Dedication

This project is for Evelyn, whose love and encouragements keeps me going just fine.

Acknowledgement

My acknowledgements go to Dr.-Ing. Wilfred N. Mwema, for his inspiration and support.

Abstract

The objective of this project was to design a wideband small-signal Microwave amplifier to operate at 2.4 GHz ISM Band, with at least 20% fractional Bandwidth and an input impedance of 300Ω. Broadband amplifier design usually involves mismatching the output and/or input impedance of the amplifier. In this project, S parameters were obtained for BFP420 at the frequency range of interest, that is, between 2.16GHz and 2.64GHz. Simultaneous conjugate matching technique was applied at the early stages of the design in order to determine if the maximum gain was large enough so that feedback topologies could be used in the network for broadbanding purposes. This was also done to determine the initial bandwidth. Broadbanding methods were then applied to the amplifier, and after several software simulations, an amplifier with a flat gain spanning the required range of frequencies was designed. The input impedance of 300Ω was achieved by examining the relationship between the transistor transconductance and the dynamic input impedance r_{be}. Shunting effect of a feedback resistor R_F at the input yielded the desired Z_{in}.

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CHAPTER 1: INTRODUCTION

1.1 Objectives

The goal of this design project was to design a wideband small-signal Microwave amplifier operated at 2.4 GHz ISM (Industrial, Scientific, and Medical) Band, and with a fractional bandwidth of 20% and an input impedance of 300Ω. The 300Ω input impedance was to act as an antenna loading seen at the base of the amplifier.
This project was chosen because of its apparent complexity and the RF design experience that would be gained by the end of the project.

1.2 Scope

The project spans the area of wideband amplifier design with special emphasis in bandwidth-improvement techniques and gain-flattening procedures. Generally, the design of microwave amplifiers (large signal or small signal) revolves around fundamental concepts such as the use of transistor S parameters and impedance matching techniques. Simultaneous conjugate matching works well for designs where maximum transducer power gain is to be achieved while constant power gain circles and noise figure circles aid in designing for fixed power gains and Low Noise Amplifiers, respectively.

Several techniques are present in literature that can be used to obtain broadband amplifiers spanning the frequency range of interest. Reactive matching or mismatch approach has the advantage that it uses lossless elements, but the resultant network suffers extremely poor impedance matching. Dissipative mismatch at the input or output of the amplifier is mostly employed where the amplifier gain is large enough to be sacrificed. This technique has the disadvantage of degrading noise figure, but it is excellent in gain-flattening and shaping. One very clear thing about this method is that resistances never really match, but merely cover up mismatch. Negative feedback method has several advantages including gain reduction and stabilization, besides broadbanding. Other methods of wideband design such as balanced amplifiers and distributed amplifiers are useful where high degree of stability, flat gain, noise figure, or where cascade networks are required. Their major drawback is their complexity.

This project combines reactive matching and negative feedback design methods in order to achieve its objectives. Since the transistor chosen for this project was found to be absolutely

stable as its S parameter relations proved, conjugate matching was first used to obtain the general view of the amplifier gain response and bandwidth. Broadband design methods were then employed. Active biasing was used to obtain the correct dc operating point.
Most microwave designs are carried out using CAD tools, and this was no exception. Both design and simulation happened in a software environment. AWR Microwave Office was used.

Literature review of this project is covered in Chapter 2 in a broad perspective. It covers the basic concepts necessary in understanding the design of Microwave Amplifiers. Chapter 3 presents the design methodology and the results of the design. Analysis of simulation results is covered in Chapter 4 while Chapter 5 carries conclusion and proposed future work.

CHAPTER 2: LITERATURE REVIEW

2.1 FUNDAMENTAL CONCEPTS IN MICROWAVE AMPLIFIER DESIGN

2.1.1 Introduction

An amplifier is a circuit designed to enlarge electrical signals. Microwave amplifiers are used mostly in telecommunication transmitters and receivers, as shown in $Fig. 1$.

Amplifier applications may require minimum noise, maximum gain, and maximum output power, best impedance matching, stability into varying loads, wide bandwidth, cascading with other circuits, and other performance factors.

Low noise amplifiers are used at the front end of receivers. They are usually approximated as small signal devices, and are usually *tuned* (i.e. they use networks at their input and output to provide a match and gain over a relatively narrow bandwidth).

Power amplifiers are used at the output of transmitters. They provide a high output power, and so cannot be approximated as small signal. They are designed using nonlinear active devices while small signal amplifiers are designed with linear active components.

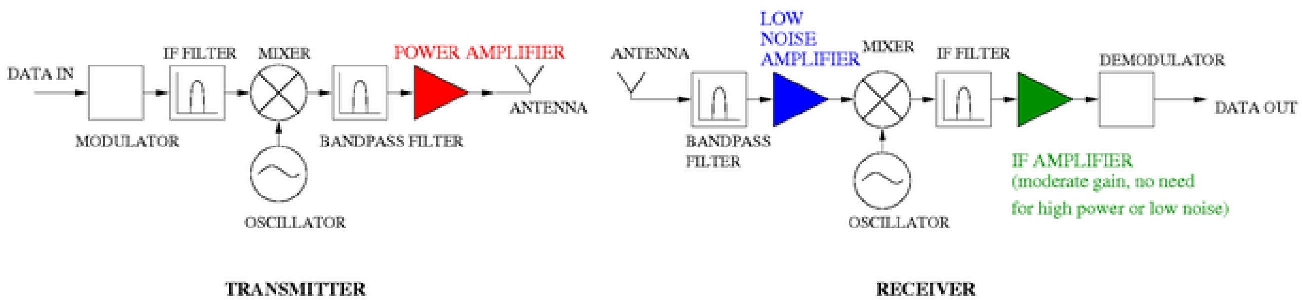

Fig 2.1 Transmitter and Receiver block diagrams showing applications of RF/Microwave LNA and Power Amplifier.

2.1.2 Scattering Parameters

Voltages and currents are difficult to measure at microwave frequencies because they are distributed values and vary with their position in microwave structures. In fact, the widely spread current in a waveguide is virtually impossible to measure directly since direct measurements ustually involve magnitude (inferred from power) and phase of a wave travelling in a specified direction, or a standing wave.

Thus, equivalent voltages and currents, and impedance and admittance matrices, become somewhat an abstraction when dealing with high frequency networks. Waves are more easily measured in microwave networks.

Scattering parameters give representation more in accord with direct measurement and the ideas of incident, reflected and transmitted waves. Fig 2.2 illustrates the concept of scattering network.

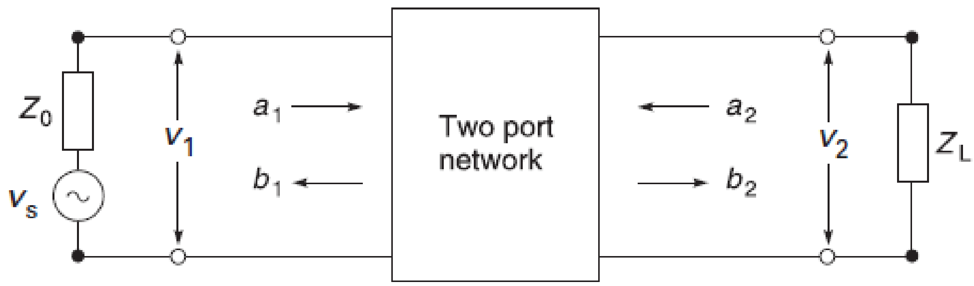

Fig 2.2 Two port scattering network with source and load.

Scattering (S) parameters characterize a network in terms of incident and reflected waves.
In fig 2.2, a_1 and a_2 represent incident voltage waves, while $b_1$and b_2 represent reflected voltage waves.
These four waves are related by the equations where S_{11}, S_{21}, $S_{12,}$ and S_{22}, are the scattering or S parameters.

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \qquad 2.1$$

$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_{2=0}}$ Input reflection coefficient with output properly terminated.

$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_{2=0}}$ Forward transmission coefficient with output properly terminated.

$S_{12} = \left.\frac{b_1}{a_2}\right|_{a_{1=0}}$ Reverse transmission coefficient with input properly terminated.

$S_{22} = \left.\frac{b_2}{a_2}\right|_{a_{1=0}}$ Output reflection coefficient with input properly terminated.

From fig 2.2, it is evident that:

$$S_{11} = \frac{b_1}{a_1} = \frac{V_1/I_1 - Z_0}{V_1/I_1 + Z_0} = \frac{Z_1 - Z_0}{Z_1 - Z_0} \quad 2.2$$

$$Z_1 = Z_0 \frac{1 + S_{11}}{1 - S_{11}} \quad 2.3$$

These equations show some of the advantages of S-parameters in the design of microwave amplifiers:

- They are simply power gains and reflection coefficients.
- They are measured under matched terminations.

$$|s_{11}|^2 = \frac{power\ reflected\ from\ the\ network\ input}{power\ incident\ on\ the\ network\ input} \quad 2.4$$

$$|s_{21}|^2 = \frac{power\ delivered\ to\ a\ Z_0\ load}{power\ available\ from\ a\ Z_0\ source} \quad 2.5$$

$$|S_{12}|^2 = Reverse\ transducer\ /insertion\ power\ gain\ with\ Z_0\ source\ and\ load \quad 2.6$$

$$|S_{22}|^2 = \frac{power\ delivered\ from\ the\ network\ output}{power\ incident\ on\ the\ network\ input} \quad 2.7$$

- S-Parameters are defined and measured relative to fixed system impedance, $Z_{0,}$usually 50Ω.
- In microwave transistors, S-parameters are determined at specific bias conditions because these parameters are bias-dependent.
- S-Parameters also depend on operating temperatures and applied signal levels. They apply to steady-state conditions only.
- Small-Signal Microwave Amplifiers are designed using S-Parameters.

2.1.3 Power Gains

For amplifiers functioning at RF and microwave frequencies, usually of interest is the input and output power relation.

Power gain is preferred for high frequency amplifiers because the impedance encountered is usually low due to parasitic capacitance.

For amplifiers functioning at lower frequency such as IF, it is the voltage gain that is of interest, since the impedance encountered is high (less parasitic).

By working with power gain RF and Microwave designers are free from the constraint of system impedance. Fig 2.3 shows the power components in an amplifier.

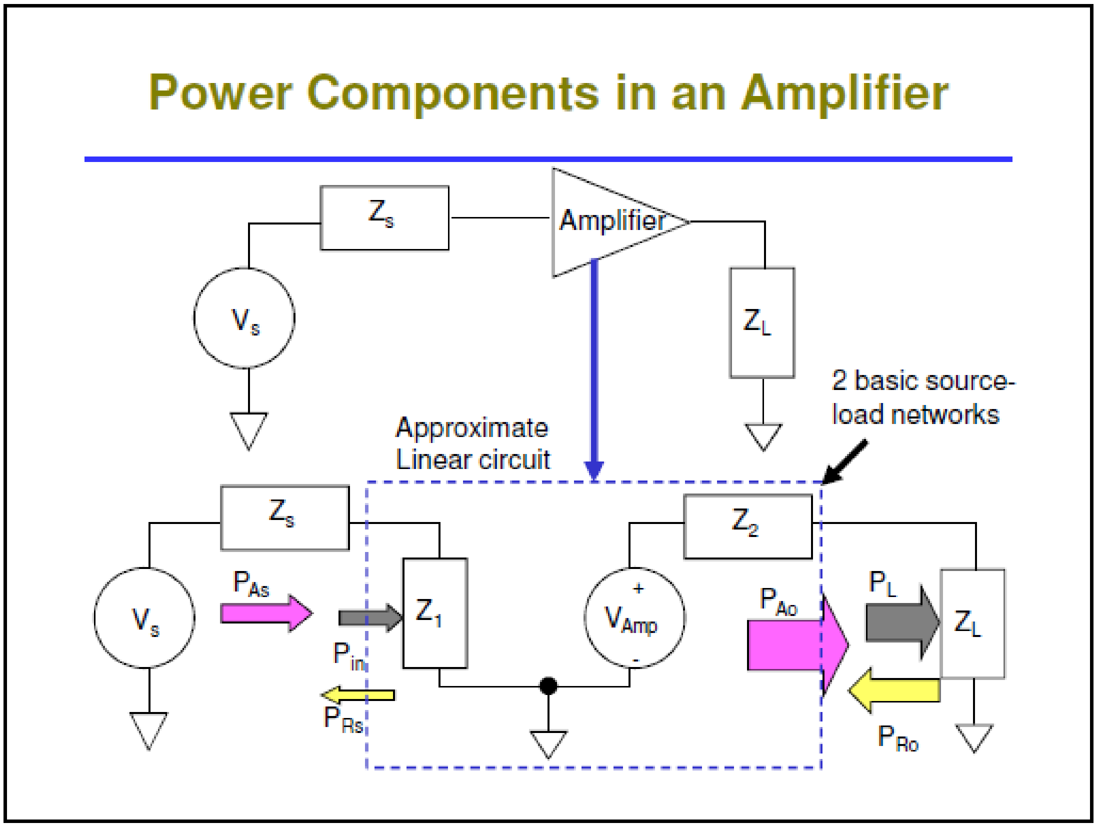

Fig. 2.3

From the power components, three types of power gain can be defined.

$$G_T = \frac{P_L}{P_{A_S}} = \frac{power\ delivered\ to\ load}{power\ available\ from\ input} \qquad 2.8$$

$$G_P = \frac{P_L}{P_{in}} = \frac{power\ delivered\ to\ load}{power\ input\ to\ amplifier} \qquad 2.9$$

$$G_A = \frac{P_{A_O}}{P_{A_S}} = \frac{power\ available\ to\ load}{power\ available\ at\ input} \qquad 2.10$$

G_T is the transducer power gain of the amplifier, G_P, the operational gain, while G_A is the available gain. G_T is the effective amplifier gain for simultaneously conjugate matched input and output ports, which leads to maximum small-signal power gain. G_P and G_A are for maximum linear output power and low noise amplifier, respectively.

- G_T is the relevant indicator of the amplifying capability of the amplifier.
- Whenever an amplifier is designed to a specific power gain, the gain of concern is the transducer power gain G_T.
- G_P and G_A are usually used in the process of amplifier synthesis to meet a certain G_T.
- An amplifier can have a large G_P or G_A and yet small G_T.

Finding the transducer power gain requires knowledge of the S-parameters, as well as the source and load terminations connected to the two-port. During linear circuit simulation, the source and load terminations are either given or computed from the circuit topology description.

The two-port's S-parameters are either specified or computed from a linear device model.

The amplifier gains defined above are functions of S-parameters, and can be written in the form;

$$G_p = \frac{|S_{21}|^2[1 - |\Gamma_L|^2]}{|1 - S_{22}\Gamma_L|^2[1 - |\Gamma_{IN}|^2]} \qquad 2.11$$

$$G_A = \frac{|S_{21}|^2[1 - |\Gamma_S|^2]}{|1 - S_{11}\Gamma_S|^2[1 - |\Gamma_{OUT}|^2]} \qquad 2.12$$

$$G_T = \frac{1 - |\Gamma_s|^2}{|1 - \Gamma_{IN}\Gamma_s|^2}|S_{21}|^2\frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} \qquad 2.13$$

$$where\ \Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad 2.14$$

Alternatively,

$$G_T = \frac{1 - |\Gamma_s|^2}{|1 - S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1 - |\Gamma_L|^2}{|1 - \Gamma_{OUT}\Gamma_L|^2} \qquad 2.15$$

$$where\ \Gamma_{OUT} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} \qquad 2.16$$

In the above expressions, Γ_{IN} represents the true input reflection coefficients of the two-port, with an arbitrary load termination Γ_L. Similarly, Γ_{OUT} stands for the output reflection coefficient of the two-port, with an arbitrary source termination connected to the input (Fig. 2.4)

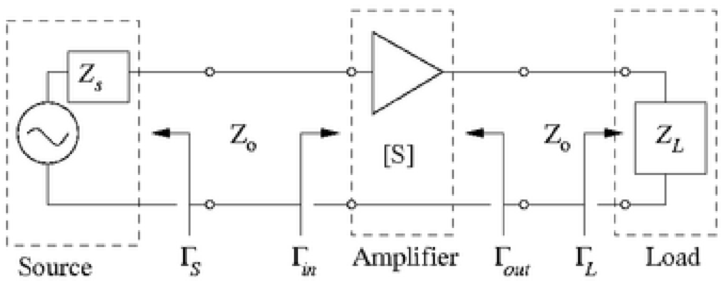

Fig. 2.4

Transducer gain can be broken up into three subexpressions:

$$G_T = \frac{1 - |\Gamma_s|^2}{|1 - \Gamma_{IN}\Gamma_s|^2}|S_{21}|^2\frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} = G_{1D}G_0G_2 \qquad 2.17$$

$where\ G_{1D} = \frac{1-|\Gamma_s|^2}{|1-\Gamma_{IN}\Gamma_s|^2}$ is the transducer gain-factor change due to the selection of $\Gamma_S and\ \Gamma_L$

$G_0 = |S_{21}|^2$ is the intrinsic gain of the amplifier and would equal the transducer power gain if both $\Gamma_s and\ \Gamma_L$ were equal.

$$G_2 = \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2}\ is\ the\ change\ of\ the\ transducer\ gain\ due\ to\ load\ selection\ \Gamma_L$$

When there is no interaction between the input ports, then ($|S_{12} = 0|$), and this introduces unilateral condition.

Under unilateral condition, $\Gamma_{IN} = S_{11}$ and $\Gamma_{OUT} = S_{22}$

Unilateral transducer gain is now given by (2.18),

$$G_{TU} = \frac{1-|\Gamma_s|^2}{|1-S_{11}\Gamma_s|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} \qquad 2.19$$

In practical microwave amplifier design, especially at frequencies above 1GHz, $S_{12} \neq 0$, and unilateral technique is not pursued.

2.1.4 Stability Analysis

Amplifier stability analysis is necessary to determine an amplifier's resistance to oscillations. In a stable amplifier, no output is produced when there is no input.

An amplifier is unstable when an output signal increases without any limit. Actually, nonlinearities do limit the maximum signal level and either set it to steady-state oscillation or stop it completely. Virtually, all RF/MW transistors are potentially unstable at some frequencies.

In low-frequency analog circuits, where transfer functions are commonly available, the Nyquist criteria provide a safe indication of system stability. System design at RF/MW frequencies is much more difficult and tedious because transfer functions are virtually never given in closed form. Hence, a thorough stability analysis is performed through a wide range of frequencies, input signal levels and external terminations.

