CALIFORNIA STATE UNIVERSITY, NORTHRIDGE

THREE STAGE KU BAND LOW NOISE AMPLIFIER AT 16 GHZ

A graduate project submitted in the partial fulfillment of the

requirements for the degree of Master of Science in

Electrical Engineering

By

Soham S. Jagalpure

May 2017

The graduate project of Soham S. Jagalpure is approved by:

Mr. Benjamin F Mallard | Date

Dr. John Valdovinos | Date

Dr. Matthew Radmanesh, Chair | Date

California State University, Northridge

Acknowledgement

I must express my gratitude to my project chair and mentor Dr. Matthew Radmanesh. He was supportive during my design project planning and implementation and also made sure I worked on it at the right pace in order to meet all the deadlines. I have learnt a lot from several courses that helped this project. I now understand the design procedure to apply and utilize all the techniques that were taught in these classes in order to meet the real-world applications.

I wish to thank my other project committee members Mr. Benjamin Mallard and Dr. John Valdovinos who took interest in my project and provided me with their valuable time to complete the project. Their positive response and project interactions gave me the required moral boost to finish my project.

I am grateful to my family for always being supportive of my work and motivating me to excel.

I also wish to acknowledge all the people who stood by me and helped me to clarify my doubts to complete this project.

Table of Contents

List of Figures

List of Tables

Abstract

THREE STAGE KU BAND LOW NOISE AMPLIFIER AT 16 GHZ

By

Soham S. Jagalpure

Master of Science in Electrical Engineering

In order to design an amplifier at a certain frequency, the gain and noise figure needs to be decided on the basis of the market research. After deciding these three values, a transistor was selected such that it helped to achieve these values. The formulas used in the design process determined the number of stages required to complete the design and to measure the performance of the amplifier.

After reviewing more than fifty transistors, the transistor FHX13/14LG was decided for the design of Low Noise Amplifier in the Ku Band at 16 GHz. The design objective was to achieve 40 dB gain with a noise figure of less than 1dB at the specified frequency. After the preliminary calculations, it was concluded that there will be three stages based on the selected transistor. The finalization of each stage was a challenge because of the trade-off issues in gain and noise figure.

In order to verify the results, National Instruments Microwave Office (student version) was used for simulation. It gave the simulation values which were very close to the hand calculations. This was also verified using MATLAB along-with an RF/MW E-Book Software* software developed by Dr. Matthew Radmanesh with his published book "Advanced RF & Microwave Circuit Design" for accuracy.

Chapter 1. Introduction

Microwave Amplifiers have been useful in many applications throughout the RF and Microwave industry. There has been a huge demand for amplifiers with a high gain and a low noise figure. The technological requirement has always been influenced by the demand in the market. The amplifiers played a significant role to build the electronic circuits in such a market. The demand came to a point where companies looked for engineers with exceptional skills in these types of design and with time, the design of the circuits shifted to higher frequency ranges. Due to technological growth and high bandwidth requirements, companies started to demand design at higher frequency ranges with higher possible gain. The major market for such designs is in Aerospace and Defense industry. With time, most developed countries have invested more in their defense leading to an increase in demand for amplifier designs. [1]

Amplification is a standout amongst all the essential applications in RF and Microwave circuit design. In the past, microwave tubes and microwave diodes which were based on the negative resistance region were used. Now, with time, the use of microwave transistors (BJT or FET) has become very popular.

Definition: "An amplifier is an electronic device which is used to increase the power intensity of any signal. To achieve this objective, careful matching of the input signal needs to be implemented with better and improved amplitude by using the energy from the power supply." [1]

An amplifier is a device which gives a modulated output. This output is much greater than the applied input. A transistor (B.J.T or an F.E.T) can be used to acquire this amplification in the design and the primary objective of such a transistor is to magnify the input signal. 'Forward Gain' can be termed as the amount of magnification which helps to decide the transistor in the design.

A typical Amplifier Block diagram is shown below:

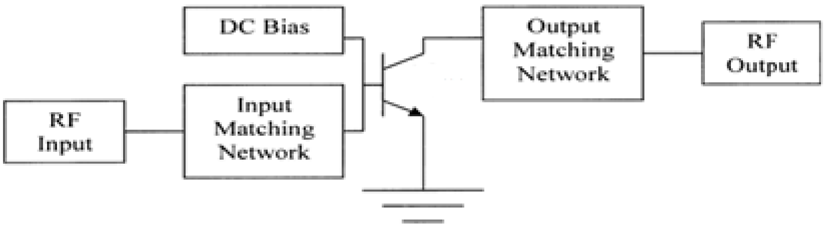


Figure 1.1 General Block Diagram of an Amplifier [1]

Low Noise Amplifier (LNA) is an amplifier which is used very often in the microwave and wireless communication. LNA helps to obtain a high gain with the least possible noise figure. [1]

As the design progressed, it was concluded that there was a need of multiple stages (possibly three) to achieve the decided goal. The design specifications are already discussed in the following chapter. The number of stages were decided on the basis of the output of a single stage Low Noise Amplifier and Maximum Gain Amplifier which will be used in a cascaded system to obtain the final objective. The higher the output of the single stage, the less number of multi-stages will be required to achieve the final objective.

