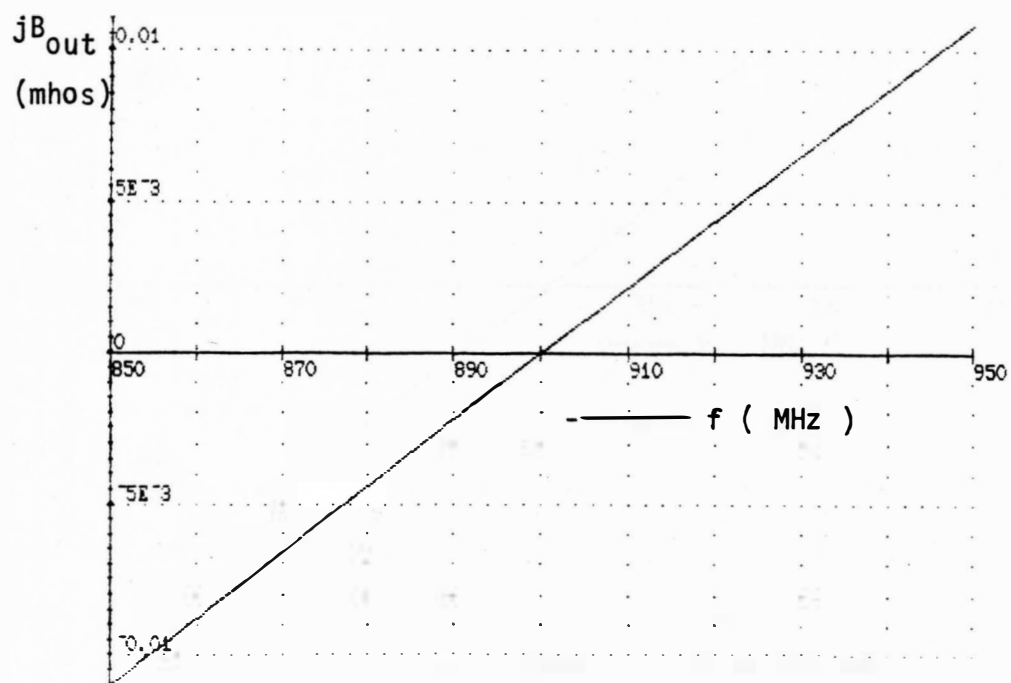


Fig.5.14 Output susceptance as a function of frequency for tuning capacitor of 5.4 pF, for transistor D.