Since true broadband nonlinear models are not also available for the active devices, RF/MW circuit stability is most conveniently evaluated at individual frequencies, based on small-signal two port parameters.

Stability analysis is carried out by assuming a small-signal amplifier, since the initial signal that causes oscillation is always very small. Stability of an amplifier is affected by the load and source impedance connected to its two ports.

Oscillations in an amplifier are unwanted for the reasons listed below:

- When oscillation takes place, the active device is pushed into its large-signal mode and the performance changes very significantly.
- The small-signal *S*-parameters are no longer valid, and therefore, the circuit design is incorrect.
- When a device oscillates it becomes noisier.
- Even if the oscillation is far below the passband of the amplifier, the newly created signal mixes with any incoming signal and shows up at the output.

- Oscillation may damage the active device.

The idea presented in the box below is intuition-based [1], though *not always correct,* and explains how oscillations may build up between an active and a passive port.

Two-port circuits may start up oscillations if reflected signals, either at the input or output port, increase their magnitudes while they are continuously reflected between an active port and its termination. Such conditions often occur far below the passband frequency of an amplifier, where the transistors have high gain and the terminations seen by the device are far from 50Ω.

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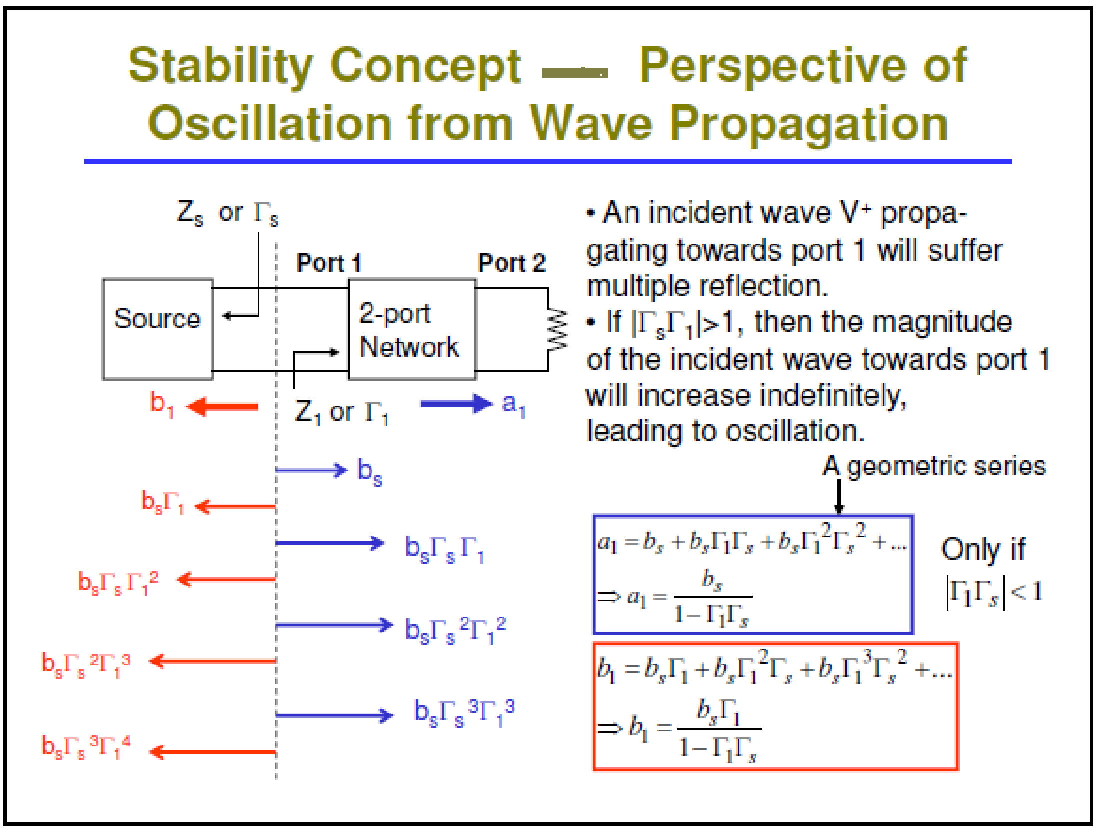

Fig 2.5

It is evident from the above analysis that for oscillations to occur,

$$|\Gamma_S\Gamma_1| > 1 \qquad 2.20$$

Since the source network is always passive,

$$|\Gamma_S| < 1 \qquad 2.21$$

Hence, for oscillations to occur,

$$|\Gamma_1| > 1 \qquad 2.22$$

And for the output network, oscillations will occur when,

$$|\Gamma_2| > 1 \qquad 2.23$$

Since, $|\Gamma_L| < 1$ 2.24

Mathematically, it is, thus, deducible (from 2.14) that to prevent oscillations,

$$|\Gamma_1| = |\Gamma_{IN}| \quad = \left|S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}\right| < 1 \qquad 2.25$$

And, $|\Gamma_2| = |\ \Gamma_{OUT}| = \left|S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}\right| < 1$ 2.26

A two-port network is said to be unconditionally stable at a given frequency if, $Re(Z_{IN}) > 0$ and $Re(Z_{OUT}) > 0$ for all passive load and source terminations [2].

If a two-port network is not unconditionally stable, it is potentially unstable, that is, some passive load and source terminations can produce input and output impedances having a negative real part [2].

In terms of reflection coefficients, equations (2.21) to (2.26) give the conditions for unconditional stability.

For any linear two-port network, there exists a stability factor (Roulette's Stability Factor), K, that gives a quick check on the circuit's stability status.

K is given in terms of S-parameters, and for unconditional stability,

$$K = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}||S_{12}|} > 1 \qquad 2.27$$

Where Δ is the determinant of the Scattering Matrix and is given by (2.28)

$$\Delta = S_{22}S_{11} - S_{12}S_{21} \qquad 2.28$$

For unconditional stability, $|\Delta| < 1$

Roulette's K factor does not indicate the relative stability of various devices. The μ −factor does.

$$\mu_1 = \frac{1 - |S_{22}|^2}{|S_{11} - \Delta(S_{22}{}^*)| + |S_{12}S_{21}|} \qquad 2.29$$

For stability, $\mu > 1$, but larger values indicate greater stability.

There is also μ_2 given by

$$\mu_2 = \frac{1 - |S_{11}|^2}{|S_{22} - \Delta(S_{11}{}^*)| + |S_{12}S_{21}|} \qquad 2.30$$

When $\mu_1 > 1$, then, $\mu_2 > 1$ and vice versa.

The μ-factors have very meaningful physical interpretations [1]: μ_1 is the distance between the center of the Smith chart and the unstable region of the load stability circle, while μ_2 shows how far the unstable region of the source stability circle is from the center of the Smith chart.

While the stability factor, K, is only an analytical definition the μ-factors show exactly how far the regions of unstable terminations are from the center of the Smith chart. If the magnitudes of the μ-factors are greater than unity, then any termination on the Smith chart may be used safely. This illustration shows the definition of μ_1, generally referred to just as μ.

Since all stability tests are based on frequency-dependent small-signal *S*-parameters, it is easy to see that two-port stability changes with frequency. Generally, active devices are stable at the very low frequencies where $|S_{12}|$ is very small, and also at the very high frequencies where $|S_{21}|$ rolls off. Unfortunately (for RF/MW amplifier designers) there is a wide range of RF/MW frequencies where the possibility of oscillation is a threat to stable operation.

The stability factor is also a function of dc bias settings and the signal level. When the applied signal level begins to compress the gain of the device, the *S*-parameters change and so does the stability factor.

For potentially unstable transistors, stability analysis is carried out graphically. When a two-port network is potentially unstable, there may be values of Γ_S and Γ_L (i.e. source and load impedances) for which the real parts of Z_{IN} and Z_{OUT} are positive [2]. These values of Γ_S and Γ_L (i.e. regions in the Smith Chart) can be determined using stability circles.

The regions where Γ_S and Γ_L produce $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$ are determined respectively. Setting the magnitudes of 2.25 and 2.26 equal to 1, and solving for Γ_S and Γ_L shows that the solutions of Γ_S and Γ_L lie on circles (stability circles) whose equations are given by(2.31) and(2.32)

$$\left|\Gamma_L - \frac{(S_{22} - \Delta S_{11}{}^*)^*}{|S_{22}|^2 - |\Delta|^2}\right| = \left|\frac{S_{21}S_{12}}{|S_{22}|^2 - |\Delta|^2}\right| \qquad 2.31$$

$$\left|\Gamma_S - \frac{(S_{11} - \Delta S_{22}{}^*)^*}{|S_{11}|^2 - |\Delta|^2}\right| = \left|\frac{S_{21}S_{12}}{|S_{11}|^2 - |\Delta|^2}\right| \qquad 2.32$$

$$where\ \Delta\ is\ as\ defined\ in\ (2.28)$$

(2.31) and (2.32) give the radii and centers of the circles where $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$ in the Γ_L and Γ_S plane, respectively.

$\Gamma_L\ values\ for |\Gamma_{IN}| = 1 (Output\ Stability\ Circle)$:

$$r_L = \left|\frac{S_{21}S_{12}}{|S_{22}|^2 - |\Delta|^2}\right| (RADIUS) \qquad 2.33$$

$$C_L = \left|\frac{(S_{22} - \Delta S_{11}{}^*)^*}{|S_{22}|^2 - |\Delta|^2}\right| (CENTRE) \qquad 2.34$$

$\Gamma_S\ values\ for\ |\Gamma_{OUT}| = 1 (Input\ Stability\ Circle)$:

$$r_S = \left|\frac{S_{21}S_{12}}{|S_{11}|^2 - |\Delta|^2}\right| (RADIUS) \qquad 2.35$$

$$C_S = \left|\frac{(S_{11} - \Delta S_{22}{}^*)^*}{|S_{11}|^2 - |\Delta|^2}\right| (CENTRE) \qquad 2.36$$

Fig 2.6 illustrates the graphical construction of stability circles where $|\Gamma_{IN}| = 1$ and $|\Gamma_{OUT}| = 1$. On one side of the stability circle boundary, in the $\Gamma_L\ plane$, $|\Gamma_{IN}| < 1$, and on the other side $|\Gamma_{IN}| > 1$. Similarly, in the $\Gamma_S\ plane$, on one side of the stability circle boundary $|\Gamma_{OUT}| < 1$, and on the other side $|\Gamma_{OUT}| > 1$.

Fig 2.7 illustrates stable and unstable regions of the output stability circle. The shaded regions are stable.

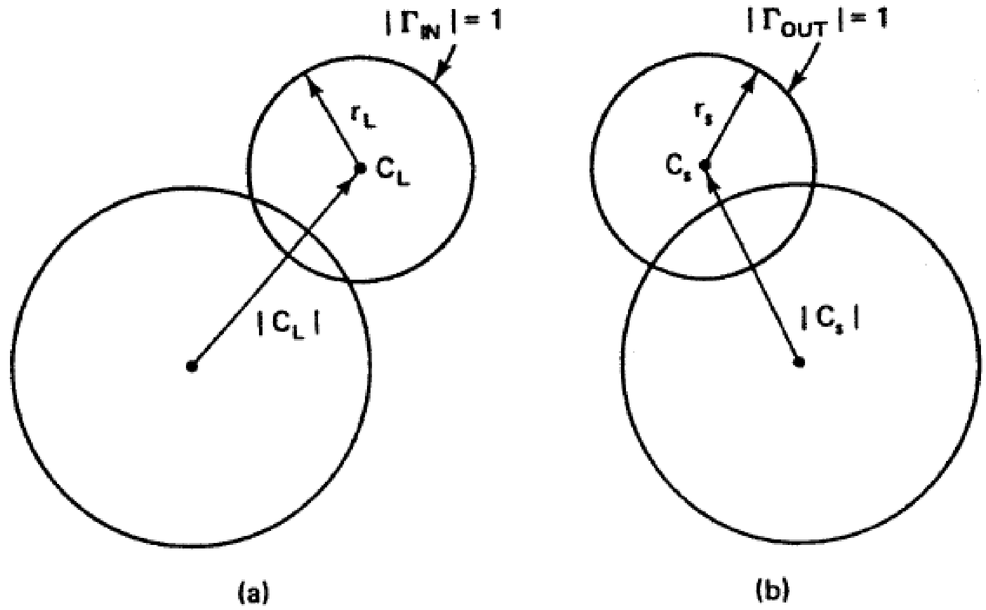

Figure 2.6 **Stability circle construction in the Smith chart: (a) Γ_L plane; (b) Γ_s plane.**

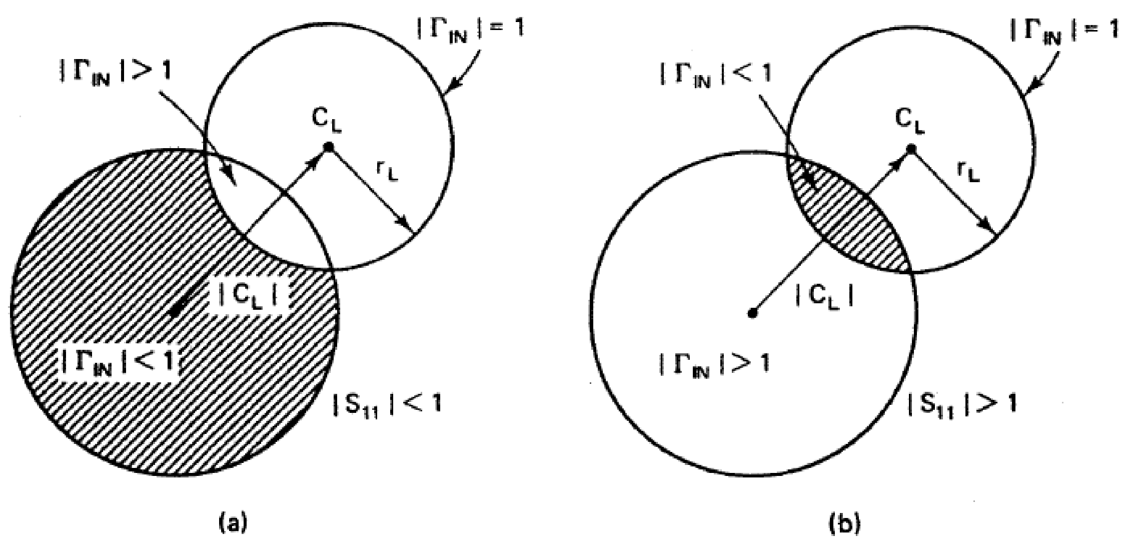

Figure 2.7 **Smith chart illustrating stable and unstable regions in the Γ_L plane.**

Conditions for Absolute Stability: No passive source or load termination can cause an amplifier to oscillate if $a, b, c, and\ d$ are all satisfied [3].

$$a. |S_{11}| < 1, |S_{22}| < 1 \quad 2.37$$

$$b. \left| \frac{|S_{21}S_{12}| - |M^*|}{|S_{11}|^2 - |\Delta|^2} \right| > 1 \quad 2.38$$

$$c. \left| \frac{|S_{21}S_{12}| - |N^*|}{|S_{22}|^2 - |\Delta|^2} \right| > 1 \quad 2.39$$

$$where\ M = S_{11} - \Delta S_{22}{}^*$$
$$N = S_{22} - \Delta S_{11}{}^*$$

2.1.5 Noise in RF/MW Circuits

Even when a two-port network is linear, the output waveform will differ from the input, because of the failure to transmit all spectral components with equal gain (or attenuation) and delay. This kind of distortion can be avoided, for instance, by input bandwidth limitation. However, noise generated in a system can still change the output waveform.

In a passive two-port noise arises only from the losses in the circuit; thermodynamic considerations indicate that such losses result in the random changes called noise. A very important consideration in a system is the amount of noise that it adds to the transmitted signal.

Noise is a random phenomenon, and at RF/MW frequencies designers prefer to deal with noise power (instead of noise voltage or noise current) that may be combined from different sources.

In MW amplifiers, a small amount of voltage can be measured at the output even without the input—this is referred to as noise power [2].

Three main causes of electrical noise:

- Thermal, or Johnson noise, caused by the thermal agitation of free electrons in conductors. It exists even when there is no current flow. It is associated with resistor white noise.

$$\overline{v_n{}^2} = 4kTR\Delta f \quad 2.40$$

- Shot, or Schottky noise, caused by the random fluctuation of current flow in semiconductors; due to current flowing across the potential barrier in PN junction. Exists only in BJTs, not FETs. Exists only when there is current flow.

$$\overline{i_n^{\,2}} = 2qI_{DC}\Delta f \qquad 2.41$$

- Flicker, or 1/f noise, caused by fluctuation in the conductivity of the medium; caused by traps associated contamination and crystal defect. It's a low frequency noise, and exists when there is a current in the circuit.

$$i_n^{\,2} = K_I \frac{I_{DC}^a}{f^b}\Delta f \qquad 2.42$$

- Burst noise: not fully understood. Low frequency noise.

$$i_n^{\,2} = K_2 \frac{I_{DC}^c}{1+\left(f/f_c\right)^2}\Delta f \qquad 2.43$$

- Avalanche noise: due to avalanche breakdown in Zener diodes. Low frequency noise.

For most electronic systems, the electrical noise usually fulfills a condition called Wide-Sense Stationary.