Chapter 2. Design Specification

The goal of this project is to design a three stage LNA with a high gain and lowest possible noise figure for an amplifier.

The given parameters and specifications that have been decided are given in Table 2.1.

Parameters	**Goal**
Frequency	16 GHz
Noise Figure	≤ 1 dB
Gain	≥ 40 dB

Table 2.1 Parameters and Specifications

Different methods were used to accomplish the final objective of the project. Initially, it was crucial to determine the LNA gain and noise figure by hand calculations. These calculations were verified using simulation on MATLAB. This was followed by building a D.C. biasing network which was then accompanied with the design of the matching networks for the input and output with the help of Smith Charts. These values are verified graphically with the help of Microwave office and ADS software. Almost all the values have been verified again by a fourth software designed by the professor himself and is provided with his published textbook “RF & Microwave Design Essentials”.

Chapter 3. Classes and Types of Amplifier

3.1 Classification based on Class [2]

Amplifiers can be classified into the following categories:

Class A amplifier: The operation of every transistor for the amplifier in this class is in the active region. This happens for the entire signal cycle. Class A amplifiers are circuits with linear operation. The transistor is biased such that it conducts for the complete input signal cycle. Due to this, these amplifiers theoretically have 50% maximum efficiency. This category is used by most small-signal and low-noise amplifiers. [2]

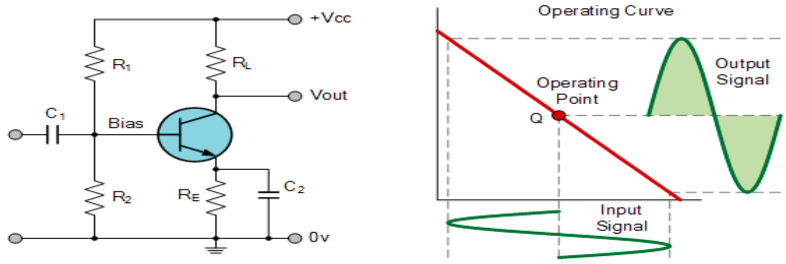


Figure 3.1 Class A Amplifier [2]

Class B amplifier: The operation of every transistor in this class is in active region for half of the signal cycle time. The transistors in this class are biased for this time period only. Typically, two complementary transistors are operated in push-pull amplifier of this class. This helps to achieve amplification for the complete cycle. Due to this, these amplifiers theoretically have 78% maximum efficiency. [2]

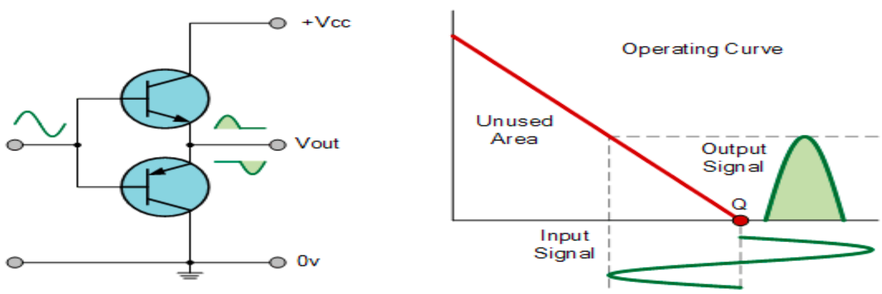


Figure 3.2 Class B Amplifier [2]

Class C amplifier: The operation of every transistor in this class is in its active region which lasts less than half of the signal cycle. Amplifier of this category functions with the transistor near cutoff for more than same time period. Typically, they use a circuit which resonant in the output. This helps to recuperate the fundamental signal. Due to this, these amplifiers are efficient 100%. It has a drawback as they need modulation for constant envelope. [2]