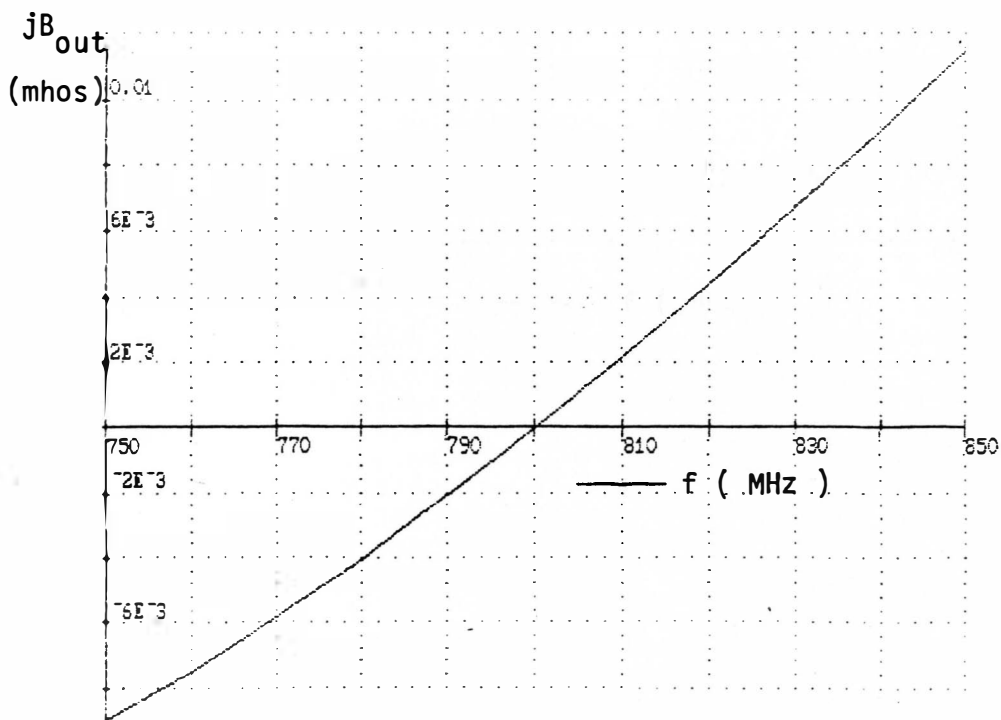


Fig.5.15 Output susceptance as a function of frequency for tuning capacitor of 9.575 pF, for transistor D.

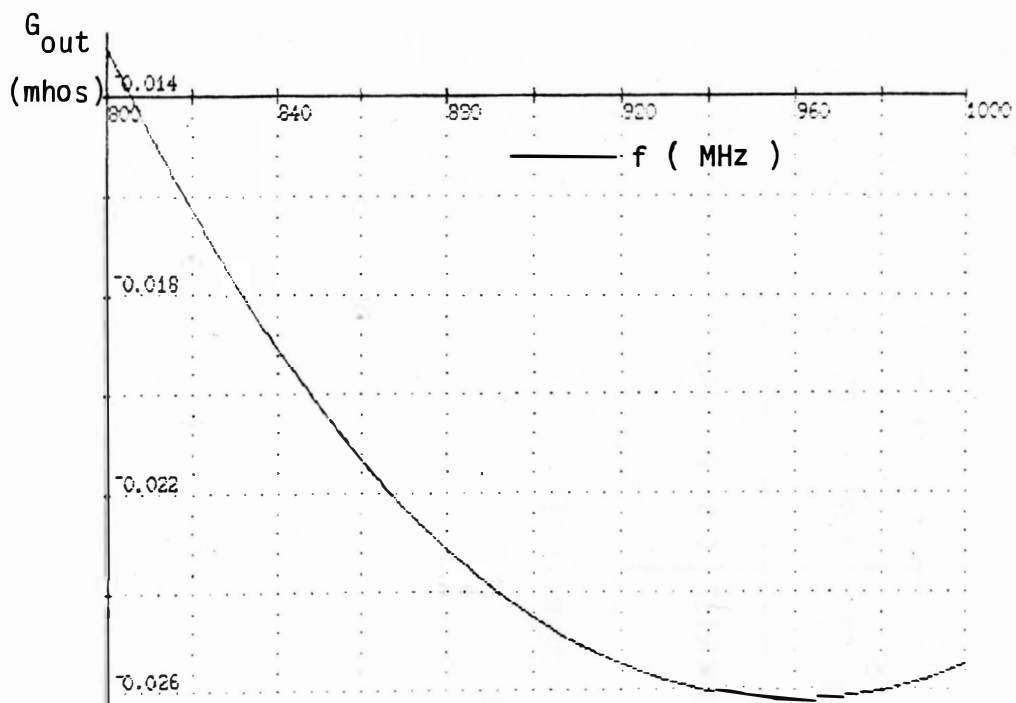


Fig.5.16 Output conductance as a function of frequency for  $D_L$  of 0.05 mts., for transistor D.

	$C_c$ (pF)	$f_T$ (GHz)	$D_L$ (mts)	$D_{STB}$ (mts)	$Q_{max}$	$Q_{min}$	$G_{o,max}$ (mmhos)	$G_{o,min}$ (mmhos)	$C_T$ max (pF)	$C_T$ min (pF)
TRAN. A	1.7	1.3	0.053	0.0207	120	96.75	-15.34	-14.2	10	2
TRAN. B	1.7	4.4	0.0558	0.01526	312.5	135.17	-53.8	-36.8	17.1	2
TRAN. C	1.0	4.0	0.06175	0.0725	187.5	94	-109	-85.9	11.05	2
TRAN. D	0.8	2.5	0.05	0.0898	92.353	80.1	-25.45	-13.0	9.55	2
TRAN. E	2.0	4.0	0.063	0.0324	225	100	-117	-82.9	12.25	2

TABLE 2. Summary of numerical results. In the table,  $C_T$  min is the value chosen for designing an oscillator.