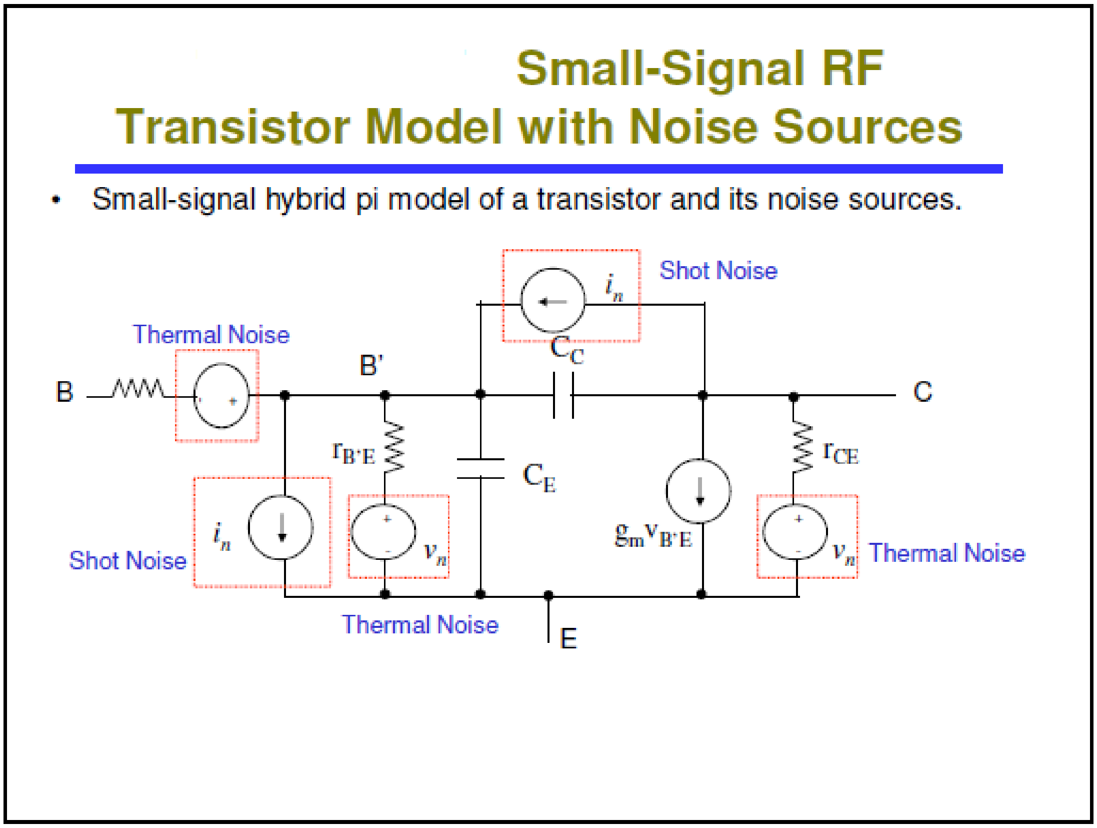

Fig 2.8

Important assumptions of noise in RF and Microwave networks:

- Amplitude of noise signal (either voltage or current) is usually small.
- The system where noise signal exists is linear.
- The noise signal is Wide-Sense Stationary and ergodic in the mean and auto-correlation.
- The PSD of the noise signal is white.
- The random variable resulting from sampling of the noise signal has Gaussian PDF.

From (2.40) *and* (2.41), it is deducible that:

- Reducing the bandwidth of the amplifier could in theory reduce the noise power at the output.
- Low noise design entails using small values of resistances.

In general,

- FETs do not have shot noise as the charge carriers in its channel do not flow through PN junction. Hence, they are usually used for amplifiers with very low noise requirement.

- Between using a discrete transistor and an integrated circuit (monolithic microwave integrated circuit, MMIC), usually a discrete transistor amplifier contribute lower noise to the systems (lower noise figure). This is evident as every component in the circuit contributes noise, the more the components; the higher is the total noise output of the circuit.
- Certain balanced configuration can reduce the noise contribution, for instance in double-balanced mixer design.
- Most RF small-signal amplifiers are also designed to be of low noise; the amplifiers introduce very little noise to the output. The amplifiers are important components in the receiver chains.

The total noise output power is composed of the amplified noise input power plus the noise output power contributed by the amplifier.

The model of a noisy two-port microwave amplifier is shown in Fig 2.8.

The noise input power can be modeled by a noisy resistor that produces thermal or Johnson noise. The rms value of the noise voltage, V_N produced by the noisy resistor, R_N over a frequency range $f_h - f_l$ is described by (2.44).

$$V_N = \sqrt{4kTBR_N} \qquad 2.44$$

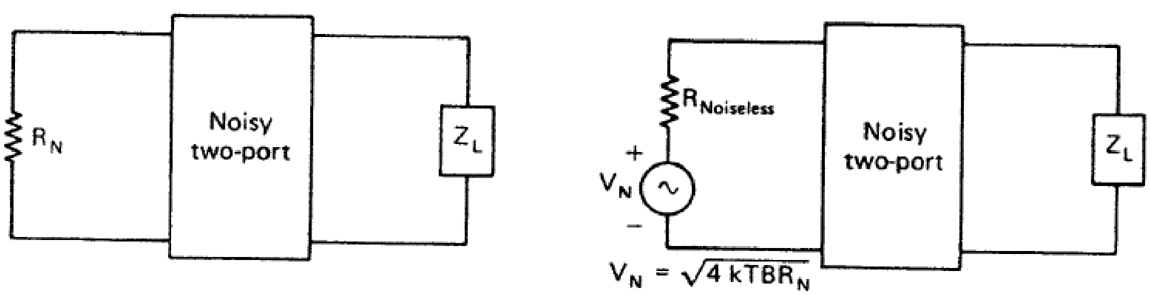

Figure 2.8 Model of a noisy microwave amplifier.

$where\ k\ is\ the\ Boltzmann\ constant (k = 1.374e-23\ J/°K),$
$T\ is\ the\ resistor\ noise\ temperature, and\ B\ is\ the\ noise\ bandwidth\ (B = f_h - f_l).$

Equation (2.44) shows that the thermal noise power depends on bandwidth and not on a given center frequency. Such a distribution of noise is called white noise.

The maximum available noise power from R_N is given by,

$$P_N = \frac{{V_N}^2}{4R_N} = kTB \qquad 2.45$$

The noise figure(F) describes quantitatively the performance of a noisy microwave amplifier. The noise figure can be expressed in the form,

$$F_{dB} = \frac{P_{N_0}}{P_{N_i}G_A} \qquad 2.46$$

$where P_{N_0} is\ the\ total\ available\ noise\ power\ at\ the\ amplifier\ output,\ P_{N_i}$
$= kTB\ is\ the\ available\ noise\ power\ due\ to\ R_N\ the\ bandwidth, and\ G_A is\ the\ available\ power\ gain.$

Since $G_A = \frac{P_{S_0}}{P_{S_i}}$ 2.47

$where\ P_{S_0}$ is the available signal power at the output $and\ P_{S_i}$ is the available signal power at the input.

Substituting (2.43) into(2.42), F becomes,

$$F_{dB} = \frac{P_{Si}/P_{Ni}}{P_{S_0}/P_{N_0}} \qquad 2.48$$

Equation(2.44) suggests that F is also defined as the ratio of the available signal-to-noise ratio at input to the available signal-to-noise ratio at output.

The noise figure of a two-port amplifier is described by,

$$F = F_{min} + \frac{r_n}{g_s}|Y_S - Y_o|^2 \qquad 2.49$$

$where\ r_n$ is the equivalent normalized noise resistance of the two-port (that is, $r_n = \frac{R_N}{Z_0}$) and $Y_S = g_s + jb_s$ represents the source admittance, and $Y_0 = g_0 - jb_0$ represents the output admittance which results in minimum noise figure.

The admittances can be expressed in terms of input and output reflection coefficients:

$$Y_S = \frac{1 - \Gamma_S}{1 + \Gamma_S} \qquad 2.50$$

$$Y_0 = \frac{1 - \Gamma_0}{1 + \Gamma_0} \qquad 2.51$$

Substituting(2.46) and (2.47) yields,

$$F = F_{min} + \frac{4r_n|\Gamma_S - \Gamma_0|^2}{1 - |\Gamma_S|^2|1 + \Gamma_0|^2} \qquad 2.52$$

F_{min} is a function of device operating current and frequency, and there is one value of Γ_0 associated with each F_{min}. The quantities F_{min} , $r_n and$ Γ_0 are noise parameters, and are given by the transistor manufacturer or are determined experimentally.

From (2.48) a noise figure parameter N_i can be defined as;

$$N_i = \frac{|\Gamma_S - \Gamma_0|^2}{1 - |\Gamma_S|^2} = \frac{F_i - F_{min}}{4r_n}|1 + \Gamma_0|^2 \qquad 2.53$$

Equation(2.49) can be used to obtain equation(2.50) which defines a family of circles with N_i as a parameter.

$$\left|\Gamma_s - \frac{\Gamma_0}{1 + N_i}\right|^2 = \frac{{N_i}^2 + N_i(1 - |\Gamma_0|^2)}{(1 + N_i)^2} \qquad 2.54$$

The family of circles defined by the above equation is known as constant noise figure circles, and the center is located at:

$$C_{Fi} = \frac{\Gamma_0}{1 + N_i} \qquad 2.55$$

The radii,

$$R_{Fi} = \frac{1}{1 + N_i}\sqrt{{N_i}^2 + N_i(1 - |\Gamma_0|^2)} \qquad 2.56$$

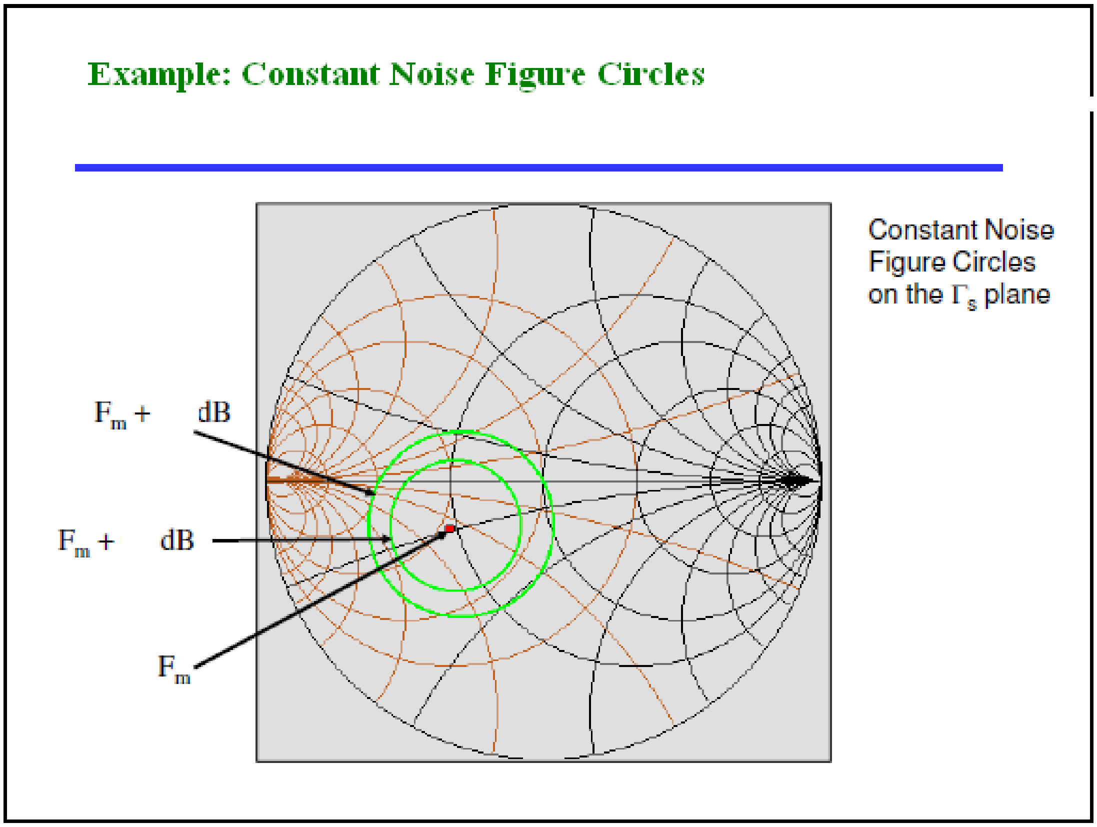

Fig 2. 10

2. 1. 6 The dc bias techniques

Most MW amplifier designers ignore dc biasing networks. While considerable effort is spent in designing for a given gain, noise figure, and bandwidth, little effort is spent in dc bias networks. The cost per decibel of MW power gain or noise figure is high, and designers cannot sacrifice amplifier performance by having poor dc bias design.

Transistor circuits require dc bias that provides the desired quiescent point. Further, it should hold the operation stable over a range of temperatures. Resistive circuits used at lower frequencies can be employed in the RF/MW range as well. However, sometimes these circuits may not work satisfactorily at higher frequencies.

For example, a resistance in parallel with a bypass capacitor is frequently used at the emitter to provide stable operation at lower frequencies. This circuit may not work at microwave frequencies because it can produce oscillation. Further, the resistance in an amplifier circuit can degrade the noise figure. Active bias networks provide certain advantages over the resistive circuits.

It is a good practice to use some form of feedback in the bias circuit to minimize the dc voltage and current variations of the device.

When power loss is critical or for large temperature changes, active biasing is employed. Active biasing offers high level of dc stability.

Since a common-emitter configuration gives a 180° phase change between collector and base at dc, any resistive connection between those terminals provides negative feedback.

Designers prefer to dc ground the RF transistor directly instead of adding a bias resistor into the common lead, even though the emitter feedback (or source feedback for FETs) is a very effective technique for dc bias stability. This is because at RF/MW frequencies, bypassing a resistor in the common ground is not easy due to component parasitic and resonances, particularly in broadband applications. At low frequencies, a bypassed emitter resistor is an important contributor to quiescent-point stability. At MW frequencies, the bypass capacitor can produce oscillations by making the input port unstable at some frequencies.

At MW frequencies, the transistor parameters that are affected most by temperature are I_{CBO}, h_{FE}, $and\ V_{BE}$. The conventional reverse current, I_{CBO}, (at low frequencies) doubles every $10^{\circ}C$ rise in temperature (2.57).

$$I_{CBO,T_2} = I_{CBO,T_1} 2^{(T_2 - T_1)/10} \qquad 2.57$$

$where\ I_{CBO,T_2}\ and\ I_{CBO,T_1}$ are the values of I_{CBO} at temperatures $T_2\ and\ T_1$ respectively. T_1 ($\sim 300K$ is usually the temperature at which the manufacturer measures I_{CBO}.

A microwave transistor has a more complicated reverse current flow composed of two components: conventional I_{CBO} and the surface current I_s.

I_s flows across the top of the silicon lattice, and increases at a rate much higher than conventional I_{CBO}.

The base-emitter voltage has a negative temperature coefficient (2.58).

$$\frac{\Delta V_{BE}}{\Delta T} \approx -2e3\, V/°\mathrm{C} \qquad 2.58$$

The dc value of h_{FE} is found to increase linearly with temperature at the rate of $0.5\%/°\mathrm{C}$.

In order to compute the change in I_C as a function of temperature in a dc bias network, I_C can be expressed as:

$$I_C = f(I_{CBO}, h_{FE}, V_{BE}) \qquad 2.59$$

And,

$$\Delta I_C = \left(\frac{\Delta I_C}{\Delta I_{CBO}}\right)\Bigg|_{\substack{\Delta h_{FE}=0 \\ \Delta V_{BE}=0}} \Delta I_{CBO} + \left(\frac{\Delta I_C}{\Delta h_{FE}}\right)\Bigg|_{\substack{\Delta I_{CBO}=0 \\ \Delta V_{BE}=0}} \Delta h_{FE} + \left(\frac{\Delta I_C}{\Delta V_{BE}}\right)_{\substack{\Delta I_{CBO}=0 \\ \Delta h_{FE}=0}} \Delta V_{BE} \qquad 2.60$$

Defining stability factors;

$$S_{I_{CBO}} = \left(\frac{\Delta I_C}{\Delta I_{CBO}}\right)\Bigg|_{\substack{\Delta h_{FE}=0 \\ \Delta V_{BE}=0}} \qquad 2.61$$

$$S_{h_{FE}} = \left(\frac{\Delta I_C}{\Delta h_{FE}}\right)\Bigg|_{\substack{\Delta I_{CBO}=0 \\ \Delta V_{BE}=0}} \qquad 2.62$$

$$S_{V_{BE}} = \left(\frac{\Delta I_C}{\Delta V_{BE}}\right)_{\substack{\Delta I_{CBO}=0 \\ \Delta h_{FE}=0}} \qquad 2.63$$

Thus, rewriting(2.60),

$$\Delta I_C = S_{I_{CBO}}\Delta I_{CBO} + S_{h_{FE}}\Delta h_{FE} + S_{V_{BE}}\Delta V_{BE} \qquad 2.64$$

For a given dc bias network, stability factors can be calculated and (2.64) used to predict variations of I_C with temperature. In a design procedure, the maximum variation of I_C in a temperature range can be selected and (2.64) be used to find the required stability factors. In turn, the stability factors, together with the Q-point location will fix the values of the resistors in the network.

Common dc bias networks that can be used at MW frequencies are given in Fig 2.1.1 Any increase in the quiescent collector current, (I_C) causes a larger voltage drop across R_1 which reduces V_{CE} which in turn reduces I_B and (I_C).

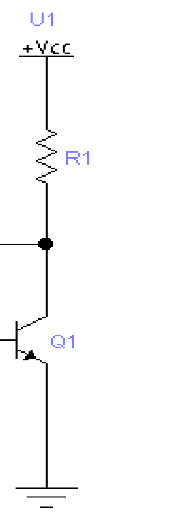

Fig 2. 1. 1 (a): Voltage Feedback Biasing

Thermal runaway is prevented when half-power principle is employed, that is, $V_{CE} = 0.5V_{CC}$.

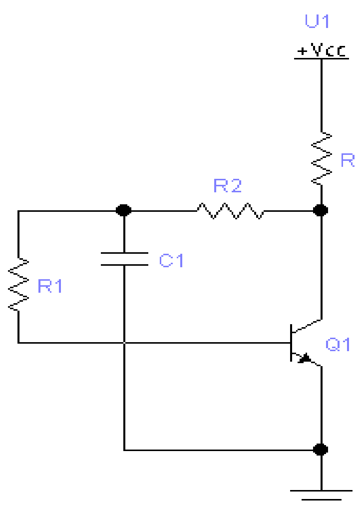

Fig 2. 1. 1 (b): Split Voltage Feedback Bias

The feedback resistor is split into two resistors, $\boldsymbol{R_1}$ and $\boldsymbol{R_2}$, with a capacitor $\boldsymbol{C_1}$ connected between its junction and chassis ground. The purpose of $\boldsymbol{C_1}$ is to prevent any output RF signal from travelling back to the input circuit. $\boldsymbol{R_2}$ is used to prevent short-circuiting the collector output signal via $\boldsymbol{C_1}$, and $\boldsymbol{R_1}$ is to prevent short-circuiting the base signal through $\boldsymbol{C_1}$.