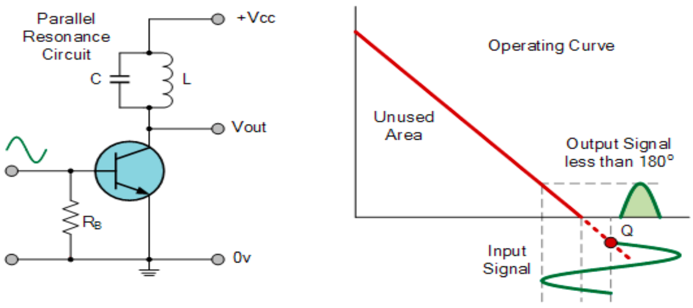


Figure 3.3 Class C Amplifier [2]

Class AB amplifier: The amplifiers in this category operates in both class A and class B for small and large signals respectively. [2]

Classes such as D, E, F, and S, which belong to high category need a switch transistor to pump a highly resonant tank circuit. This helps to obtain a high efficiency. Most of the transmitters operating in the UHF range or above need a class A, AB, or B power amplifiers. This is due to the requirement of the low distortion products. [2]

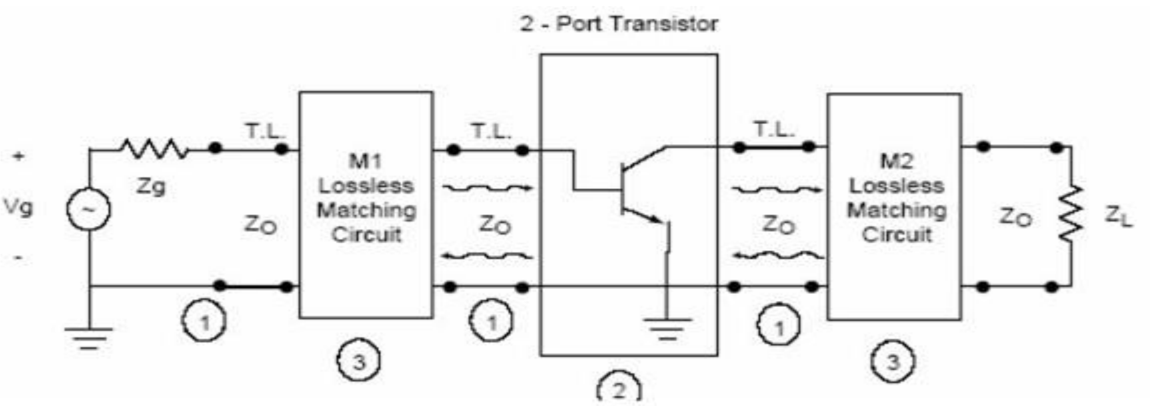


Figure 3.4 Block Diagram of a Single Stage Amplifier [2]

Figure 3.4 shows the matching networks M $_1$ and M $_2$ used to match the network with the transistor. The important function of matching network as the name suggests is to match the network with the transistor so that there is no reflection in the circuit. This is important because reflection which gets introduced due to mismatch causes a decrease in the gain, and if this reflection is removed, the circuit performs such that it gives the best possible gain.

3.2 Classification based on the level of signal. [2]

There are two categories and can be defined as:

3.2.1 Small Signal Amplifier- Definition: "Is a method of analysis of an active circuit in which it is assumed that the signals deviate from (or fluctuate to either side of) the steady bias levels by such a small amount that only a small part of the operating characteristic of the device is covered and thus the operation is always linear." [2]

This is further categorized in two types based on the design

3.2.1.A D.C. circuit design – Linear operation is needed from a transistor to work in small signal condition. This lead to the operation of amplifier in the class A type. The Q-point of the D.C. characteristics needs to be chosen in the mid-range of the graph I_C vs V_{CE} or I_D vs V_{DS} for BJT or FET respectively. [2] This ensures the transistor operation in the active region mode.

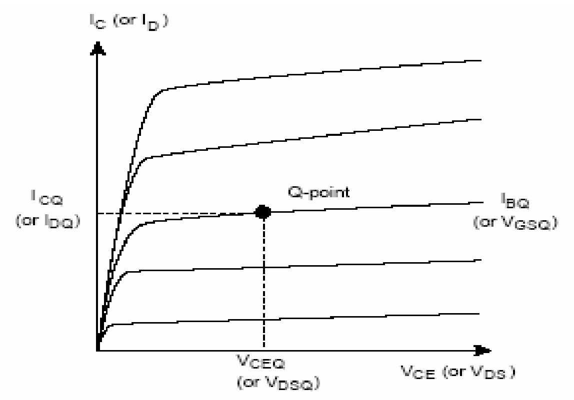


Figure 3.5 Transistor FET and BJT curve characteristic [2]