$R_L$  is the transformed load resistance which is 500 ohms

$dB_{out}/df$  is the slope of  $B_{out}$  at  $f_o$

For transistor A, the external Q values for different oscillation frequencies were:

For  $f_o = 800$  MHz,  $Q_{ext}$  was 105.13.

For  $f_o = 900$  MHz,  $Q_{ext}$  was 96.75.

For  $f_o = 1000$  MHz,  $Q_{ext}$  was 120.

Similar steps were carried out for the other three transistors. The results obtained for these transistors are tabulated in table 2.

In the table, the last row gives the results of a hypothetical transistor, designated as Transistor E. This transistor has the same parameter values as Transistor C except for the output capacitance  $C_c$ . The value of  $C_c$  was increased from 1 pF to 2 pF. The design and evaluation steps were carried out for the other four transistors. It can be seen from the table that the value of  $C_c$  does have an effect on feedback line length, stub line length, tuning capacitor value and the external Q of the circuit. A statistical analysis was done to find out the significance of this effect, as explained in the following chapter.



## CHAPTER 6

### ANALYSIS AND CONCLUSION

The work reported was carried out in two stages. In the first stage, a high frequency model of a transistor was developed and the validity of the model was tested using S and Y parameters for four different transistors. The validity was proved as explained in Chapter Three.

In the second stage, the high frequency model of the transistor was used in the ac design of the VCO and the numerical results were obtained as tabulated in the previous chapter.

It can be seen from the table 2 that it is very difficult to arrive at a conclusion regarding the dependencies of feedback line length, stub line length, tuning capacitor value and the external Q value of the circuit on  $f_T$  and Cc. Hence, a statistical analysis was carried out to determine the dependencies of the above mentioned values on  $f_T$  and Cc. A multiple regression analysis was done for each of the values separately, considering each value as a dependent variable and  $f_T$  and Cc as independent variables. The results of the analysis are as follows. In the results DL and Dstb are in meters, Cc is in pico farads and  $f_T$  is in GHz.

1) Dependency of feedback line length on  $f_T$  and Cc :

Initially a linear model was used to determine the dependency. The model used was given as follows -

$$DL = A + B Cc + C f_T \quad \text{where}$$

A is a constant and B and C are the coefficients of Cc and f<sub>T</sub>. The results obtained for this model are shown in figure 6.1. from which the linear model can be written as

$$DL = 0.04745461 + 0.000057782 Cc + 0.00227270 f_T.$$

In the figure, there are two test values which show the validity of the linear model. They are the MODEL F value and R-SQUARE value. The MODEL F value gives the overall utility of the model and the R-SQUARE value gives the sample multiple coefficient of determination. From the figure it can be seen that the value of F obtained was 0.36 at the observed significance level of 0.7644. This means the dependency of feedback line length on f<sub>T</sub> and Cc can be concluded with only about 23.56 percent. In other words it can be said that the feedback line length does not depend linearly on f<sub>T</sub> and Cc. Later many more models were tried to find the dependency of feedback line length on f<sub>T</sub> and Cc. The best model that showed the dependency was a quadratic model in Cc, whose results are shown in figure 6.2. The model is given as follows -

$$DL = A_1 + B_1 Cc + C_1 Cc * Cc$$

In terms of the coefficient values from figure 6.2 the model can be written as follows-

$$DL = -0.05855556 + 0.19725 Cc - 0.07694444 Cc^2$$

SAS 1  
 11 20 FRIDAY, JULY 24, 1987  
 GENERAL LINEAR MODELS PROCEDURE

DEPENDENT VARIABLE DL

SOURCE	DF	SUM OF SQUARES	MEAN SQUARE	
MODEL	2	0.00003123	0.00001561	
ERROR	1	0.00004390	0.00004390	
CORRECTED TOTAL	3	0.00007513		