Any increase in collector current ΔI_C results in a decrease in V_{CE}, V_{BB} and I_B which in turn counteracts any further increase in I_C. This circuit produces lower resistance values, and therefore is, more compatible with thick or thin film resistors. For good stability factors designers usually assume that;

$$I_{BB} \gg I_B \text{ and } V_{BB} \approx 0.1V_{CC}$$

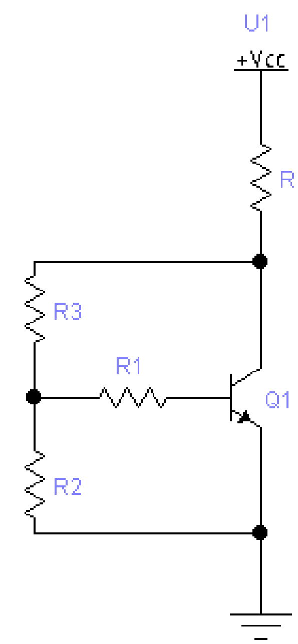

Fig 2.1.1 (c): Voltage Feedback and Constant Current Bias

2.2 BROADBAND AMPLIFIER DESIGN

Although there are no set rules to consider, an amplifier is generally considered to be narrowband when its bandwidth is less than 20% of the center frequency [1]. The design of broadband amplifiers introduces difficulties which require careful considerations. Some of these difficulties are:

- The variations of $|S_{21}|$ and $|S_{12}|$ with frequency. Typically, $|S_{21}|$ decreases with frequency at the rate $6dB/octave$ and $|S_{12}|$ increases with frequency at the same rate. The variation of $|S_{21}S_{12}|$ with frequency is important since the stability of the circuit depends on this quantity.
- The scattering parameters S_{11} and S_{22} are also frequency dependent and their variations are important over a broad range of frequencies.

- There is degradation in the noise figure and VSWR in some frequency range of the broadband amplifier.

There are several techniques used in the design of broadband amplifiers; use of compensated matching networks, use of negative feedback, resistive matching, balanced amplifiers, and traveling wave amplifiers (distributed amplifiers).

1. The technique of compensated matching networks involves mismatching the input and output networks to compensate for the variation with frequency of $|S_{21}|$. The matching networks are designed to give the best input and output VSWR. However, because of the broad bandwidth, the VSWR is optimum around certain frequencies, and a balanced amplifier design may be required. The design of compensated matching networks can be done analytically with the help of Smith Charts. However, the use of computer is usually required for the complex analytical procedures involved. Proper analytical procedures produce results that can be simulated by CAD methods. The matching networks can also be designed using network synthesis techniques. The design of compensated matching networks to obtain gain flatness results in impedance mismatching that can significantly degrade the input and output VSWR.
2. Negative feedback can be used in broadband amplifiers to provide a flat gain response and to reduce output and input VSWR. It also controls the amplifier performance due to variations in S-parameters from transistor to transistor. As the bandwidth requirements of the amplifier approach a decade of frequency, gain compensation based on matching networks only is very difficult, and negative feedback techniques are used. A disadvantage of negative feedback is that it degrades the noise figure and reduces the maximum power gain available from a transistor.
3. The resistive-matching networks are independent of frequency and hence can be used to design broadband amplifiers. The upper limit is determined from the frequencies when the resistances cease to work due to associated parasitic elements [4]. However, the noise figure of such amplifiers may be unacceptable.

CHAPTER 3: DESIGN METHODOLOGY

3.1 Introduction

A Small-Signal Amplifier maintains small-signal operation, linearity and steady state in the design frequency range.

When operating at the design frequency, transmission line theory comes into the picture. The high frequency and short wavelength of microwave energy make for difficulties in analysis and design of microwave components and systems. Matching of the input and output of the transistor must be considered and designed around. A typical block diagram of a single-stage RF amplifier is shown below.

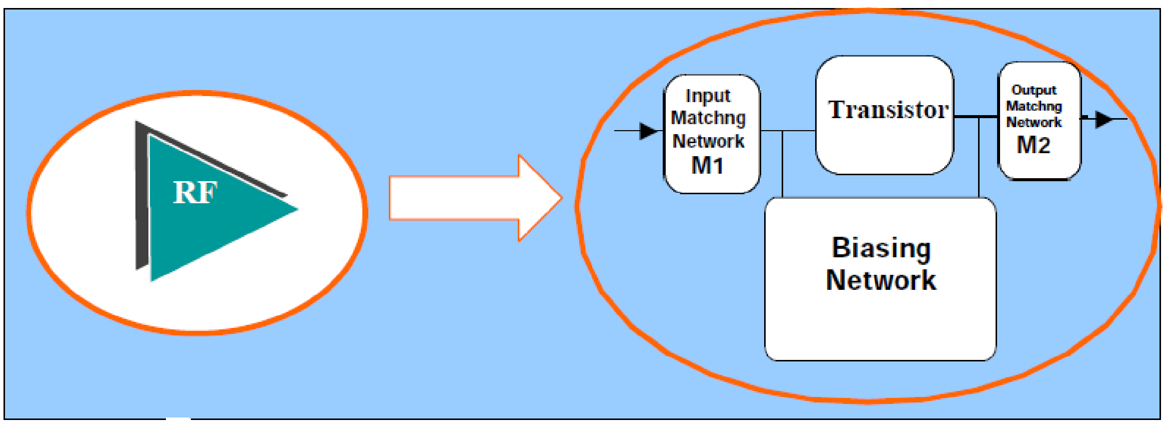

Figure 3.1 General Microwave Amplifier Topology

This was the basic topology that was adhered to through the design procedure. The basic design flow for this topology is as follows:

- Choose Microwave Transistor based on design specifications
- Design a DC Biasing circuit for desired operation.
- Design the Input and Output Matching Circuits based on the desired type of amplifier: Low-Noise Amp, High-Gain Amp, or High-Power Amp, et cetera.

Because this design was that of a small signal device, there is a more specific design flow to that can be summarized as in fig 3.2

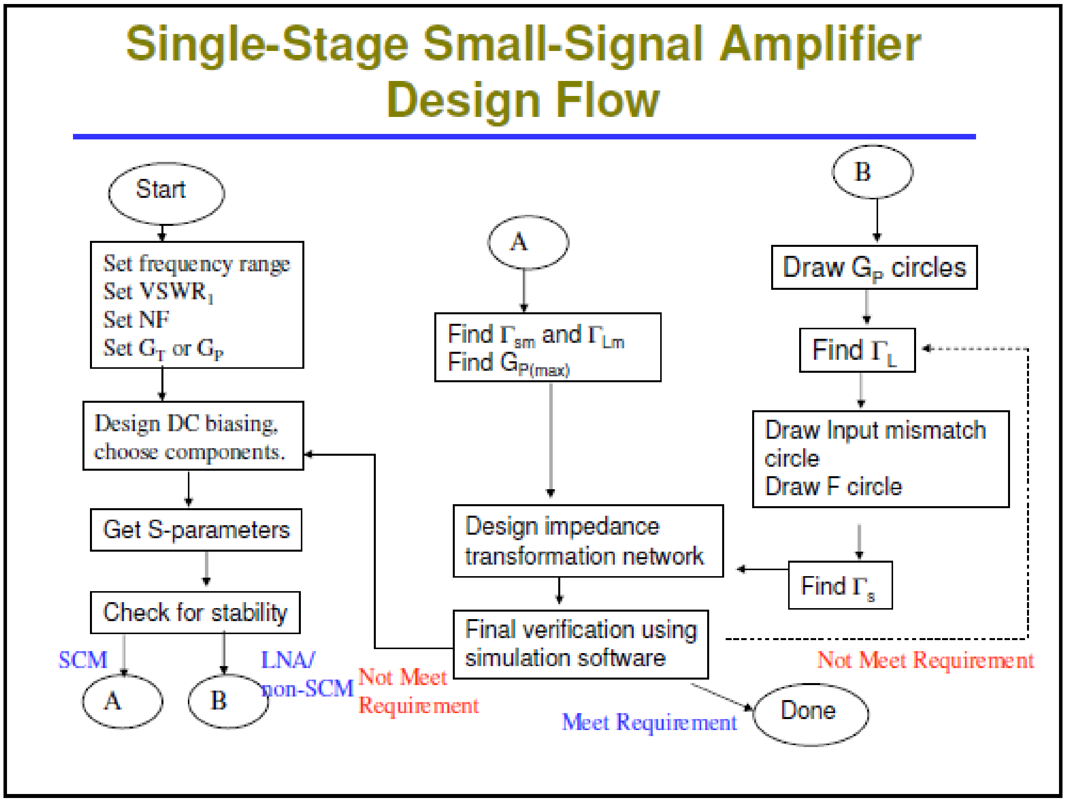

Fig 3.2

3.2 Project Specifications

The overall target specifications of the amplifier design were as follows;

- Operation Frequency @2.4GHz.
- Input impedance 300Ω.
- Bandwidth 0.48GHz

The transistor selected for this project was BFP420, a wideband Low Noise BJT with $f_T = 23GHz$.

3.3 Design Environment: Microwave Office

It was obvious from the start that the amplifier would need to be designed in the software environment if it was actually to be built. There are several software packages in the industry that are used for the design and simulation of RF circuits. The one chosen for this design was

Applied Wave Research's Microwave Office. Microwave Office is one of the top three industry standard RF design and simulation packages which made it very attractive. Learning the use and capabilities of the software through the design process turned out to be very time consuming but the experience gained with the software would no doubt be invaluable in an RF career.

3.4 D.C. bias network

Active biasing was chosen for this design.

The two circuits of fig 3.3 are convenient for creating the transfer characteristics of BFP 420, and also for finding the voltage of the base-emitter junction for a specified base current.

In fig 3.3 (d), the desired quiescent point was at $V_{CE} = 3.25$ and $I_C = 4.88$ mA. The corresponding base current (shown in mA at the right side of the plot) was slightly less than 0.05 mA. This value of I_B yielded V_{BE} of about 0.8154V as indicated in fig 3.3 (c).

The bias point selected was appropriate for low-noise and low-power microwave applications [2].

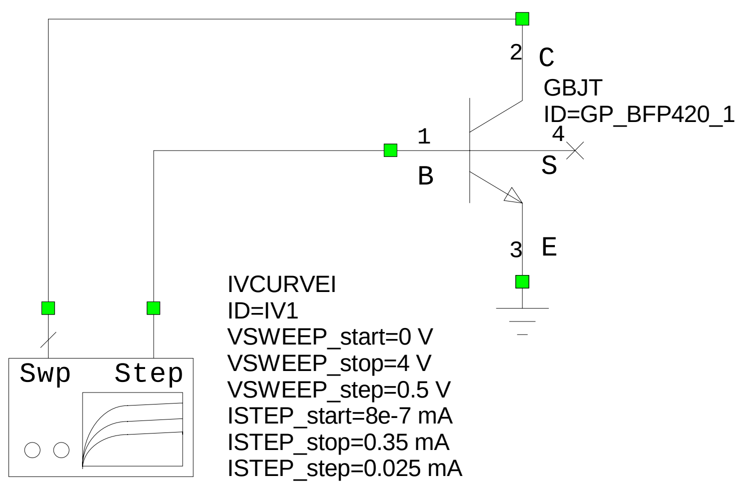

Fig 3.3 (a). Microwave Office Curvetracer set up for obtaining IV CURVES

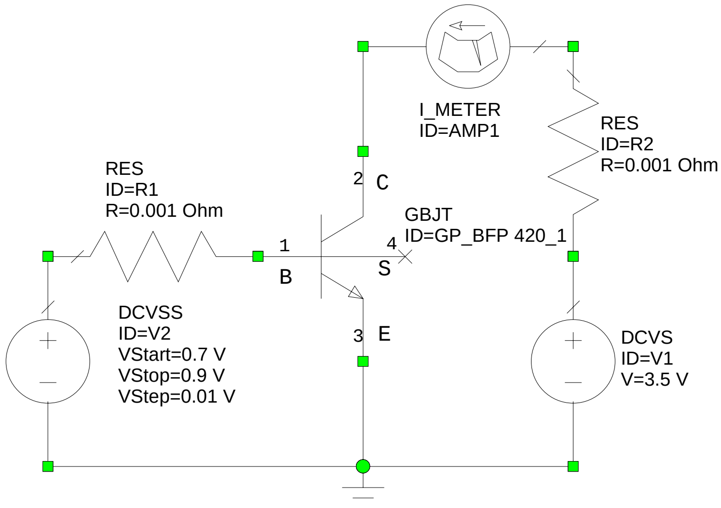

Fig 3.3(b). Circuit set up for determining I_B vs V_{BE} Curve

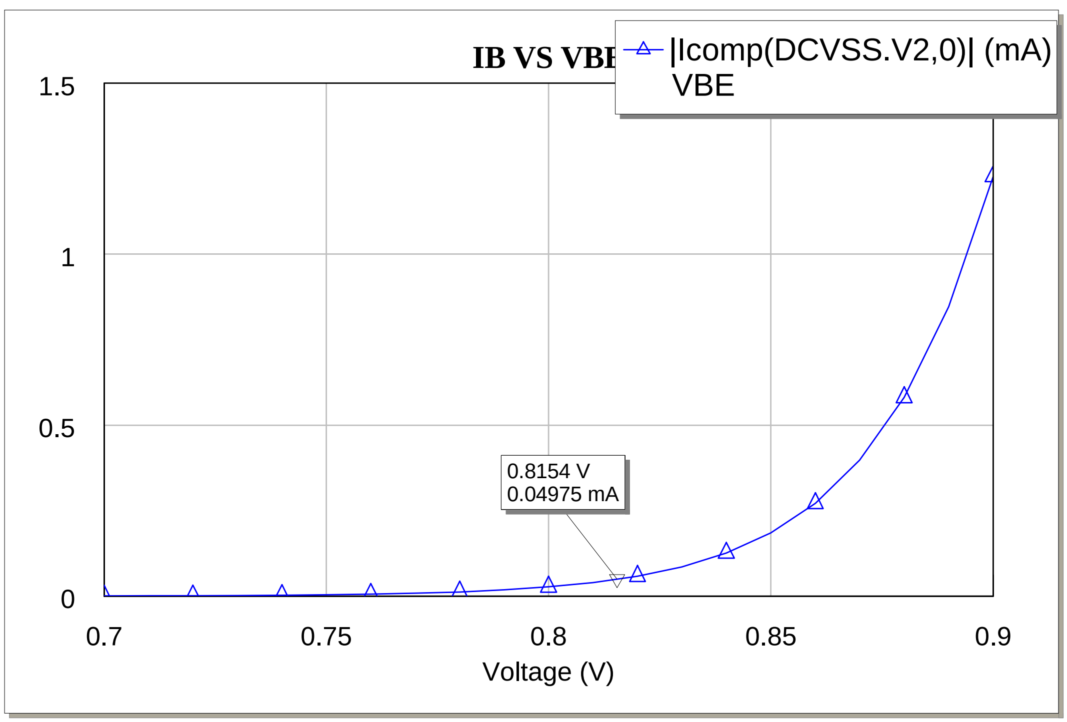

Fig 3.3(c) . Plotting $I_B$vs V_{BE}at V_{CE}= 2V helps to determine the exact V_{BE} for I_B=0.3mA

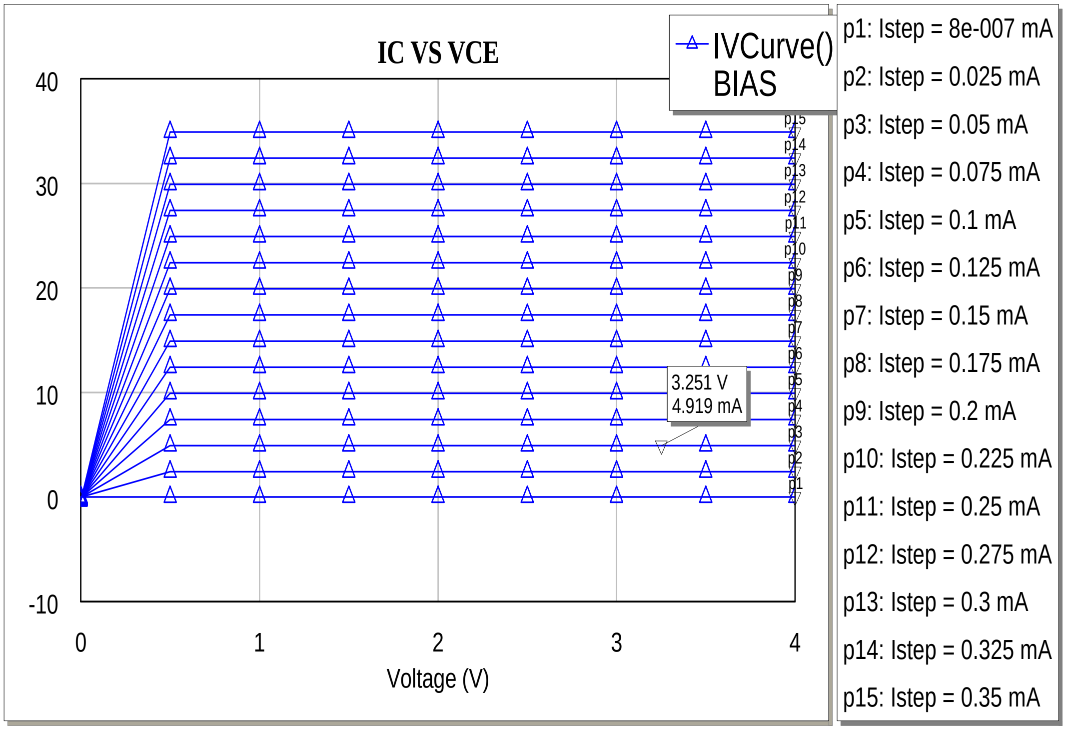

Fig 3.3(d). dc transfer characteristics of BFP420 obtained from simulation of fig. (b)

The active bias configuration is given in fig 3.4 with the bias currents and voltages annotated. A pnp BJT was used to stabilize the operating point of the BFP420. The bypass capacitors $C_1,\ C_2, and\ C_3$ were typically $0.001\mu F$ disk capacitors. The radio frequency (RFC) chokes are typically made of two or three turns of No. 36 enameled wire on 0.1-in air core. The network operated as follows: if the current through L_1 increased, the current through R_3 increased. This caused the emitter-base voltage of the pnp transistor to decrease, and result in a decrease in its emitter current. The decrease in the pnp emitter current reduced the collector and the base currents of the MW transistor, an act which in turn produced the desired stability.

The silicon diode compensated for the diode temperature dependency of the base-collector junction of B1.