This D.C. biasing circuit is completely isolated from the RF signals and needs to be connected to the RF circuit. This creates minimum interaction and overcomes any leakage for the RF/microwave signals that are going to the output from the input side. An evaluation must be performed to confirm that there is no inappropriate coupling of the elements with the D.C. source or ground. This can be done by various methods as mentioned below:

a. Connecting an inductor in the circuit which is also known as an RF choke (RFC). This is connected between the RF/MW circuit and D.C. source. Practically, “Ferrite bead” is one of the simplest and commonly method to implement an RF choke. [2]

b. Connecting a capacitor (as a load) of very high value at the quarter-wave transformer. This commendably shorts any residual RF/MW signal that might leak into the D.C. circuit. An open circuit is created due to the use of this high valued capacitor at the input end near the RF/MW circuitry. [2]

c. Connecting a quarter-wave transformer between the RF/MW circuit and the D.C. source. Preferably, the characteristic impedance (Z_O) should be very high (i.e., $Z_O >> 1$) for a transformer. This will help to create a high impedance path for any RF/MW signal flowing in the circuit. [2]

3.2.1.B RF/ MW circuit design – (explained in the next chapter) [2]

3.2.2 Large Signal Amplifier: Definition: “Is a method of analysis of an active circuit under high amplitude signals that traverse such a large part of the operating characteristics of a device that nonlinear portions of the characteristic are usually encountered causing nonlinear operation of the device.” [2]

3.3 Classification based on Types of Amplifier [1]

Numerous design types of amplifiers are stated below:

3.3.1 Narrow Band Amplifiers (NBA) Design: The amplification for this design occurs around 10% bandwidth from the center frequency. [1]

3.3.2 High-Gain Amplifier (HGA) Design: They work on specific gain. This gain is not essentially the same as maximum available gain. [1]

For this design, a constant input and output gain circles are drawn on the basis of the required gain. Γ_S and Γ_L are selected such that they lie on the respective constant-gain circles. [1]

3.3.3 Maximum Gain Amplifier (MGA) Design: It is also known as a special case of HGA. [1] The input and the output gain circles get reduced to a point as follows:

$\Gamma_S = \Gamma_{IN}{*}$

$\Gamma_L = \Gamma_{OUT}{*}$

Such a modification leads further to two situations. [1]

3.3.3.A Unilateral Design: Before understanding the Unilateral design, first there is a need to understand the Unilateral Figure of Merit denoted by U. The U helps us to understand the error present in our analysis. [1]

In terms of magnitude, $\frac{G_T}{G_{TU,max}} = \frac{1}{(1-X)^2}$ where $X = \frac{S_{12} S_{21} \Gamma_S \Gamma_L}{(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_L)}$

The ratio of gain G_T to $G_{TU,max}$, is measured in dB and needs to be inside the error range that is tolerable. This is as shown below: [1]

$$\frac{1}{(1+|X|)^2} \leq \frac{G_T}{G_{TU}} \leq \frac{1}{(1-|X|)^2}$$

which for Unilateral Case can also be written as,

$$\frac{1}{(1+U)^2} \leq \frac{G_T}{G_{TU,max}} \text{ dB} \leq \frac{1}{(1-U)^2}$$

Where $\frac{1}{(1+U)^2}$ and $\frac{1}{(1-U)^2}$ are maximum tolerable errors (in dB) and value of U is given by

$$U = \frac{|S_{12}|\,|S_{21}|\,|S_{11}|\,|S_{22}|}{(1-|S_{11}|^2)(1-|S_{22}|^2)}$$

G_T = Transducer Gain $\qquad$ $G_{TU,\,MAX}$ = Maximum Available Transducer Gain

If the value of U ($\frac{G_T}{G_{TU,max}}$) < 0.06, it is a Unilateral Case. Else, we use the formulas for Bilateral Case which is mentioned later.

Thus, the MGA design equations is: [1]

$$\Gamma_S = S_{11}^* \text{ and } \Gamma_L = S_{22}^*$$

This state is used to estimate the transducer maximum gain ($G_{TU,\,max}$).

$$G_{TU,\,MAX} = G_{S,\,MAX} \cdot G_O \cdot G_{L,\,MAX} = \frac{1}{1-|S_{11}|^2} \cdot |S_{21}|^2 \cdot \frac{1}{1-|S_{22}|^2}$$

3.3.3.B Bilateral Design: In this case, Γ_S needs to be selected exactly equal to Γ_{MS}. Similar is the case for Γ_L to convert it to Γ_{ML}. [1]

Γ_{MS} and Γ_{ML} values can be calculated by the equations mentioned below and can be used to design the required matching networks.