MODEL F = 0.36 PR F = 0.7644

R-SQUARE	C.V.	ROOT MSE	DL MEAN	
0.415692	12.0163	0.00662550	0.05513750	

SOURCE	DF	TYPE I SS	F VALUE	PR	F
CC	1	0.00000000	0.00	0.9994	
FT	1	0.00003123	0.71	0.5539	

SOURCE	DF	TYPE III SS	F VALUE	PR	F
CC	1	0.00000022	0.00	0.9551	
FT	1	0.00003123	0.71	0.5539	

PARAMETER	ESTIMATE	T FOR HO PARAMETER=0	PR	T	STD ERROR OF ESTIMATE
INTERCEPT	0.04745461	3.30	0.1872		0.01437221
CC	0.00057782	0.07	0.9551		0.00818490
FT	0.00227270	0.84	0.5539		0.00269449

Fig.6.1 SAS printout for dependency of feedback line length on  $f_T$  and  $C_c$  linearly.

SAS 1  
 11 22 FRIDAY, JULY 24, 1987  
 GENERAL LINEAR MODELS PROCEDURE

DEPENDENT VARIABLE DL

SOURCE	DF	SUM OF SQUARES	MEAN SQUARE
MODEL	2	0.00007121	0.00003560
ERROR	1	0.00000392	0.00000392
CORRECTED TOTAL	3	0.00007513	

MODEL F = 9.08 PR F = 0.2284

R-SQUARE	C.V.	ROOT MSE	DL MEAN
0.947822	3.5908	0.00197990	0.05513750

SOURCE	DF	TYPE I SS	F VALUE	PR	F
CC	1	0.00000000	0.00	0.9980	
CC*CC	1	0.00007121	18.17	0.1467	

SOURCE	DF	TYPE III SS	F VALUE	PR	F
CC	1	0.00007100	18.11	0.1469	
CC*CC	1	0.00007121	18.17	0.1467	

PARAMETER	ESTIMATE	T FOR H0 PARAMETER=0	PR	T	STD ERROR OF ESTIMATE
INTERCEPT	-0.05855556	-2.18	0.2740		0.02688373
CC	0.19725000	4.26	0.1469		0.04634652
CC*CC	-0.07694444	-4.26	0.1467		0.01805342

Fig.6.2 SAS printout for dependency of feedback line length on Cc quadratically.

From the figure it can be seen that the value of  $F$  increased greatly by using the quadratic model. Therefore it can be concluded that the feedback line length seems to depend on the square of the output capacitance  $C_c$ .

## 2) Dependence of stubline length on $f_T$ and $C_c$ :

From figure 6.3 it can be seen that the stub line length strongly depends on both  $f_T$  and  $C_c$  linearly. It can also be concluded that the increase in  $f_T$  and  $C_c$  values does decrease the stub line length. The model used is given as follows -

$$D_{stb} = A_2 + B_2 C_c + C_2 f_T$$

and from the results the linear model for  $D_{stb}$  can be written as follows-

$$D_{stb} = 0.15788836 - 0.07959994 C_c - 0.00158801 f_T$$

## 3) Dependence of tuning capacitor value on $f_T$ and $C_c$ :

From figure 6.4 it can be concluded with only about 70 percent confidence level that the tuning capacitor value does depend on  $f_T$  and  $C_c$ . The model used is given as follows -

$$C_T = A_3 + B_3 C_c + C_3 f_T$$

From the results obtained, the linear model for  $C_T$  can be written as follows-