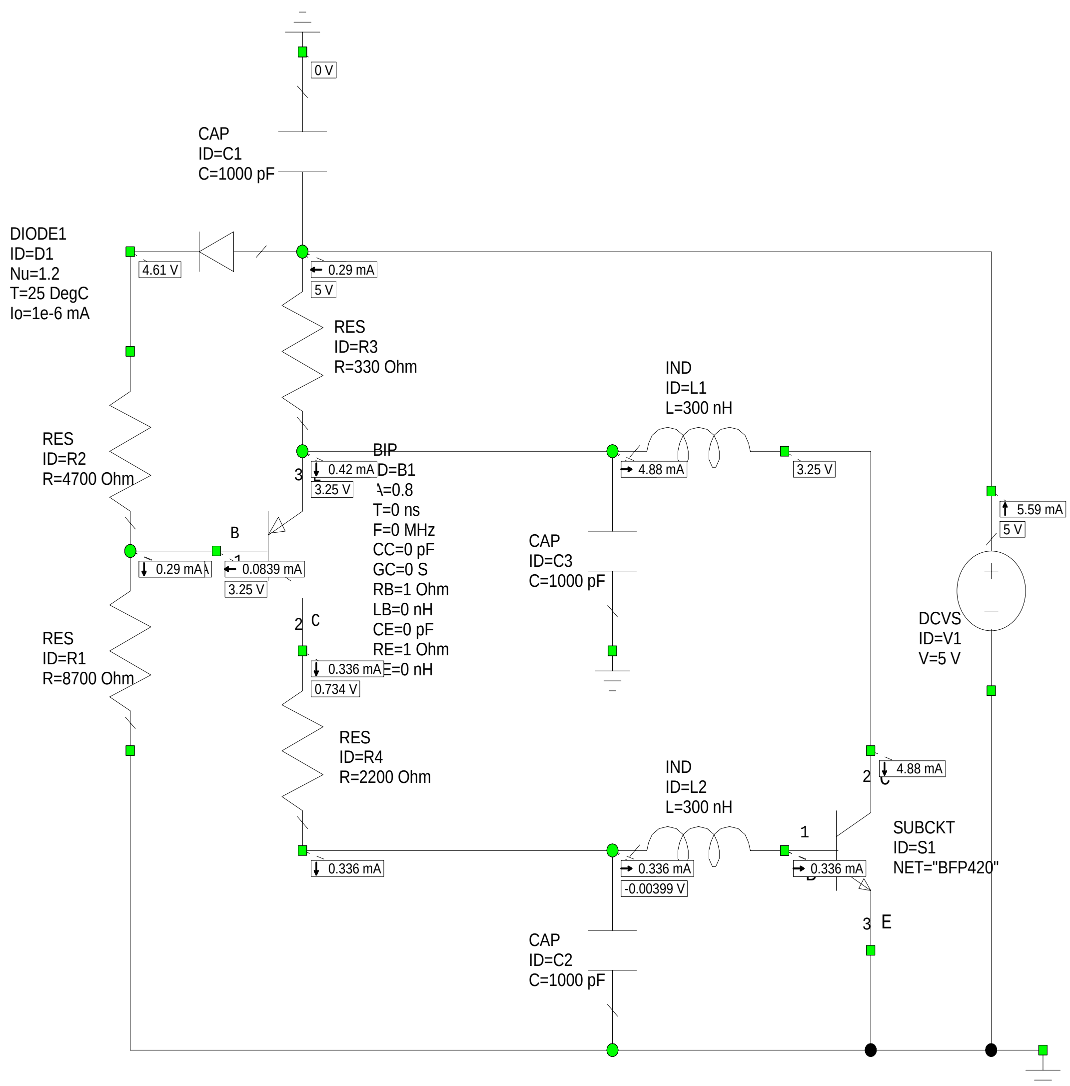

Fig 3.4 Active bias circuit for BFP420 showing the bias currents and voltages

3.5 Stability Analysis

The S parameters for BFP420 were obtained for $V_{CE} = 3.5V \; and \; I_C = 30mA$ (see Appendix 1, for the S parameter file) at the minimum upper cut off frequency of the design, that is, at $f_2 = 2.64GHz$

Using equations 2.27 and 2.28, the stability parameters Δ and K and were calculated and found to be $0.3178\angle 55.3°$ and 1.0783, respectively. Since, $|\Delta| < 1$ and the stability factor $K > 1$, the transistor was unconditionally stable.

Calculations of equations (2.38) and (2.39) yielded 1.066 (>1) and 1.089 (>1), respectively, thus, proving that the amplifier was absolutely stable for all passive source and load terminations.

3.6 Impedance Matching

In microwave amplifier design, source and load stability circles are usually drawn to determine the stable regions of the Smith Chart where impedance matching circuits could be designed with the correct input and output reflection coefficients. This ensures proper matching for stability, as explained in section 2.1.4. When designing broadband microwave amplifier, constant gain circles can be used to selectively increase or decrease the basic transducer gain between $f_1 \; and \; f_2$ until a good match is obtained. The source and load terminations $\Gamma_S \; and \; \Gamma_L$ are found to lie on specified gain circles, and this offers a variety of circuit options and a good chance at gain flattening and stability. The disadvantage of this method of broadbanding is poor impedance matching, resulting into large VSWRs, since the signals that are not transmitted are reflected.

In this design, since the transistor satisfied the conditions for absolute stability, conjugate matching was used to design the first matching circuit and then broadbanding methods were used to obtain a flat gain in the desired bandwidth.

Simultaneous conjugate matching is usually used to obtain maximum power gain from an amplifier, and this happens when;

$$\Gamma_{IN} = \Gamma_S{}^* \qquad 3.6.1$$

And,

$$\Gamma_{OUT} = \Gamma_L{}^* \qquad 3.6.2$$

From (2.25) and (2.26) it is obvious that,

$$\Gamma_S^* = S_{11} + \frac{S_{21}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad 3.6.3$$

$$\Gamma_{OUT}^* = S_{22} + \frac{S_{21}S_{12}\Gamma_S}{1 - S_{11}\Gamma_S} \qquad 3.6.4$$

Fig 3.5 explains simultaneous conjugate matching.

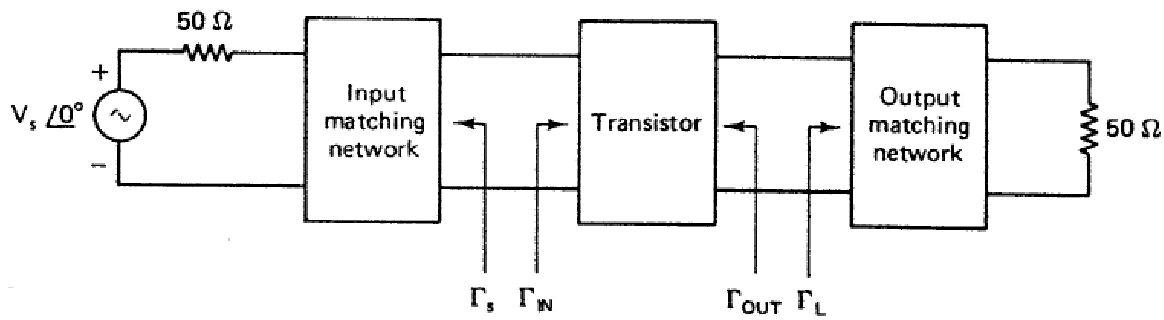

Fig 3.5 Simultaneous Conjugate Matching

Solving (3.5.3) and (3.5.4) simultaneously gives the values of Γ_S (Γ_{MS}) and Γ_L (Γ_{ML}) required for simultaneous conjugate matching.

$$\Gamma_{MS} = \frac{B_1 \pm \sqrt{{B_1}^2 - 4|C_1|^2}}{2C_1} \qquad 3.6.5$$

$$\Gamma_{ML} = \frac{B_2 \pm \sqrt{{B_2}^2 - 4|C_2|^2}}{2C_2} \qquad 3.6.6$$

Where,

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \qquad 3.6.7$$

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \qquad 3.6.8$$

$$C_1 = S_{11} - \Delta {S_{22}}^* \qquad 3.6.9$$

And,

$$C_2 = S_{22} - \Delta {S_{11}}^* \qquad 3.6.10$$

If $\left|{B_1}/{2C_1}\right| > 1$ and $B_1 > 0$ in (3.6.5) then the solution with the minus sign produces $|\Gamma_{MS}| < 1$ and the solution with the plus sign produces $|\Gamma_{MS}| > 1$. If $\left|{B_1}/{2C_1}\right| > 1$ and $B_1 < 0$ in (3.6.5) then the solution with the plus sign produces $|\Gamma_{MS}| < 1$ and the solution with the minus sign produces $|\Gamma_{MS}| > 1$. Similar considerations apply to (3.6.7).

The condition that $K > 1$ is only a necessary condition for unconditional stability. Therefore, a simultaneous conjugate match having unconditional stability is possible if $K > 1$ and $|\Delta| < 1$. Since $|\Delta| < 1$ implies$B_1 > 0$ and $B_2 > 0$ the minus signs must be used in (3.6.5) and (3.5.6) when calculating simultaneous conjugate match for an unconditionally stable two-port network.

With the BFP420 S Parameters the following values were obtained for the quantities defined in (3.5.5) to (3.5.10).

$$B_1 = 0.9936 \qquad 3.6.11$$

$$B_2 = 0.8044 \qquad 3.6.12$$

$$C_2 = 0.2495\angle - 36.13° \qquad 3.6.13$$

$$C_1 = 0.4686\angle 160.7° \qquad 3.6.14$$

$$\Gamma_{MS} = 0.5475\angle - 160.7° \qquad 3.6.15$$

$$\Gamma_{ML} = 0.3477\angle 36.13 \qquad 3.6.16$$

A Smith Chart was then used to find the input and output lossless matching networks using (3.5.15) and (3.5.16).

The maximum transducer gain obtained from conjugate matching is given by,

$$G_{T,MAX} = \left|\frac{S_{21}}{S_{12}}\right| \left(K - \sqrt{(K^2 - 1)}\right) \qquad 3.6.17$$

$$= 17.06dB$$

The circuit obtained from the design method followed so far is presented in fig 3.6

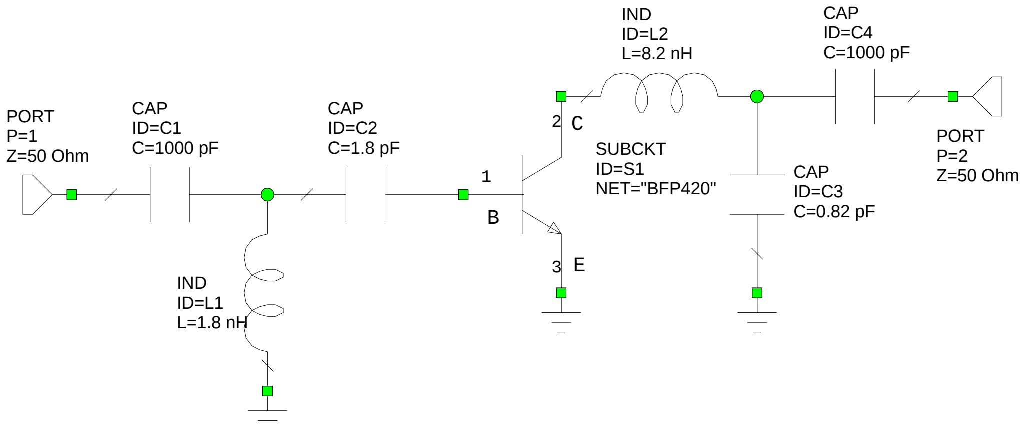

Fig 3.6 Microwave Amplifier

After the impedance matching design, the amplifier was found to have a narrow bandwidth. Broadband design and gain flattening techniques were then employed. Negative feedback as a technique used in wideband amplifier design [see section 2.2] was designed and incorporated in the network in order to achieve gain flattening in the desired frequency range. Appropriate feedback resistorR_F was determined from the transistorS_{21} and the desired input impedance [3].

$$R_F = Z_{in}[1 + |S_{21}|] \qquad 3.6.18$$

For $Z_{in} = 50\Omega$, $R_F = 327\Omega$ (The nearest standard value is 330Ω).

To compensate for the gain roll off, an inductance in series with the feedback resistance was necessary. This inductance is determined from the upper cut off frequency by the equation,

$$L_F = \frac{R_F}{2\pi f_{3dB}} \qquad 3.6.19$$

At $f_{3dB} = 2.64GHz$, $L_F \approx 22nH$.

The amplifier in fig 3.7 was obtained after incorporating both feedback components, R_F and L_F.

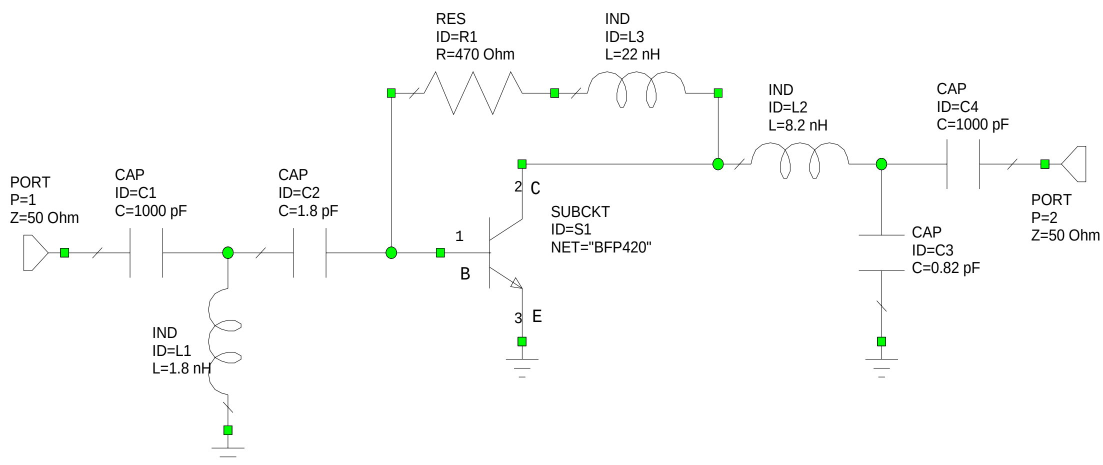

Fig 3.7 Microwave Amplifier with Feedback

Reactive matching network was also employed at the output in order to improve the gain at high frequency, and to reduce the feedback effect to some degree. Fig 3.8 gives the amplifier circuit with the reactive matching network.

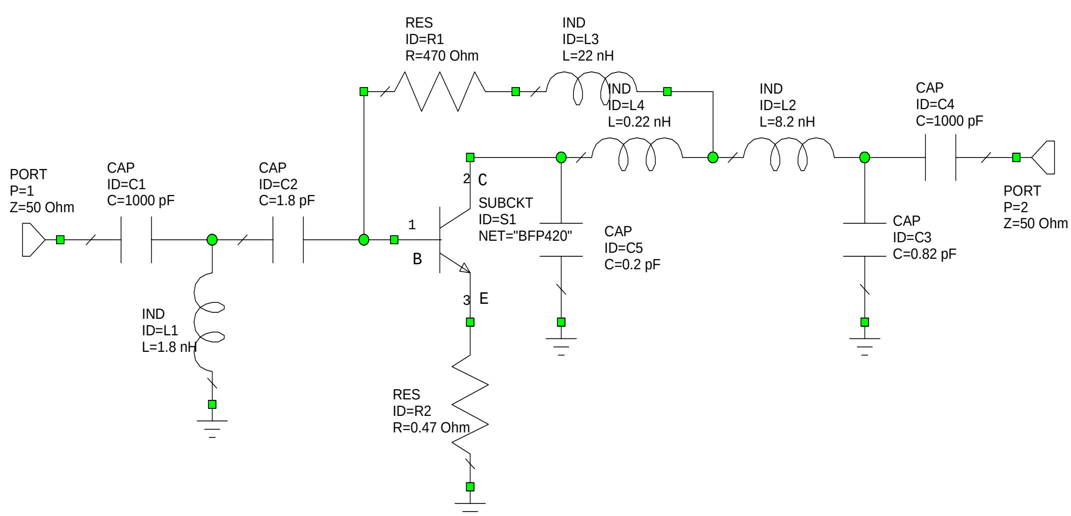

Fig 3.8 Microwave Amplifier with Feedback and Reactive Broadbanding elements.

In order to achieve the input impedance of 300Ω the relationship between transistor transconductance g_m and the dc short circuit-current gain β was utilized.

Finally, the lumped elements of fig 3.8 were converted to distributed elements. There are several methods for converting lumped elements to distributed elements, as explained in ref. 6. The methods employed here involved single-stub matching using reflection coefficient at the input and direct application of the formulas involving capacitance and inductance susceptance [6] at the output. Conversion of the $8.2nH$ and the $22nH$ series inductances were impossible through the direct application of the formulas due to the restrictions imposed on series inductances [6]. In transforming series inductors the resistances in series with the inductance is also transformed due to transmission-line effects. This is not a desirable phenomenon. It is to be noted that this conversion was only impossible at the operation frequency 2.4, but could have been carried out at much higher frequencies. Actually, $8.2nH$ is the maximum realizable inductance at $6GHz$.

The equation (3.5.20) transforms a given inductance into equivalent electrical length.

$$\beta l = \frac{\omega L/Z_0}{\sqrt{1-(\omega L/Z_0)^2}} \qquad 3.6.20$$

The series $0.22nH$-inductor was realizable at the frequency of operation. An attempt to convert the output matching circuit to distributed circuit using Richard's Transformation and Kuroda's identities failed because these two concepts are based on commensurate lowpass filters.

Transforming the dc blocking capacitances to lumped parameters was impossible analytically but could have been done in Microwave Office using Artwork Cells. This was not done, though, due to time limit; the Software was new to the designer, and learning it was extremely time-consuming. The dc blocking capacitances were also too large to be replaced with Chipcaps available in Microwave Office.

The transmission line lengths were eventually converted to microstrip lines, lengths in mils. The microstrip substrate used was Duroid 5880 with $\varepsilon = 2.2, h = 31$mil.

The transformed Microwave amplifiers are given in figs (3.9) and (3.10).