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

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

$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2$ $\qquad$ $B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$

$C_1 = S_{11} - \Delta S^*_{22}$ $\qquad$ $C_2 = S_{22} - \Delta S^*_{11}$

The transducer has a maximum gain ($G_{T,\,max}$) given by:

$$G_{T,MAX} = G_{S,MAX} \cdot G_{O} \cdot G_{L,MAX} = \frac{1}{1-|\Gamma_{MS}|^2} \cdot |S_{21}|^2 \cdot \frac{1-|\Gamma_{ML}|^2}{|1-S_{22}\,\Gamma_{ML}|^2}$$

3.3.4 Low-Noise Amplifier (LNA) Design: These amplifiers help to amplify by maintaining the noise to a defined lower limit. The primary objective is to avoid exceeding the noise figure limit while attaining the highest possible gain. Practically, it is very difficult to obtain both. Therefore, it is required to decide a goal before proceeding with the design. [1]

In this type of amplifier design, the objective is not to exceed a specified noise figure value while achieving the highest possible gain. We trade-off between gain and noise figure since they cannot be achieved simultaneously. Under a given design requirement with a specific noise figure and an exact gain value, there is a need to carry out the following design procedure. [1]

Step 1. Calculate the assigned gain values for the matching networks at input and output.

Step 2. Plot the constant-gain circle and the constant-noise circle on the same Smith Chart for the source.

Step 3. Choose a source constant-gain circle to intercept the desired constant noise-figure circle.

Step 4. The input matching network can be designed using Γ_S (selected such that it is anywhere on the constant circle for gain. This point should lie between the intercept points and inside the constant noise figure circle).

Step 5. Finalize the amplifier circuit with the design of the matching network for the output. This is done by plotting the load constant-gain circle with the allocated gain from the above step, and by choosing a Γ_L on this circle.

3.3.5 Minimum Noise Amplifier is a special case of LNA. The noise figure circles are reduced to a single point (Γ opt) leading to the reduction of the design process to a single design choice. [1]

$$\Gamma_S = \Gamma_{opt}$$

After completing the first five steps of the design, the design of the matching networks can be initialized for the input and output.

After satisfying the above condition, in order to get the best VSWR at the output, choose:

$$\Gamma_L = \Gamma_{OUT}{}^* = \left(S_{22} + \frac{S_{12}\, S_{21}\, \Gamma_{opt}}{1 - S_{11}\, \Gamma_{opt}}\right)$$

The matching networks for the input and output can be easily designed by Γ_S and Γ_L . [1]

In the design, since the unilateral figure of merit U was very high, there was a need to modify the above method such that based on the Γ_{MS} and Γ_{ML} values, Γ_S and Γ_L values were calculated. After getting Γ_S and Γ_L, the values of the Noise Figure circle was calculated to make sure that the design gives noise within the design specification limit. This process had to be repeated i.e. with repetitive iteration to get a stable Γ_S and Γ_L value before finalizing it. This was followed by the matching network for the LNA for the input and output side, thus completing a single stage LNA. [1]

The next chapter explains each step of the design in detail.