$$C_T = 0.04883285 + 4.6581326 C_c + 1.90835735 f_T$$

## 4) Dependence of external $Q$ value on $f_T$ and $C_c$ :

From figure 6.5 it can be concluded that there is a strong linear dependence of external  $Q$  value on  $f_T$  and

```

SAS
11 22 FRIDAY, JULY 24, 1987
GENERAL LINEAR MODELS PROCEDURE
DEPENDENT VARIABLE DST
SOURCE          DF          SUM OF SQUARES          MEAN SQUARE
MODEL          2          0.00415424          0.00207712
ERROR          1          0.00000065          0.00000065
CORRECTED TOTAL          3          0.00415489

MODEL F =          3188.58          PR F = 0.0125

R-SQUARE          C.V.          ROOT MSE          DST MEAN
0.999843          1.6284          0.00080711          0.04956500

SOURCE          DF          TYPE I SS          F VALUE          PR          F
CC          1          0.00413899          6353.75          0.0080
FT          1          0.00001525          23.41          0.1298

SOURCE          DF          TYPE III SS          F VALUE          PR          F
CC          1          0.00415179          6373.40          0.0080
FT          1          0.00001525          23.41          0.1298

PARAMETER          ESTIMATE          T FOR HO          PR          T          STD ERROR OF
PARAMETER=0          ESTIMATE
INTERCEPT          0.15788836          90.18          0.0071          0.00175080
CC          -0.07959594          -79.83          0.0080          0.00099707
FT          -0.00158801          -4.84          0.1298          0.00032824

```

Fig.6.3 SAS printout for dependency of stubline length on  $f_T$  and  $C_c$  linearly.

```

SAS
11 22 FRIDAY, JULY 24, 1987
GENERAL LINEAR MODELS PROCEDURE

DEPENDENT VARIABLE C

SOURCE          DF          SUM OF SQUARES          MEAN SQUARE
MODEL          2          28670.89633339          14335.44816669
ERROR          1          208.98112336           208.98112336
CORRECTED TOTAL 3          28879.87745675

MODEL F =          68.60          PR F = 0.0851

R-SQUARE          C.V.          ROOT MSE          Q MEAN
C.992764          8.1174          14.45617942          178.08825000

SOURCE          DF          TYPE I SS          F VALUE          PR          F
CC          1          7546.39227614          36.11          0.1050
FT          1          21124.50405725          101.08          0.0631

SOURCE          DF          TYPE III SS          F VALUE          PR          F
CC          1          9777.52873499          46.79          0.0924
FT          1          21124.50405725          101.08          0.0631

PARAMETER          ESTIMATE          T FOR HO          PR          T          STD ERROR OF
PARAMETER=0          ESTIMATE
INTERCEPT          -160.99422792          -5.13          0.1225          31.35873018
CC          122.15451936          6.84          0.0924          17.85864558
FT          59.10872222          10.05          0.0631          5.87911344

```

Fig.6.4 SAS printout for dependency of tuning capacitor value on  $f_T$  and  $C_c$  linearly.

SAS 1  
 11 23 FRIDAY, JULY 24, 1987  
 GENERAL LINEAR MODELS PROCEDURE

DEPENDENT VARIABLE CT

SOURCE	DF	SUM OF SQUARES	MEAN SQUARE	
MODEL	2	33.47759654	16.73879827	
ERROR	1	3.41490346	3.41490346	
CORRECTED TOTAL	3	36.89250000		

MODEL F = 4.90 PR F = 0.3042

R-SQUARE	C.V.	ROGT MSE	CT MEAN	
C.907436	15.4964	1.84794574	11.92500000	

SOURCE	DF	TYPE I SS	F VALUE	PR	F
CC	1	11.45833333	3.36	0.3181	
FT	1	22.01926321	6.45	0.2388	

SOURCE	DF	TYPE III SS	F VALUE	PR	F
CC	1	14.21833546	4.16	0.2901	
FT	1	22.01926321	6.45	0.2388	

PARAMETER	ESTIMATE	T FOR H0 PARAMETER=0	PR	T	STD ERROR OF ESTIMATE
INTERCEPT	0.04883285	0.01	0.9922		4.00861321
CC	4.65821326	2.04	0.2901		2.28288589
FT	1.90835735	2.54	0.2388		0.75153208

Fig.6.5 SAS printout for dependency of external  $Q$  value on  $f_T$  and  $C_c$  linearly.