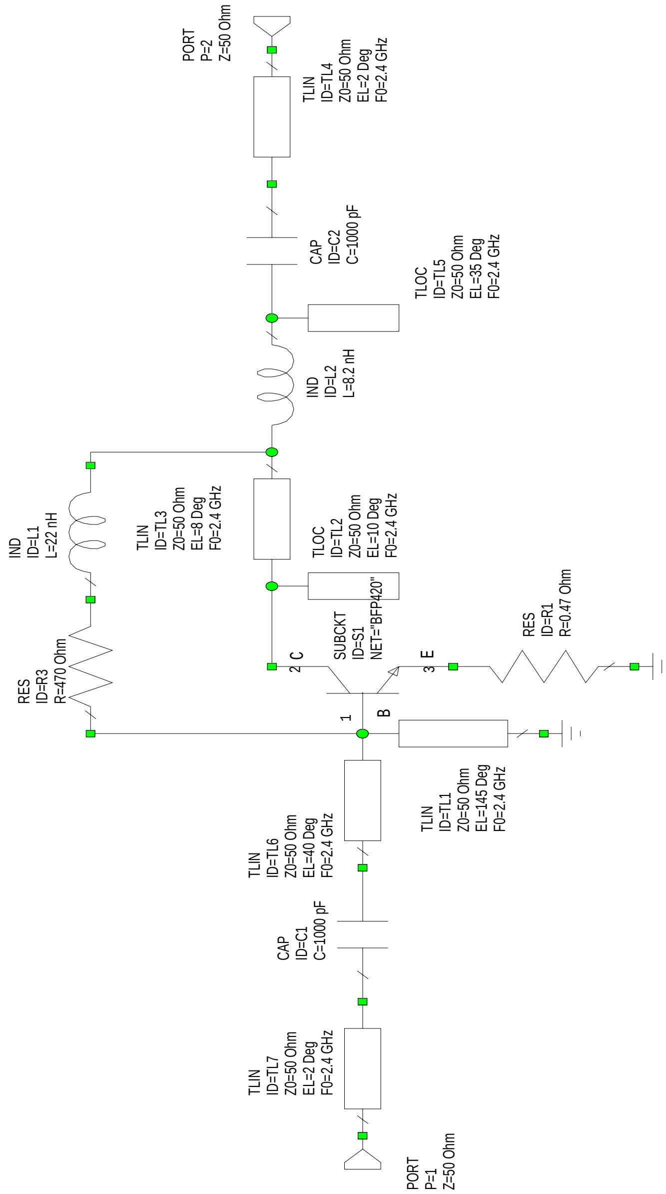

Fig 3.9 Distributed Network in electrical lengths.

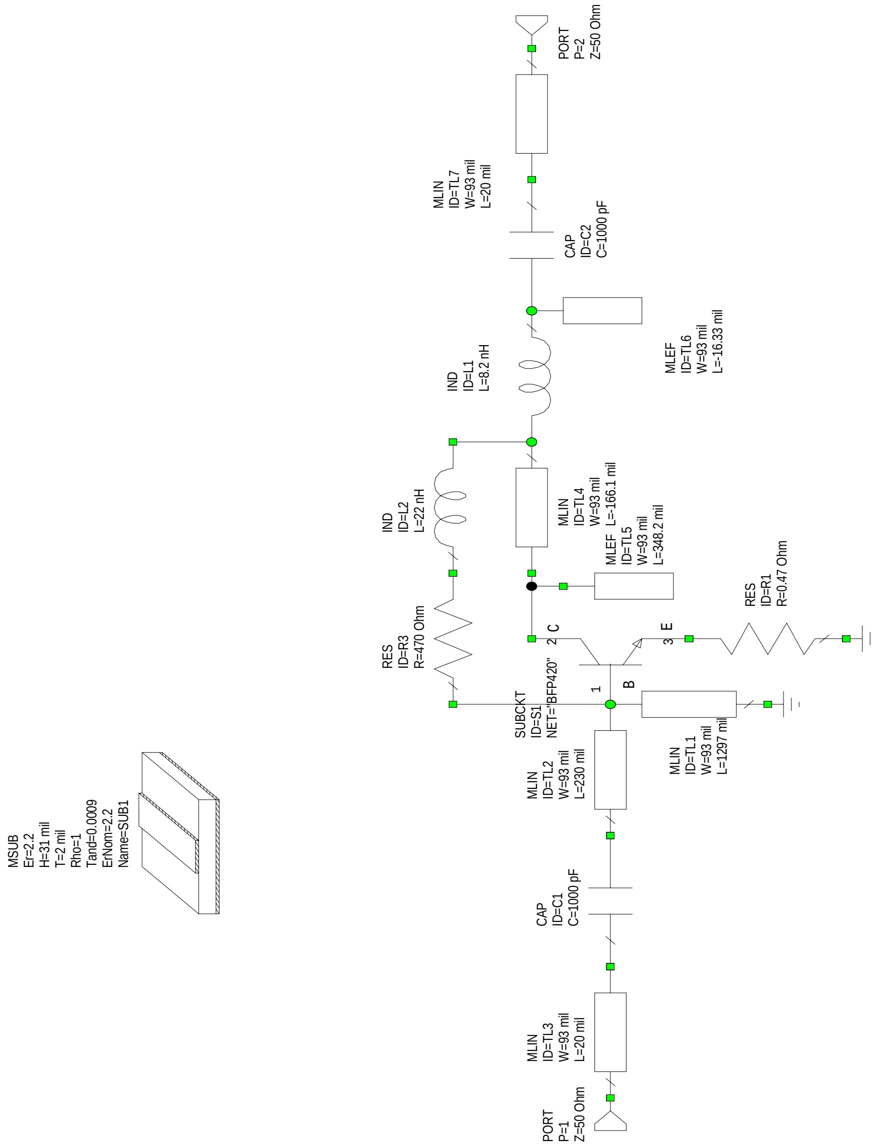

Fig 3.10 Amplifier Schematic with Microstrip transmission lines

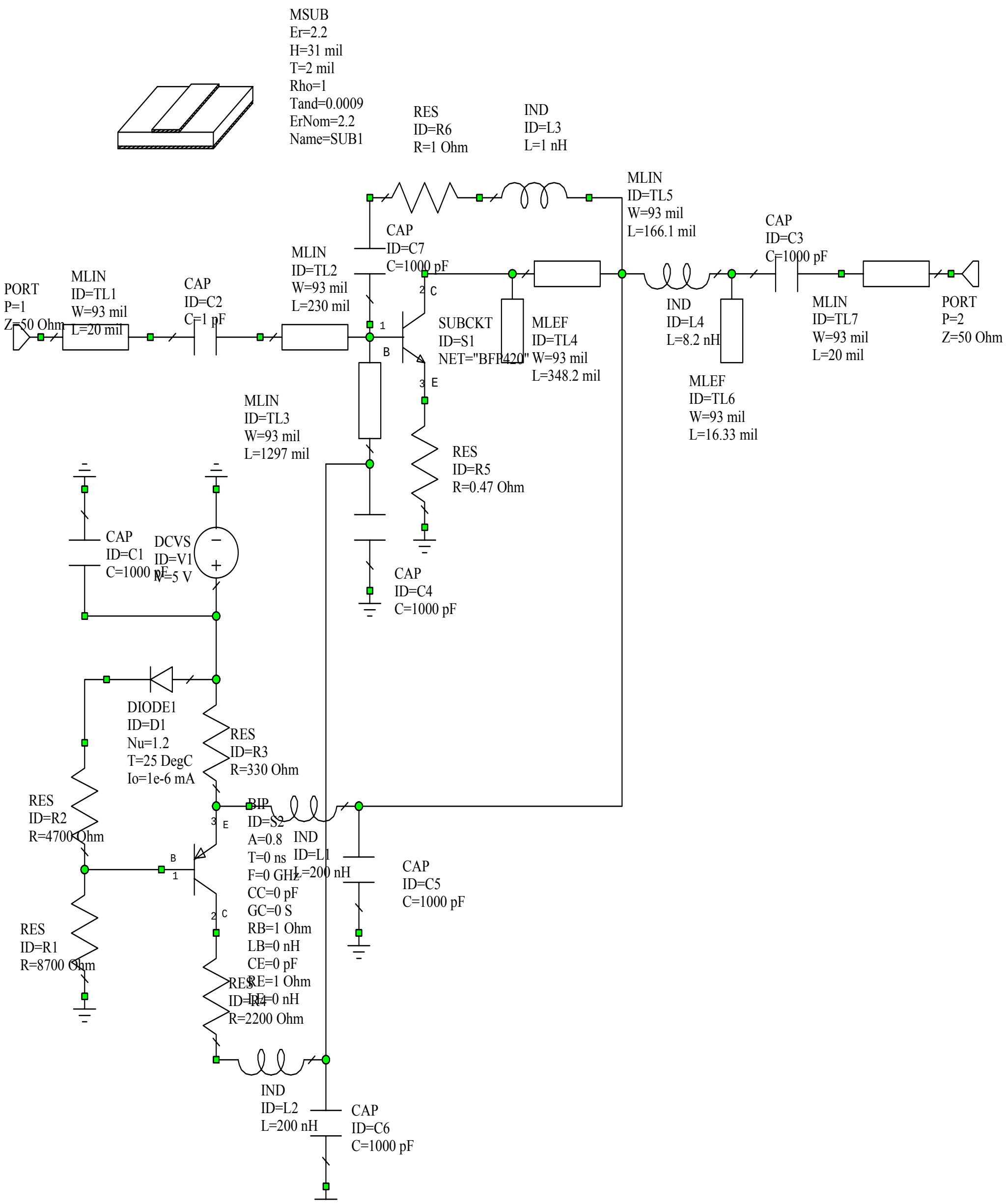

Fig 3.11 Complete amplifier schematic with dc biasing. Active biasing was used. Capacitances C1 to C7 are high Q dc blocks while the inductances L1 and L2 are RF Chokes.

CHAPTER 4: RESULTS AND ANALYSIS

4.1 Input impedance

The desired input impedance of 300Ω was achieved as follows;

The transistor transconductance is given by (4.1)

$$g_m = \frac{1 + |S_{21}|}{Z_0} \qquad 4.1$$

With $Z_0 = 50$ and $|S_{21}| = 5.539$, $g_m = 0.1308S$

Now, $r_{be} = \frac{\beta}{g_m} = \frac{100}{0.1308} = 764.2\Omega$

From the circuit the emitter feedback resistance $R_E = 0.47\Omega$,

Input impedance seen at the base of the transistor before the input matching network was added is then given by (4.2)

$$Z_{in} = r_{be} + \beta R_E \qquad 4.2$$

$$Z_{in} = 811\Omega$$

It is clear from this result that shunting the transistor base with an impedance of 476Ω (standard 470Ω) produces an input impedance of 300Ω.

Transistor parasitic inductances should have been considered in the above analysis but they could be limited practically to the bonding inductance, and this makes them have practically no effect on input impedance until well above very high frequencies (vhf).

The feedback resistance, besides ensuring gain-flattening at the centre frequency, also acts to limit input impedance to the desired 300Ω.

This analysis assumes that $\beta_{dc} = \beta_{ac}$.

4.2 Broadband Design

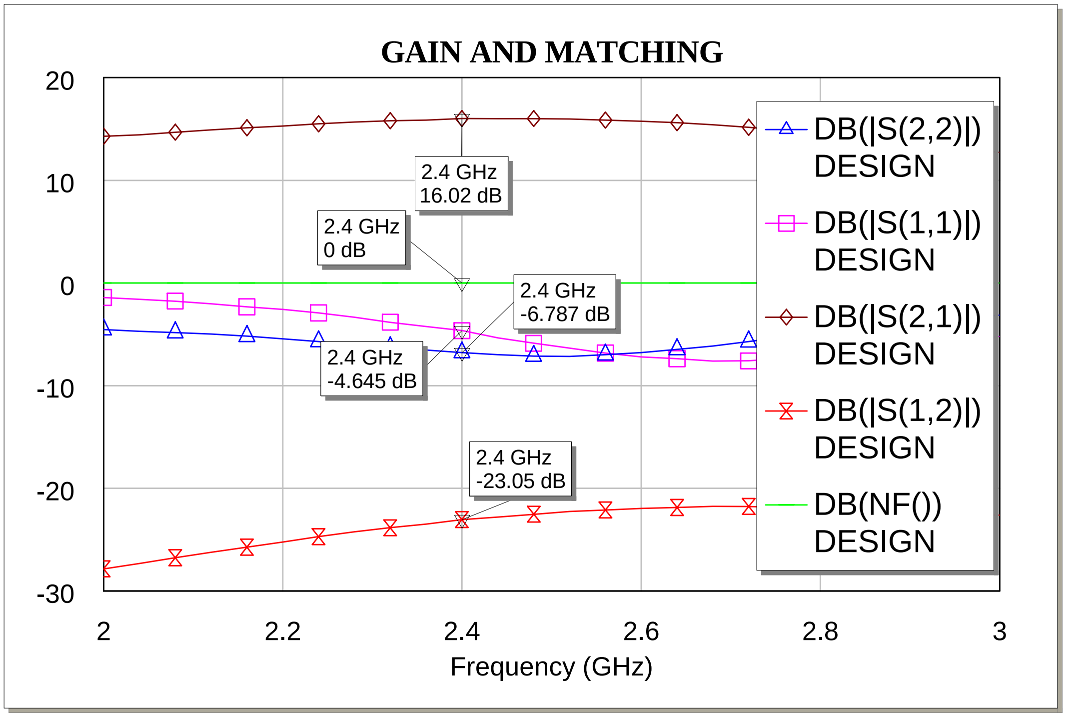

Fig 4.1 Frequency-dependent gain, matching, and noise performance of the circuit in fig 3.6

In fig 4.1, the 16.02db gain is a good approximate from the calculated G_{TMAX} of $17.06dB$ in 3.6.17. The frequency-dependent gain $|S_{21}|^2$ is relatively flat and spans a wide bandwidth, covering the Project Specification Range of $0.48GHz$. The input reflection coefficient and VSWR are about 0.6 and 4, respectively. The output reflection coefficient and VSWR are about 0.46 and 2.7, respectively. The output matching seems good but the plots of Z_{11} and Z_{22} fig 4.2 reveal that the output loading is capacitive since the $0.82pF$ capacitance shunts the output. The input impedance of this circuit is about $r_{be} = \frac{\beta}{g_m} \approx 764\Omega$. This is not the desired Z_{in}. Z_{11} is to a good approximation of 50Ω.

The noise figure is 0dB since there are no resistances in the circuit.

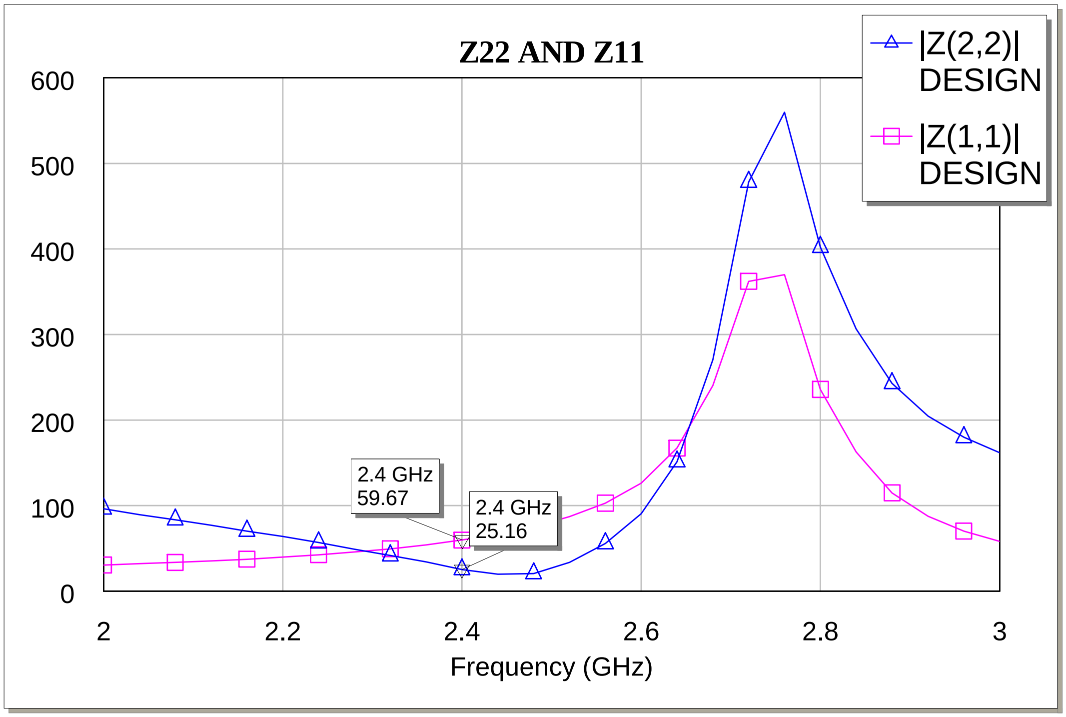

Fig 4.2 Poor output matching of the circuit in fig 3.6

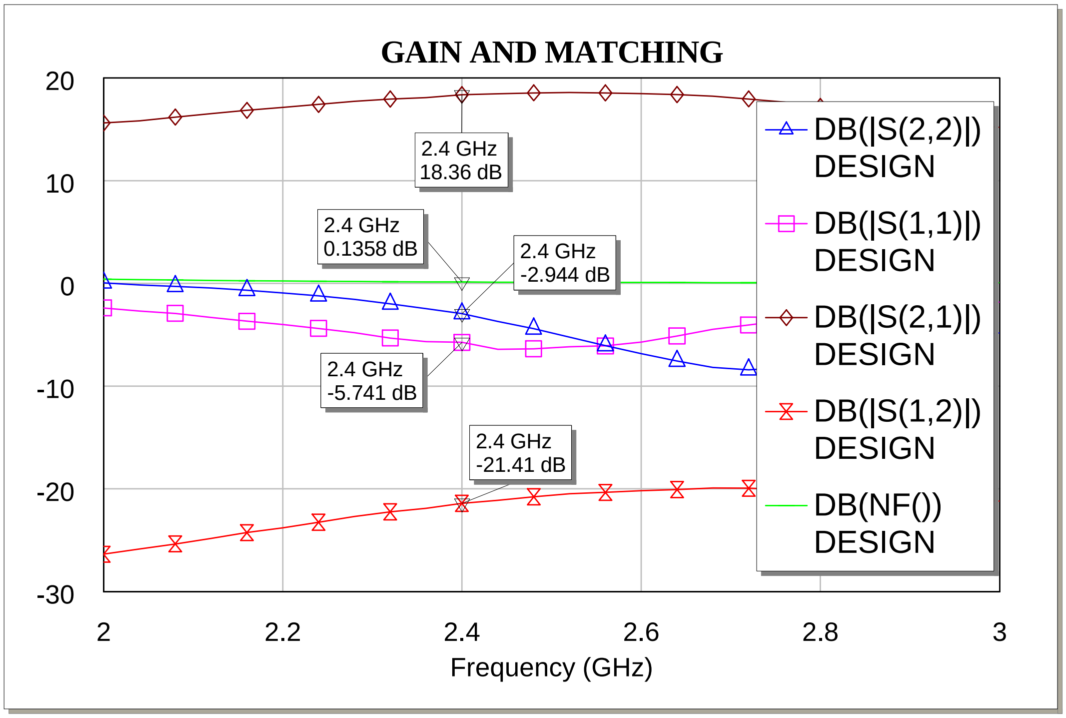

Fig 4.3 Frequency-dependent gain, matching and noise performance of the circuit in fig 3.7

Fig 4.3 reveals the plots of the frequency-dependent gain, matching and noise performance of the circuit in fig 3.7. This circuit in fig 3.7 has the feedback resistance of 470Ω that shunts the input r_{be} to 300Ω. The input VSWR has improved to about 3 while the output VSWR has worsened to approximately 0.7. From fig 4.4 the amplifier Z_{22} has decreased severely possibly due to the shunting effect of R_F at the output, which results poorer matching than before. There is gain improvement due to R_F loading at the input.