Chapter 4. Steps to Design an Amplifier

4.1 Introduction

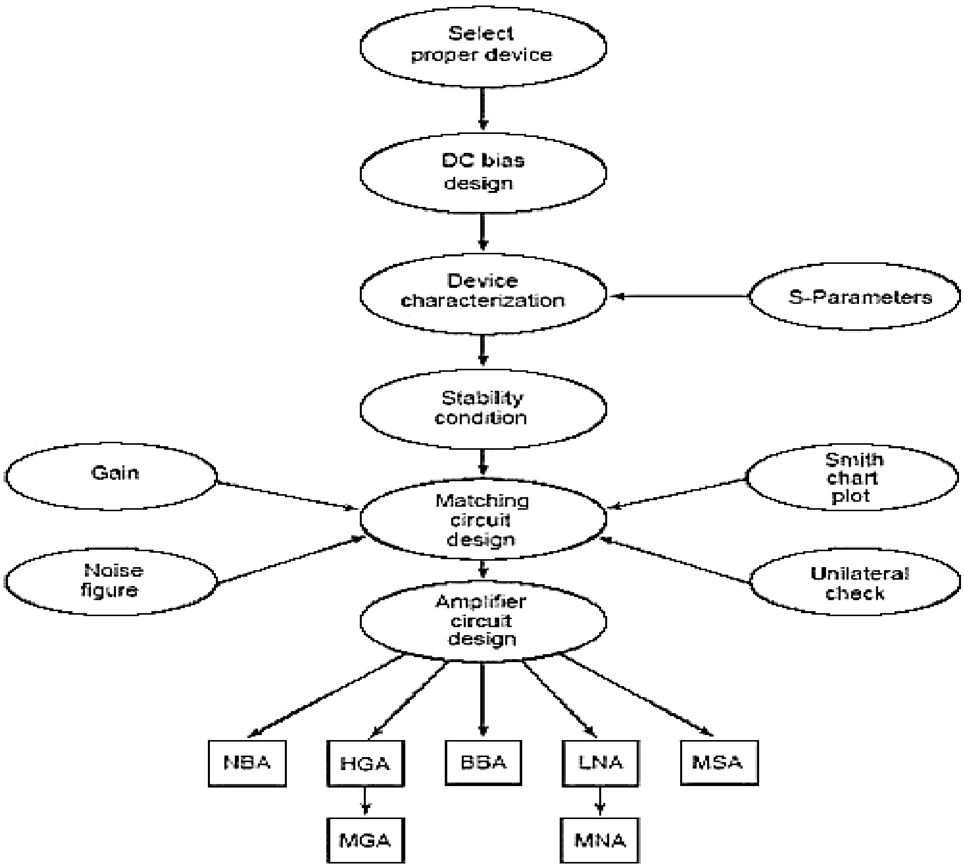


Figure 4.1: Design Steps for any Amplifier [1]

There is a need to follow the steps as mentioned below to design the microwave circuit as mentioned in Chapter 15 from the book "RF and Microwave Design Essentials" by Prof. Dr. Matthew Radmanesh.

The design steps are mentioned in the order mentioned in the above Figure 4.1 along with the detailed explanation of each step for easy understanding. [1]

This is the first step. From the specifications of the amplifier, if gain (G) is given then there is a need to select the transistor which has a gain higher than the result obtained by dividing the S-parameters S_{21} and S_{12} ($| S_{21} / S_{12} | > G$). This is operated in the desired frequency range. Another possibility is to check the given value of the noise figure F_0, there is a need to make sure that the selected transistor is such that the F_0 is higher than the minimum noise figure, F_{min} ($F_0 > F_{min}$) from the datasheet of the selected transistor. [1]

4.2 D.C. Biasing

This is the second step. The transistor needs to be biased on the basis of the curves of I_C - V_{CE} for BJT or I_D - V_{DS} for FET. FET is biased in the I_D - V_{DS} curves of the transistor. [1]

The book "RF Microwave and Design Essentials" by Professor Dr. Matthew Radmanesh shows a suitable D.C. biasing method that needs to be selected from appendix K.