Cc. Increase in the values of  $f_T$  and Cc seem to increase the value of Q. The model used is given as follows -

$$Q = A4 + B4 Cc + C4 f_T$$

From the results the model for Q can be written as follows-

$$Q = -160.99422792 + 122.15451936 Cc + 59.10872222 f_T$$

### FUTURE WORK

The first area for the next graduate student to work on is the improvement of the high frequency model of a transistor. In the present research work, an empirical rule was used, based on Vendelin's model, to account for the distributed nature of Cc and  $r_{bb'}$ . The ratios of different capacitor and resistor values in terms of Cc and  $r_{bb'}$ , respectively were based on the values given for a particular transistor. A future graduate student can work on developing a more accurate rule to account for the distributed nature of Cc and  $r_{bb'}$ . The second area to work on is the dependence of feedback length on the transistor parameters, especially on Cc. From the present research work, it can be concluded that the feedback line length does not seem to depend on the  $f_T$  of the transistor which was assumed before the start of this work. However it appears that the feedback line length does depend on the output capacitance of the transistor. Hence it seems to be worthwhile for the next student to study the effect

of  $C_c$  on feedback line length, for a larger number of transistors than used in this research. In addition, other parameters, such as  $h_{fe}$ , should be tested.

A P P E N D I X  
-----

APL FUNCTION TO CALCULATE S AND Y PARAMETERS.

```

▽TRANSIST[0]▽
[0] M←TRANSIST F;A;B;C;D;MI
[1] W←0.2×F
[2] WT←0.2×FT
[3] C1←0.294643×CC
[4] C2←0.3125×CC
[5] C3←0.25×CC
[6] C4←0.142857×CC
[7] RBB←TC÷CC
[8] R1←0.02469×RBB
[9] R2←0.4321×RBB
[10] R3←0.54321×RBB
[11] GM←IC÷2E
[12] REE←1÷GM
[13] RP←HFE×REE
[14] CP←1÷(REE×WT)
[15] A←RB÷((RB×RB)+(W×W)×(LB×LB))
[16] B←(W×LB)÷((RB×RB)+(W×W)×(LB×LB))
[17] C←RE÷((RE×RE)+(W×W)×(LE×LE))
[18] D←(W×LE)÷((RE×RE)+(W×W)×(LE×LE))
[19] M← 2 7 7 0
[20] M[1;1;1]←A
[21] M[1;1;3]←-A
[22] M[1;2;2]←1÷R4
[23] M[1;2;6]←GM

```

contd.-----

[24]  $M[1; 2; 7] \leftarrow -GM$   
 [25]  $M[1; 3; 1] \leftarrow -A$   
 [26]  $M[1; 3; 3] \leftarrow A + (1 \div R1)$   
 [27]  $M[1; 3; 4] \leftarrow ^{-1} \div R1$   
 [28]  $M[1; 4; 3] \leftarrow ^{-1} \div R1$   
 [29]  $M[1; 4; 4] \leftarrow (1 \div R1) + (1 \div R2)$   
 [30]  $M[1; 4; 5] \leftarrow ^{-1} \div R2$   
 [31]  $M[1; 5; 4] \leftarrow ^{-1} \div R2$   
 [32]  $M[1; 5; 5] \leftarrow (1 \div R2) + (1 \div R3)$   
 [33]  $M[1; 5; 6] \leftarrow ^{-1} \div R3$   
 [34]  $M[1; 6; 5] \leftarrow ^{-1} \div R3$   
 [35]  $M[1; 6; 6] \leftarrow (1 \div R3) + (1 \div RP)$   
 [36]  $M[1; 6; 7] \leftarrow ^{-1} \div RP$   
 [37]  $M[1; 7; 6] \leftarrow (^{-1} \div RP) - GM$   
 [38]  $M[1; 7; 7] \leftarrow (1 \div RP) + GM + C$   
 [39]  $M[2; 1; 1] \leftarrow -B$   
 [40]  $M[2; 1; 3] \leftarrow B$   
 [41]  $M[2; 2; 2] \leftarrow W \times (C1 + C2 + C3 + C4)$   
 [42]  $M[2; 2; 3] \leftarrow -W \times C1$   
 [43]  $M[2; 2; 4] \leftarrow -W \times C2$   
 [44]  $M[2; 2; 5] \leftarrow -W \times C3$   
 [45]  $M[2; 2; 6] \leftarrow -W \times C4$   
 [46]  $M[2; 3; 1] \leftarrow B$   
 [47]  $M[2; 3; 2] \leftarrow -W \times C1$   
 [48]  $M[2; 3; 3] \leftarrow -B + (W \times C1)$