Increase in bandwidth is apparent from the frequency-dependent gain curve. This is due to the feedback resistance R_F . Shunt feedback network is one of the ways of broadening the amplifier bandwidth. The noise figure has degraded to 0.1358 due to R_F influence.

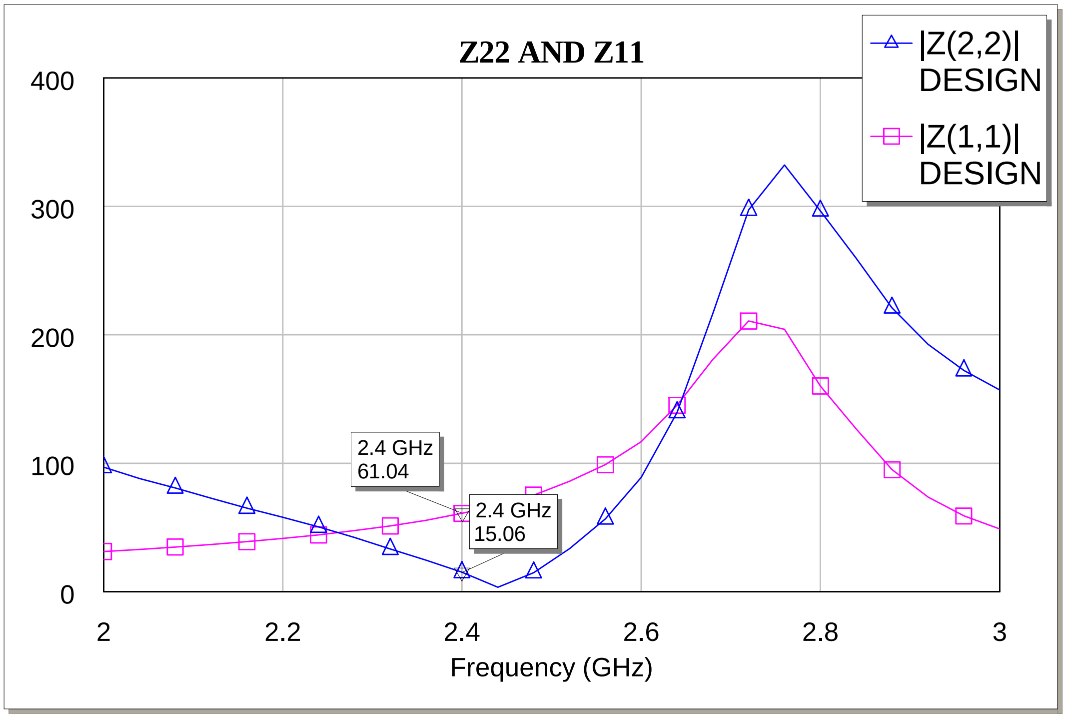

Fig 4.4 Z_{22} of the circuit in fig 3.7 has reduced severely, possibly due to the shunting effect of R_F at the output which results in poorer matching. It remains capacitive.

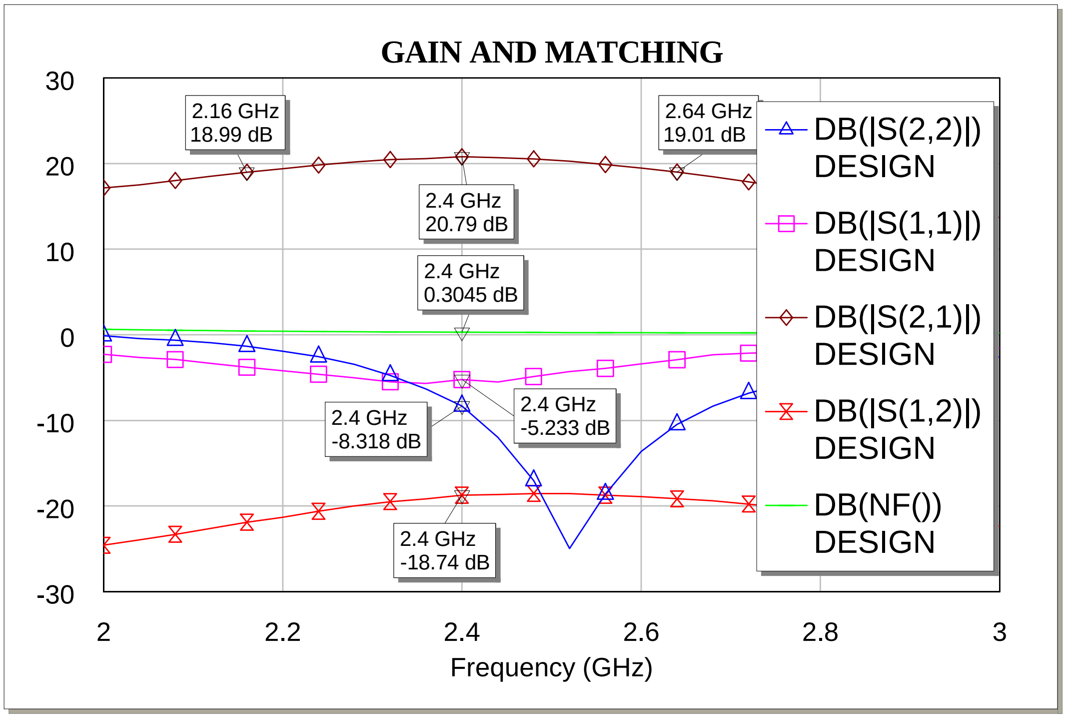

Fig 4.5 Frequency-dependent gain, matching and noise performance of the circuit in fig 3.8

In fig 4.5 the frequency range of interest is more defined with f_1 and f_2 being equal at approximately 19dB. This effect was achieved through the reactive matching network containing shunt capacitance, $C5$ and series inductance $L4$. In a broadband microwave amplifier design, if the goal is to cover the frequency range between f_1 and f_2 then f_2 should be 1 to 2dB below G_{TMAX} [1]. This fact is well portrayed in fig 4.5.

The reactive matching network has improved the output matching and VSWR. It was also used for gain-shaping at the upper cut off frequency. The emitter feedback resistance degraded the noise figure further to 0.3045dB and worsened the input mismatch. It also helped in gain-shaping at the lower cut off frequency.

Fig 4.6 shows the overall effect of the emitter resistance and the output reactive matching network on impedance.

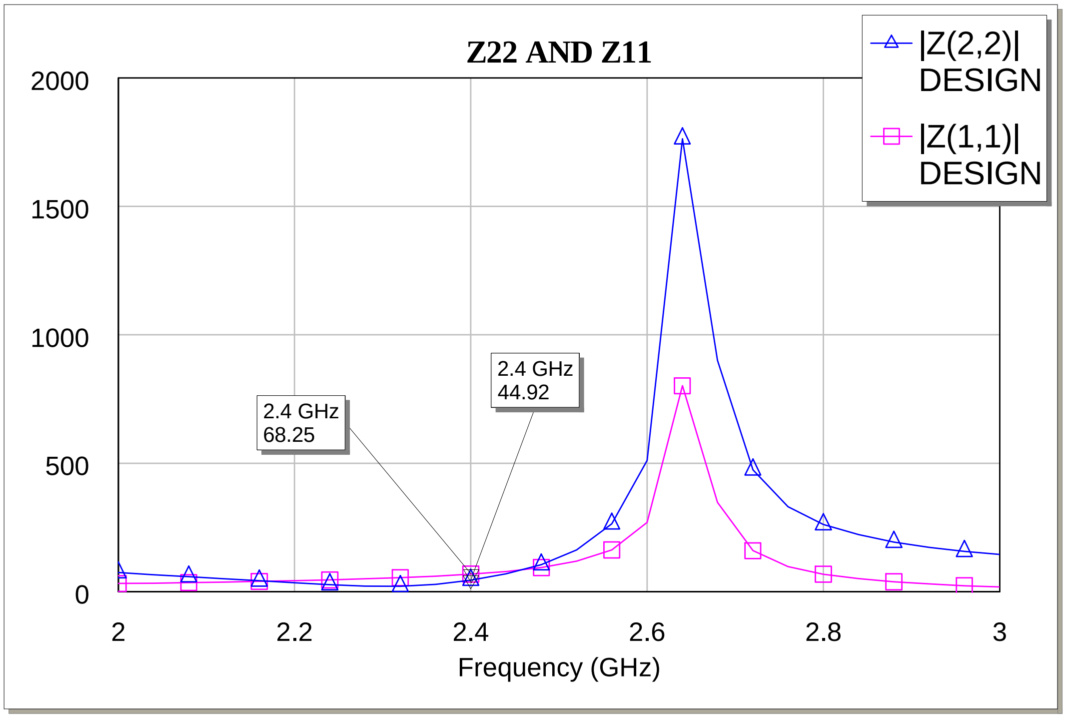

Fig 4.6 Overall effects of emitter resistance and the output reactive matching network on impedance. The 0.47Ω emitter resistance increased input mismatch while the output reactive matching network improved the output matching.

Fig 4.7 shows the frequency-dependent gain, matching and noise performance of the distributed microwave amplifier in fig 3.9.

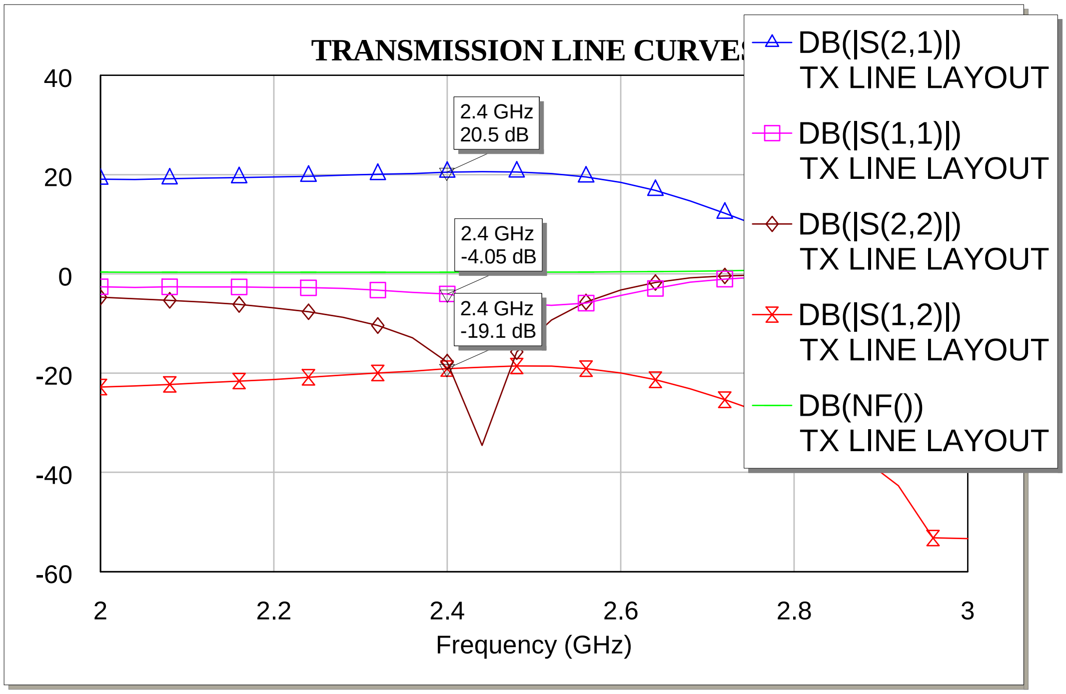

Fig 4.7 Frequency-dependent gain, matching and noise performance of the circuit in fig 3.9

In fig 4.7 gain-flattening was achieved at low frequencies. The output VSWR has a low value of about 0.11. The input VSWR is approximately 0.63. These values are tolerable at the frequency of range interest. Noise figure has remained fairly constant.

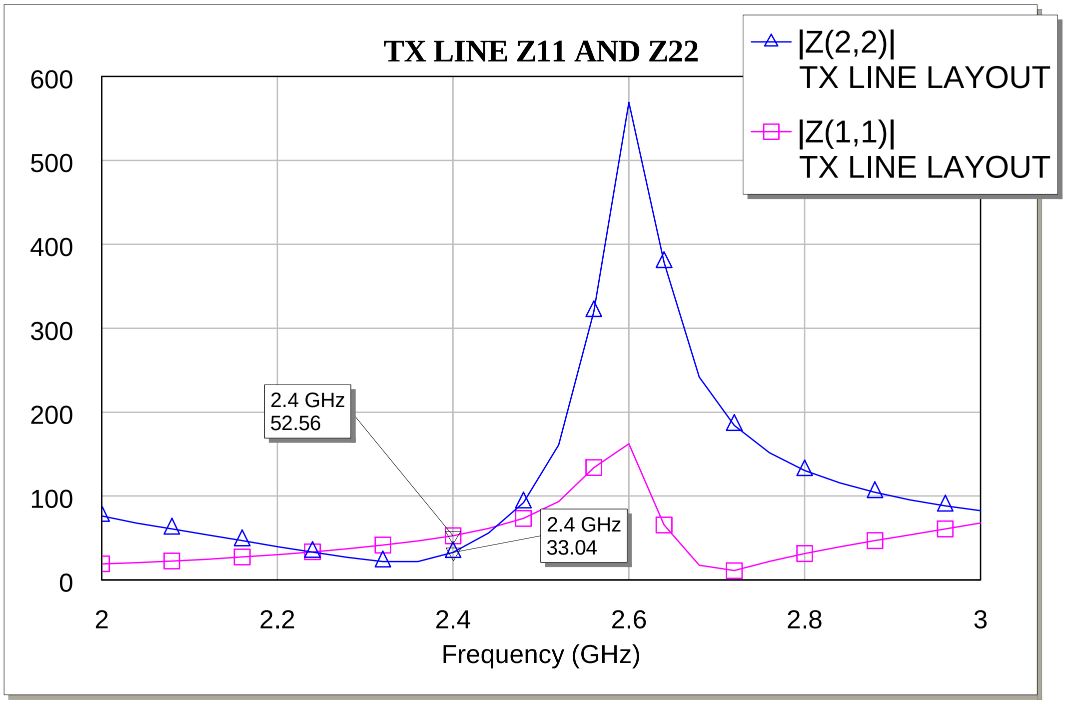

Fig 4.8 Output mismatch for the distributed amplifier. The difference in these impedances and those in fig 4.6 is due to the errors of computational approximations when converting lumped to distributed elements.

Fig 4.9 shows the frequency-dependent gain, matching and noise performance of the microwave amplifier in fig 3.10 using microstrip transmission lines.

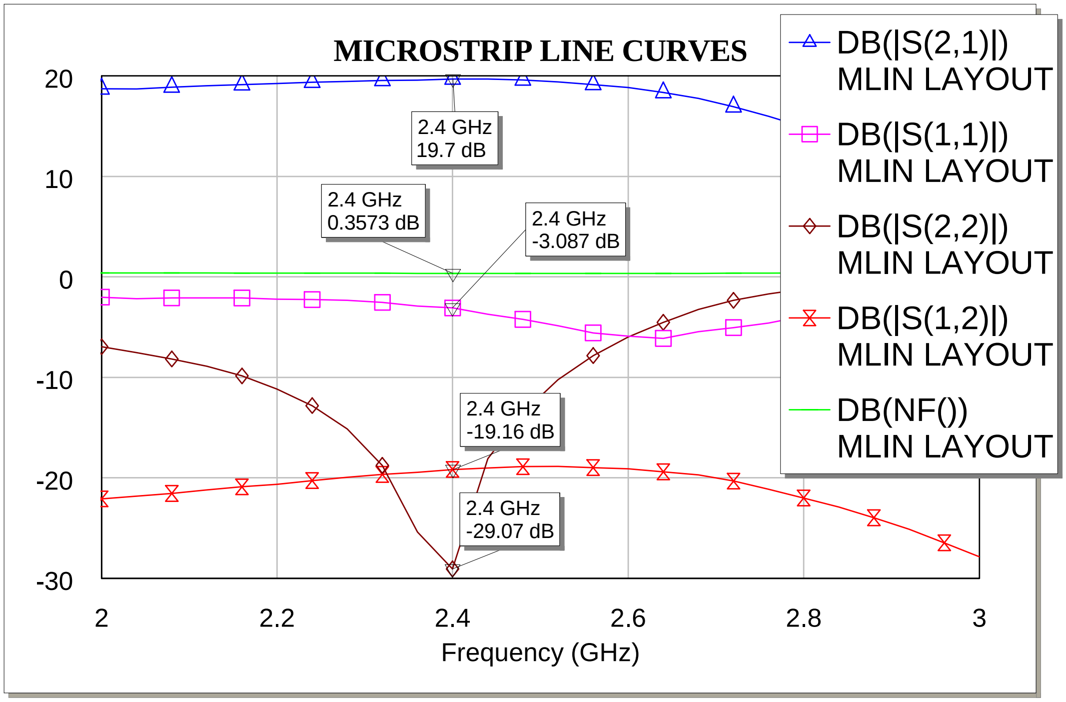

Fig 4.9 Frequency-dependent gain, matching and noise performance of the circuit in fig 3.10

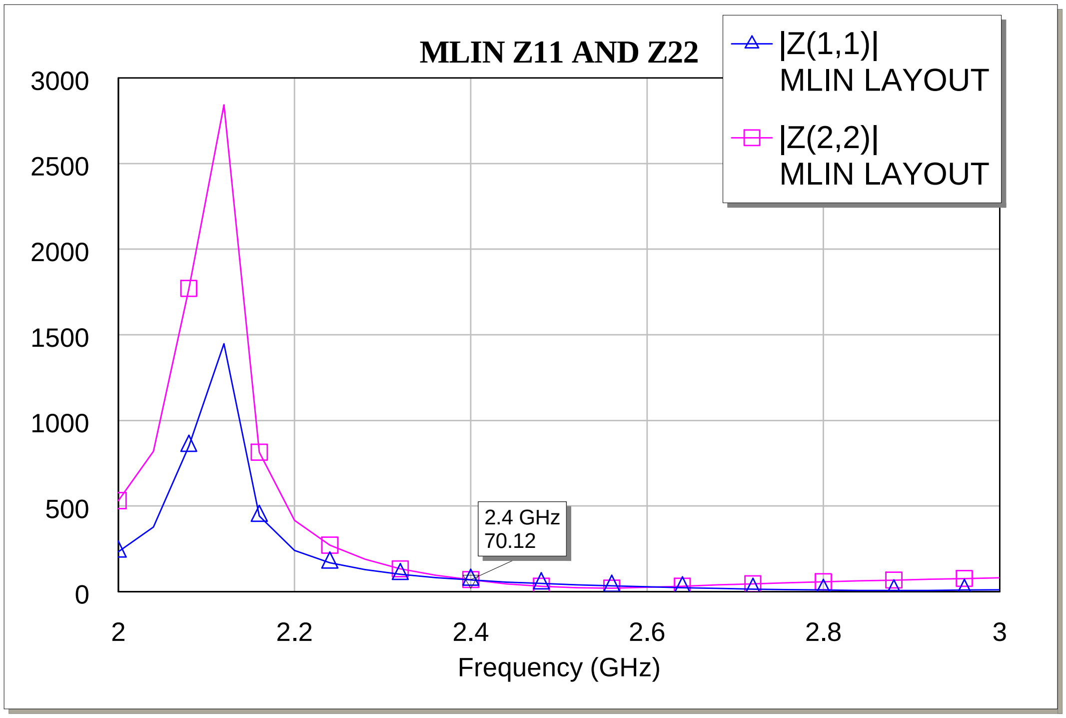

Fig 4.10 Final simulation results for input and output matching. $Z_{11} = Z_{22}$

CHAPTER 5: CONCLUSION S AND FUTURE WORK

5.1 Conclusion

Designing an amplifier to operate at microwave frequencies is a tiresome and time-consuming task because of the complexities involved in stabilization, dc biasing, and design of impedance matching networks and final layout of the circuit. One of the difficulties experienced lay in the transformation of lumped networks to distributed networks, especially where a lumped network contained series inductors. Several reference materials in the field of Microwave Engineering strongly advise on the use of Smith Chart in designing short- or open-circuit stubs to replace lumped networks. But what about in situations where open- or short-circuit stubs can no longer accurately represent a lumped network? Or what if a reactive lossless network (lowpass or highpass) is incorporated into the amplifier for broadbanding purposes as was done in this project?