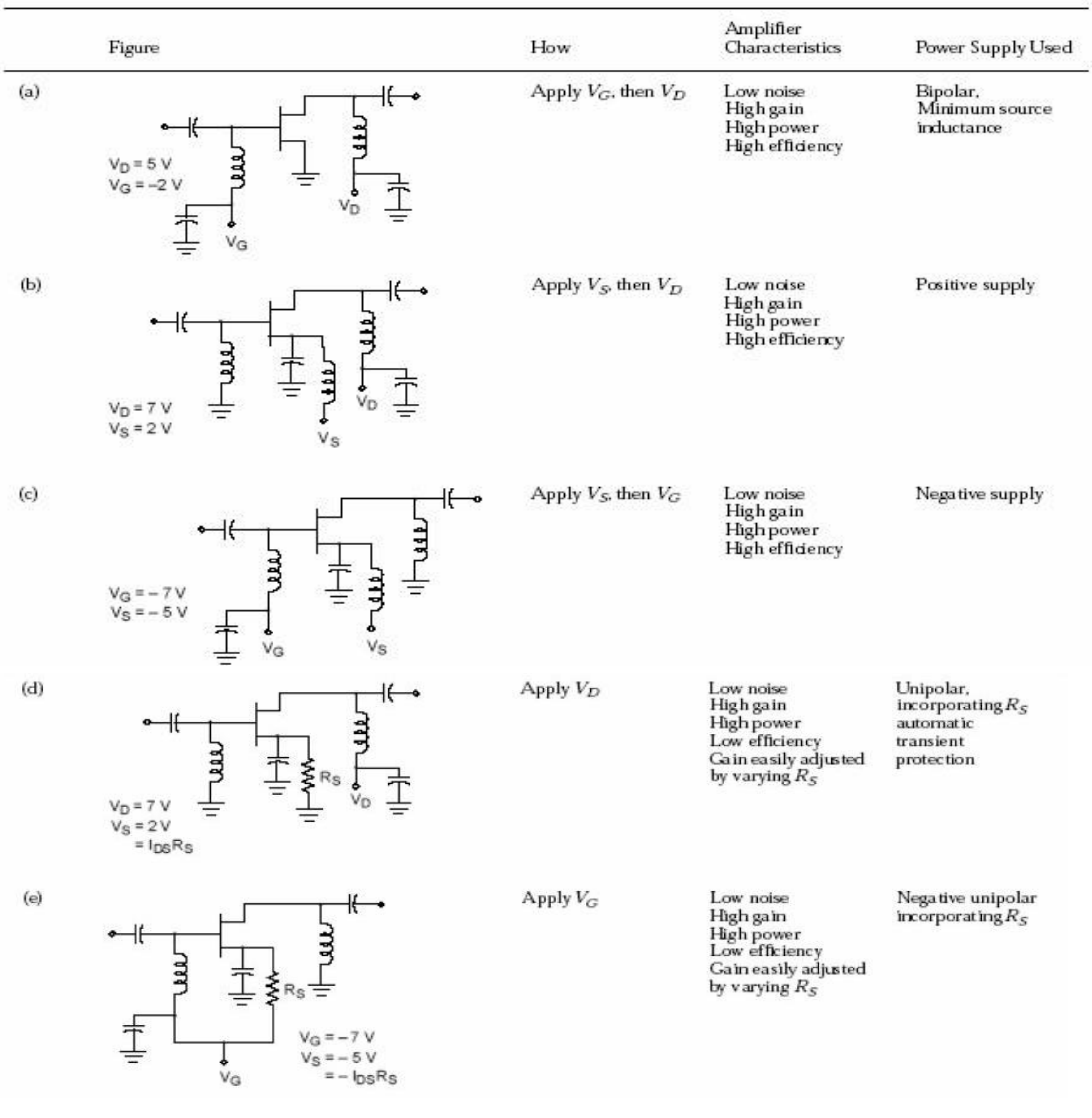


Figure	How	Amplifier Characteristics	Power Supply Used
(a) $V_D = 5$ V, $V_G = -2$ V	Apply V_G, then V_D	Low noise High gain High power High efficiency	Bipolar, Minimum source inductance
(b) $V_D = 7$ V, $V_S = 2$ V	Apply V_S, then V_D	Low noise High gain High power High efficiency	Positive supply
(c) $V_G = -7$ V, $V_S = -5$ V	Apply V_S, then V_G	Low noise High gain High power High efficiency	Negative supply
(d) $V_D = 7$ V, $V_S = 2$ V $= I_{DS}R_S$	Apply V_D	Low noise High gain High power Low efficiency Gain easily adjusted by varying R_S	Unipolar, incorporating R_S automatic transient protection
(e) $V_G = -7$ V, $V_S = -5$ V $= -I_{DS}R_S$	Apply V_G	Low noise High gain High power Low efficiency Gain easily adjusted by varying R_S	Negative unipolar incorporating R_S

Figure 4.2: D.C. Biasing Techniques [1]

Option (e) was used from all the above techniques for D.C. biasing in the design.

4.3 Device Characterization

This is the third step. The selected Q-point helps to decide the S-parameters. These parameters are specified in the data sheet. [1]

4.4 Stability Test

This is the fourth step. At the desired frequency, the condition of stability is verified. In general, Stability can be termed as the ability of an amplifier to maintain its effectiveness in its nominal operating characteristics. This maintained effectiveness is present in spite of large changes in the environment. These changes can be physical temperature, signal frequency, source or load conditions, etc. [1]

The Stability test is done by two methods as follows.

A) Analytical Method

Then based on these definitions, a two-Port network will be unconditionally stable if, and only if, either one of the following mathematically equivalent criteria are satisfied: [1]

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

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

I) Method # 1: Three-Parameter Test

$$K > 1, \quad \text{and} \quad \frac{1 - |S_{11}|^2}{|S_{12}\, S_{21}|} > 1 \quad \text{and} \quad \frac{1 - |S_{22}|^2}{|S_{12}\, S_{21}|} > 1$$

The three parameter test can be modified into a two-parameter test which is also known as the “K - Δ Test” and is mentioned below. [1]

II) Method # 2: Two-Parameter Test (K - Δ Test)

$$K > 1 \text{ and } |\Delta| < 1$$

This test is the most commonly used and popular because of its simplicity and ease of calculations. [1]

III) Method # 3: Two-Parameter Test (K - β_1 Test)

$$K > 1 \text{ and } \beta_1 > 0$$

These three criteria are equivalent to each other. Therefore, if a device satisfies any one of these three criteria, the other two are also satisfied.