contd.-----

```

[49]  M[2;4;2]←-W×C2
[50]  M[2;4;4]←W×C2
[51]  M[2;5;2]←-W×C3
[52]  M[2;5;5]←W×C3
[53]  M[2;6;2]←-W×C4
[54]  M[2;6;6]←W×(CP+C4)
[55]  M[2;6;7]←-W×CP
[56]  M[2;7;6]←-W×CP
[57]  M[2;7;7]←(W×CP)-D
[58]  MI←(CMINVM)÷50
[59]  A1←MI[;1;1]-(1,0)
[60]  A2←MI[;1;1]+(1,0)
[61]  A3←MI[;2;2]+(1,0)
[62]  A4←MI[;2;2]-(1,0)
[63]  A5←MI[;1;2]CMUL MI[;2;1]
[64]  A6←((A2 CMUL A3)-A5)
[65]  S11←((A1 CMUL A3)-A5)CDIV A6
[66]  S12←2×MI[;1;2]CDIV A6
[67]  S21←2×MI[;2;1]CDIV A6
[68]  S22←((A2 CMUL A4)-A5)CDIV A6
[69]  S11A←POLAR S11
[70]  S11A[2;]←180×S11A[2;]÷01
[71]  S12A←POLAR S12
[72]  S12A[2;]←180×S12A[2;]÷01
[73]  S21A←POLAR S21

```

contd.-----

```

[74] S21A[2;]←180×S21A[2;]÷01
[75] S22A←POLAR S22
[76] S22A[2;]←180×S22A[2;]÷01
[77] MIA←(MI[;1;1])×50
[78] MIB←(MI[;1;2])×50
[79] MIC←(MI[;2;1])×50
[80] MID←(MI[;2;2])×50
[81] A7←MIA CMUL MID
[82] A8←MIB CMUL MIC
[83] A9←A7-A8
[84] Y11←MID CDIV A9
[85] Y12←-MIB CDIV A9
[86] Y21←-MIC CDIV A9
[87] Y22←MIA CDIV A9
[88] Y← 2 2 2 e0
[89] X←02×F×DL÷V
[90] Y[2;1;1]←-Y0÷30X
[91] Y11L←Y[;1;1]
[92] Y[2;1;2]←Y0×30X
[93] Y12L←Y[;1;2]
[94] Y[2;2;1]←Y0×30X
[95] Y21L←Y[;2;1]
[96] Y[2;2;2]←-Y0÷30X
[97] Y22L←Y[;2;2]
[98] Y11C←Y11L+Y11

```

contd.-----

```

[99]  Y11CA←Y11C×500
[100] Y12C←Y12L+Y12
[101] Y12CA←Y12C×500
[102] Y21C←Y21L+Y21
[103] Y21CA←Y21C×500
[104] Y22C←Y22L+Y22
[105] Y22CA←Y22C×500
[106] AA←(1,0)-Y11CA
[107] BB←(1,0)+Y11CA
[108] CX←(1,●)-Y22CA
[109] DD←(1,0)+Y22CA
[110] EE←Y21CA CMUL Y12CA
[111] FF←(BB CMUL DD)-EE
[112] S11T←((AA CMUL DD)+EE)CDIV FF
[113] S12T←(-2×Y12CA)CDIV FF
[114] S21T←(-2×Y21CA)CDIV FF
[115] S22T←((BB CMUL CX)+EE)CDIV FF
[116] S11P←POLAR S11T
[117] S11P[2;]←180×S11P[2;]÷01
[118] S12P←POLAR S12T
[119] S12P[2;]←180×S12P[2;]÷01
[120] S21P←POLAR S21T
[121] S21P[2;]←180×S21P[2;]÷01
[122] S22P←POLAR S22T
[123] S22P[2;]←180×S22P[2;]÷01
[124] YOUT←Y22C-((Y12C CMUL Y21C)CDIV Y
11C)