Much time was spent by the designer in researching these areas that cast severe uncertainty on the project. Even renowned authors of the ilk of D. M. Pozar, Collin R.E., Vendelin, Besser and Gonzalez, to name a few, simply brushed away the concept of replacing series inductances with transmission lines as complex and practically unrealizable. But Pieter [6] attacked the subject and provided approximations to replace a series inductor with a series transmission line. But at a specified frequency, there was a limit to the values of inductances that could be approximated to transmission lines. For instance, at 2.4 GHz, the 8.2 nH-inductor at the output matching of the amplifier designed in this project was too large to be replaced with a transmission line.

Another problem was in feeding active-bias dc signal into RF Circuit when the RF Circuit had only open-circuited stubs. Les Besser [2] finally solved this problem but only partially.
In the final microstrip layout, the schematic elements like the capacitances and inductances are not shown. These elements are not in the layout database of Microwave Office, and Artwork Cells could have been created for them, but the designer ran out of the time due to complicated personal issues.

The objectives of the project, especially in the wideband amplifier design were met, and the experience gained is worth a lifetime to the designer.

5.2 Future Work

Even though the objectives of this project were met, it was done in a software environment, and effort should be put in the future to build it in a lab environment. Research could also be carried out in determining how RF Circuits with only open-circuited stubs should be actively biased. Another area that needs attention is the transformation of series inductances to series transmission lines. This project can also be done again using other methods of broadbanding such as resistive matching, balanced amplifiers and distributed amplifiers.

APPENDIX 1

BFP420 S PARAMETER DATAFILE

```
!  Infineon Technologies AG
! BFP420 , Si-NPN RF-Transistor in SOT 343
!  Vce=3.5 V,  Ic=30.0 mA
! Common Emitter S-Parameters:              21. August 2003
#  GHz  S  MA  R  50
! f           S11            S21            S12            S22
! GHz      Mag     Ang      Mag   Ang      Mag   Ang      Mag   Ang
2.000000  0.381879  176.610   7.314802   71.969  0.057116   60.667
0.291791  -59.290
 2.037500  0.377096  176.766   7.140318   71.715  0.058182   59.965
0.289513  -59.517
 2.075000  0.384885  175.860   7.027829   70.747  0.059163   60.011
0.286455  -60.104
 2.112500  0.378726  175.007   6.925057   70.080  0.060317   60.200
0.282909  -60.318
 2.150000  0.378102  173.200   6.789431   69.568  0.061103   60.066
0.280328  -60.479
 2.187500  0.382368  173.504   6.682805   68.752  0.061986   59.676
0.277890  -61.082
 2.225000  0.385068  172.346   6.554921   68.252  0.063143   59.186
0.275580  -61.325
 2.262500  0.382607  171.209   6.463558   67.639  0.064177   59.115
0.273401  -61.673
 2.300000  0.387280  169.843   6.331814   67.162  0.065169   58.756
0.269800  -62.264
 2.337500  0.380597  169.237   6.254433   66.537  0.065879   58.622
0.267417  -62.420
 2.375000  0.391433  169.063   6.126288   66.091  0.066803   58.235
0.265024  -62.624
 2.412500  0.392562  167.685   6.069014   65.177  0.068235   58.320
0.262300  -63.225
 2.450000  0.382384  167.533   5.961854   64.430  0.068831   58.272
0.261200  -63.545
 2.487500  0.390501  166.618   5.871130   64.069  0.069876   57.541
0.258597  -63.768
 2.525000  0.392700  166.069   5.767929   63.331  0.071023   57.246
0.256246  -64.099
 2.562500  0.392891  165.311   5.684379   62.652  0.071863   57.687
0.254703  -64.611
 2.600000  0.391605  164.635   5.619301   62.157  0.073084   57.382
0.251862  -64.811
 2.637500  0.396084  163.975   5.538932   61.611  0.073822   57.017
0.249570  -65.339
 2.675000  0.392244  162.116   5.467404   61.044  0.075265   56.691
0.247213  -65.283
 2.712500  0.388583  161.527   5.395897   60.572  0.076518   56.305
0.245738  -65.896
 2.750000  0.396166  161.200   5.312917   59.658  0.076968   56.208
0.244171  -66.198
 2.787500  0.391322  160.549   5.224392   59.132  0.077691   55.683
0.240606  -66.491
 2.825000  0.395937  160.541   5.178797   58.934  0.078963   55.387
0.239769  -66.932
 2.862500  0.394462  159.454   5.112419   58.308  0.079965   55.152
0.238862  -67.211
```

2.900000	0.395388	158.687	5.042406	57.701	0.081005	54.669	0.236559	-67.635
2.937500	0.396995	157.220	4.979839	57.385	0.081706	54.729	0.235023	-68.086
2.975000	0.391256	157.201	4.915117	56.407	0.082889	54.565	0.232612	-68.385
3.012500	0.398343	156.442	4.861039	55.944	0.083817	54.536	0.230720	-68.840
3.050000	0.396967	155.472	4.786260	55.621	0.085167	54.104	0.229606	-69.149
3.087500	0.393155	155.873	4.724222	54.779	0.085597	53.519	0.227959	-69.594
3.125000	0.399198	154.575	4.675677	54.409	0.086937	53.306	0.226667	-69.686
3.162500	0.400166	153.658	4.632051	53.442	0.088220	53.167	0.225634	-70.286
3.200000	0.402587	153.608	4.576778	52.961	0.089322	53.146	0.223830	-70.482
3.237500	0.403350	153.332	4.522041	52.639	0.090522	52.603	0.222348	-71.183
3.275000	0.403094	151.533	4.471821	52.063	0.091104	52.393	0.220730	-71.398
3.312500	0.401236	151.314	4.428153	51.294	0.091741	51.910	0.218587	-72.062
3.350000	0.398213	150.563	4.378623	51.102	0.092901	51.669	0.217187	-72.463
3.387500	0.401266	150.323	4.321409	50.306	0.093618	51.319	0.216382	-72.819
3.425000	0.403309	149.296	4.265532	49.908	0.095217	51.154	0.215472	-73.193
3.462500	0.405833	148.759	4.233951	49.129	0.096220	50.747	0.214258	-73.917
3.500000	0.405960	147.635	4.184242	48.520	0.096997	50.400	0.212025	-74.150
3.537500	0.404165	146.928	4.132979	48.261	0.098117	49.933	0.211188	-74.859
3.575000	0.393723	146.845	4.089454	47.148	0.098914	49.442	0.210210	-75.574
3.612500	0.397837	145.909	4.016462	46.828	0.099510	49.527	0.207412	-76.349
3.650000	0.402909	146.067	3.954053	46.346	0.100496	49.125	0.203745	-77.012
3.687500	0.394366	146.252	3.905630	46.054	0.101016	48.957	0.200602	-77.006
3.725000	0.399343	146.326	3.874707	45.888	0.102417	49.096	0.198721	-76.757
3.762500	0.402796	146.364	3.858441	45.374	0.103262	48.660	0.199178	-76.907
3.800000	0.405588	145.341	3.815202	44.948	0.104772	48.342	0.197176	-77.240
3.837500	0.408848	144.288	3.799782	44.105	0.105989	47.976	0.196992	-77.810
3.875000	0.410781	143.857	3.750574	43.950	0.106848	47.686	0.196284	-77.966
3.912500	0.416303	143.171	3.718667	43.558	0.107827	47.512	0.194869	-78.533
3.950000	0.414628	142.723	3.684369	42.771	0.108626	46.924	0.194045	-79.232
3.987500	0.414461	142.130	3.667516	42.361	0.109865	46.700	0.192887	-79.743

4.025000	0.416217	142.278	3.640495	41.893	0.111270	46.642	0.191075	-79.674
4.062500	0.418846	140.926	3.604968	41.271	0.111866	46.116	0.189645	-80.553
4.100000	0.418080	141.072	3.563019	40.696	0.112786	46.044	0.189429	-80.998
4.137500	0.419690	139.716	3.552358	40.429	0.114750	45.361	0.188874	-81.592
4.175000	0.421384	139.829	3.516522	39.676	0.114705	45.275	0.185730	-81.815
4.212500	0.421025	138.609	3.468038	39.367	0.116420	45.042	0.186200	-82.134
4.250000	0.420938	137.824	3.451227	38.624	0.117080	44.646	0.184319	-82.714
4.287500	0.431959	137.173	3.417767	38.090	0.118043	44.466	0.182223	-83.567
4.325000	0.426874	136.976	3.396016	37.760	0.118686	44.219	0.181202	-83.946
4.362500	0.423527	136.375	3.370511	36.790	0.120433	43.480	0.181521	-84.530
4.400000	0.426971	136.040	3.326403	36.602	0.120652	43.411	0.179019	-84.914
4.437500	0.430559	135.062	3.315910	36.186	0.122227	43.025	0.178302	-85.189
4.475000	0.429156	134.588	3.289286	35.386	0.122644	42.489	0.177320	-86.329
4.512500	0.432126	133.366	3.260016	34.859	0.124243	42.272	0.175804	-86.582
4.550000	0.434866	133.373	3.220724	34.500	0.125153	42.045	0.174299	-87.079
4.587500	0.434215	132.448	3.208251	34.283	0.125843	41.781	0.172799	-87.302
4.625000	0.431262	132.775	3.178978	33.517	0.127413	41.354	0.171701	-87.897
4.662500	0.434115	131.283	3.162050	33.132	0.128159	41.097	0.170638	-88.386
4.700000	0.438047	131.079	3.134383	32.796	0.128855	40.640	0.169590	-89.144
4.737500	0.443037	131.434	3.113074	32.090	0.129918	40.566	0.168437	-89.356
4.775000	0.436970	130.265	3.077331	31.676	0.131215	40.056	0.167969	-90.115
4.812500	0.437837	129.760	3.070535	31.146	0.131677	39.654	0.166615	-90.415
4.850000	0.440166	128.725	3.054003	30.660	0.133182	39.297	0.165317	-90.857
4.887500	0.443159	128.039	3.009982	30.522	0.134093	38.959	0.163708	-91.709
4.925000	0.439428	127.914	2.994307	29.723	0.135127	38.734	0.162420	-91.906
4.962500	0.450480	127.722	2.969569	29.370	0.136234	38.734	0.160051	-92.576
5.000000	0.447854	126.324	2.946631	29.128	0.137723	38.175	0.159511	-93.194
5.037500	0.446512	126.232	2.922664	28.243	0.137825	37.687	0.158758	-93.924
5.075000	0.445504	125.834	2.897340	27.885	0.138897	37.548	0.158418	-94.101
5.112500	0.448659	124.720	2.887779	27.172	0.139997	37.211	0.157785	-94.921

5.150000	0.452343	124.789	2.845660	26.798	0.139871	36.925	0.157012	-95.113
5.187500	0.446588	124.390	2.834391	26.554	0.141698	36.953	0.154789	-95.901
5.225000	0.451720	123.529	2.817340	25.795	0.142542	36.031	0.154408	-96.585
5.262500	0.447092	122.732	2.798691	25.322	0.143521	35.849	0.153045	-97.286
5.300000	0.448599	122.826	2.794906	24.975	0.144807	35.516	0.152140	-97.625
5.337500	0.454096	121.875	2.778000	24.350	0.146060	35.349	0.151668	-98.344
5.375000	0.451103	120.335	2.749838	23.957	0.146360	34.944	0.149152	-98.792
5.412500	0.456274	120.093	2.747726	23.554	0.148178	34.827	0.148857	-99.181
5.450000	0.447066	120.010	2.713755	22.613	0.148543	34.190	0.147908	-100.401
5.487500	0.448973	119.422	2.695732	22.421	0.149236	33.982	0.146318	-100.664
5.525000	0.452682	119.605	2.673561	22.125	0.151583	33.572	0.145702	-101.725
5.562500	0.453908	118.627	2.661426	21.510	0.151776	33.348	0.144610	-101.914
5.600000	0.453079	118.125	2.632365	20.807	0.152810	33.025	0.143556	-102.317
5.637500	0.459533	117.136	2.625380	20.728	0.153526	32.719	0.141629	-103.133
5.675000	0.456034	117.152	2.617550	20.239	0.155126	32.508	0.142006	-103.718
5.712500	0.458703	117.483	2.595424	19.702	0.155128	32.007	0.140723	-104.467
5.750000	0.460899	115.225	2.569927	18.974	0.156802	31.944	0.140140	-105.223
5.787500	0.456281	115.446	2.552457	18.801	0.157882	31.391	0.139058	-106.131
5.825000	0.461250	116.139	2.538169	18.588	0.158819	31.295	0.136940	-106.397
5.862500	0.457688	115.438	2.525414	17.671	0.160270	30.859	0.136993	-107.047
5.900000	0.460347	114.196	2.514629	17.149	0.160535	30.581	0.135552	-107.731
5.937500	0.463787	113.556	2.491738	16.458	0.161450	29.872	0.134497	-108.559
5.975000	0.470643	112.819	2.495312	16.419	0.163359	29.942	0.133427	-108.987
6.012500	0.468073	113.067	2.474895	16.081	0.164006	29.471	0.132640	-109.698
6.050000	0.466099	112.387	2.441695	15.499	0.163821	29.198	0.130591	-110.613
6.087500	0.463745	112.517	2.439775	14.929	0.165750	29.034	0.131653	-111.256
6.125000	0.464268	111.299	2.425971	14.414	0.166165	28.269	0.130309	-111.912
6.162500	0.472125	111.099	2.411627	14.181	0.167241	28.198	0.128967	-112.713
6.200000	0.467844	110.188	2.400690	13.773	0.168801	27.792	0.128642	-113.491
6.237500	0.471909	109.918	2.390102	13.269	0.170243	27.504	0.127573	-113.819

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 6.275000  0.464194  110.185   2.387476   12.842  0.170872   27.455
0.126938 -114.392
 6.312500  0.474002  109.956   2.373894   12.796  0.172482   27.297
0.126146 -114.753
 6.350000  0.467337  108.831   2.352309   12.071  0.173414   26.774
0.124420 -115.552
 6.387500  0.467320  109.010   2.326250   11.591  0.175099   26.680
0.124882 -116.459
 6.425000  0.478196  108.930   2.343459   11.307  0.175932   26.394
0.125465 -116.579
 6.462500  0.477289  107.669   2.334940   10.443  0.177266   25.621
0.124214 -118.434
 6.500000  0.480259  107.461   2.315410   10.470  0.178416   25.284
0.123793 -118.893
 6.537500  0.483847  106.968   2.311909    9.617  0.180079   24.999
0.123315 -120.280
 6.575000  0.479631  106.746   2.284239    9.272  0.180569   24.626
0.122726 -121.388
 6.612500  0.489055  106.780   2.278744    8.673  0.180821   24.175
0.121081 -122.594
 6.650000  0.482788  105.589   2.263740    8.434  0.182674   23.891
0.121008 -122.800
 6.687500  0.486068  105.510   2.243917    7.474  0.184039   23.498
0.120356 -124.112
 6.725000  0.483213  104.039   2.250078    6.955  0.183734   23.127
0.119176 -124.658
 6.762500  0.486637  103.539   2.235638    7.148  0.185720   22.876
0.118207 -125.376
 6.800000  0.489611  103.312   2.222017    6.324  0.186483   22.242
0.118004 -126.837
 6.837500  0.494012  103.124   2.224525    6.133  0.187839   22.183
0.117059 -127.226
 6.875000  0.485476  101.665   2.193551    5.528  0.188947   21.541
0.115603 -129.201
 6.912500  0.493758  102.417   2.203129    5.156  0.189313   21.266
0.114801 -129.613
 6.950000  0.490705  101.476   2.172854    4.627  0.190558   21.037
0.114238 -129.966
 6.987500  0.488954  100.994   2.178612    3.846  0.191925   20.829
0.115001 -131.576
 7.025000  0.490414  100.632   2.161848    3.697  0.192623   20.309
0.113229 -132.408
!
!                Infineon Technologies AG, Munich
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