```

\*\*\*\*\*

## OSCILLATOR CIRCUIT.

```

▽OSC[Q]▽
[0] XZ←OSC F;M
[1] XZ← 1 2 ρ0
[2] M←TRANSIST F
[3] YCTR←(0, W×CTR)
[4] YCT←(0, W×CT)
[5] YCT1←(0, W×CT1)
[6] YST←(0, -Y0÷30(W×DSTB÷V))
[7] Y11CM←Y11C+YCT
[8] Y12CM←Y12C
[9] Y21CM←Y21C
[10] Y22CM←Y22C+YCTR+YST+YCT1
[11] Y11CMP←Y11CM×500
[12] Y12CMP←Y12CM×500
[13] Y21CMP←Y21CM×500
[14] Y22CMP←Y22CM×500
[15] AAA←(1, 0)-Y11CMP
[16] BBB←(1, 0)+Y11CMP
[17] CXC←(1, 0)-Y22CMP
[18] DDD←(1, 0)+Y22CMP
[19] EEE←Y21CMP CMUL Y12CMP
[20] FFF←(BBB CMUL DDD)-EEE
[21] S11CM←((AAA CMUL DDD)+EEE)CDIV FFF
[22] S12CM←(-2×Y12CMP)CDIV FFF
[23] S21CM←(-2×Y21CMP)CDIV FFF
[24] S22CM←((BBB CMUL CXC)+EEE)CDIV FFF
[25] S11CMP←POLAR S11CM
[26] S11CMP[2;]←180×S11CMP[2;]÷01
[27] S22CMP←POLAR S22CM
[28] S22CMP[2;]←180×S22CMP[2;]÷01
[29] S12CMP←POLAR S12CM
[30] S12CMP[2;]←180×S12CMP[2;]÷01
[31] S21CMP←POLAR S21CM
[32] S21CMP[2;]←180×S21CMP[2;]÷01
[33] YOUT1←Y22CM-((Y21CM CMUL Y12CM)CDI
V Y11CM)

```



APL FUNCTIONS TO CALCULATE OPTIMUM FEEDBACK LINE LENGTH  
STUB LINE LENGTH AND CARRY OUT FREQUENCY RESPONSE.

▽FREQRSP[0]▽

```
[0] Z←FREQRSP F;K;N;M
[1] N←ρF←F×1000000 ⋄ K←1
[2] Z← 2 0 ρ0
[3] L1:M←TRANSIST F[K] ⋄ Z←Z,((2,1)ρ(,Y
22)) ⋄ →(N₂K←K+1)/L1
```

▽FBLENGTH[0]▽

```
[0] Z←FBLENGTH U;K;N;XZ;M
[1] N←ρU ⋄ K←1
[2] Z← 2 0 ρ0
[3] L1:DL←U[K] ⋄ XZ←OSC F ⋄ Z←Z,((2,1)ρ
(,YOUT1)) ⋄ →(N₂K←K+1)/L1
```

▽STLENGTH[0]▽

```
[0] SL←STLENGTH UL;K;N;Z
[1] N←ρUL ⋄ K←1
[2] SL← 2 0 ρ0
[3] L1:DSTB←UL[K] ⋄ XZ←OSC F ⋄ SL←SL,((
2,1)ρ(,YOUT1)) ⋄ →(N₂K←K+1)/L1
```

▽COVER[0]▽

```
[0] Z←COVER F;K;N;XZ;M
[1] N←ρF←F×1000000 ⋄ K←1
[2] Z← 2 0 ρ0
[3] L1:XZ←OSC F[K] ⋄ Z←Z,((2,1)ρ(,YOUT1
)) ⋄ →(N₂K←K+1)/L1
```

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