Thus, a two - Port network will be unconditionally stable if and only if any one of the above three criteria are satisfied.

For a unilateral transistor, we have $S_{12} = 0$ making K infinite > 1. Also, $| \Delta | = | S_{11} S_{22} |$

Since K > 1 has already been satisfied, therefore, in order to satisfy the condition for unconditional stability we desire $| \Delta | < 1$, which requires:

a. $| S_{11} | < 1$

b. $| S_{22} | < 1$

for all passive values of Z_S and Z_L.

IV) Method # 4: Single-Parameter Test (μ-Parameter Test)

The “two-parameter test” (“K - Δ test”) described above, provides a set of mathematical conditions on two parameters for unconditional stability and only indicates whether a device is stable or not. The “two-parameter” test cannot be used to show the degree of stability of a device with respect to any other similar device due to the constraints that are imposed on those two parameters. [1]

To determine both, the unconditional stability of a device and the stability with respect to other devices, another method had been created which combines the “K - Δ parameters” into a ‘single-parameter test’. It is known as “μ - parameter test”. The parameter “μ” is defined as:

$$\mu = \frac{1 - |S_{11}|^2}{|S_{22} - S_{11}^* \Delta| + |S_{12} S_{21}|}$$

For unconditional stability, the following must be satisfied:

$$\mu > 1$$

Furthermore, if "device A" has a parameter "μ_A" which is greater than "μ_B" corresponding to "device B" [1], i.e. $\mu_A > \mu_B$

Then "device A" is said to be more stable than "device B".

The above equation indicates that a device with a larger value of "μ" is more desirable for an amplifier design since it implies a greater degree of stability. [1]

B) Graphical Method

For stable systems, next step is implemented directly. If the condition does not get satisfied, there is a need for the circles of stability at the input and output to locate the stable regions. [1]

A graphical method is used for this. The S-parameters are used in the determination of the input and output stability circles.

The input stability circle drawn in the Γ_S - plane is obtained from the equations:

$$|\Gamma_S - C_S| = R_S$$

Where,

$$D_S = |S_{11}|^2 - |\Delta|^2$$

$$C_S = \frac{(S_{11} - \Delta S_{22}{}^*)^*}{D_S}$$

$$R_S = \left| \frac{S_{12} * S_{21}}{D_S} \right|$$

The output stability circle drawn in Γ_L - plane is obtained from the equations:

$$|\Gamma_L - C_L| = R_L$$

Where:

$$D_L = |S_{22}|^2 - |\Delta|^2$$

$$C_L = \frac{(S_{22} - \Delta\ S_{11}{}^{*})^{*}}{D_L}$$

$$R_L = \left| \frac{S_{12} * S_{21}}{D_L} \right|$$

For LNA, the focus usually is on the input stability circle. The output stability circle is not usually considered since a stable input is bound to give an output within the stable region.

A special case of unconditional stability for the given specifications are kept into consideration:

For $|S_{11}| < 1$ and $|S_{22}| < 1$, the amplifier circuit will be stable unconditionally when either of the following two conditions hold true:

a) Both stability circles fall outside the Smith Chart completely, or
b) Both stability circles enclose the Smith Chart completely.

Therefore, the unconditional stability can be concisely stated in mathematical form for all passive source and load impedances as below:

$$\left| |C_L| - R_L \right| > 1 \qquad \text{for } |S_{11}| < 1$$

$$\left| |C_S| - R_S \right| > 1 \qquad \text{for } |S_{22}| < 1$$

4.5 Important Design Parameters

This is the fifth step and involves Unilateral plot, Gain and Noise and Mismatch Factor.

4.5.1 Unilateral Figure of Merit [1]

The crucial test at input and output of an LNA is to design the matching network such that they render close to zero. The LNA needs to be matched to the source and load ports. This leads to two scenarios:

Unilateral Case: The unilateral design is followed if $S_{12} = 0$.

If $S_{12} \neq 0$, then there is a need to estimate the unilateral figure of merit also known as U. This will help to calculate the range of error. For small range i.e. around ±5dB, then the unilateral design is used (if $U < 0.06$). If not, then the bilateral design is followed (if $U > 0.06$).

$$U = \frac{|S_{11}||S_{21}||S_{12}||S_{22}|}{(1-|S_{11}|^2)(1-|S_{22}|^2)}$$

The formula to evaluate the range of error is,

$$\frac{1}{(1+U)^2} < R < \frac{1}{(1-U)^2}$$

When the transistor is unilateral, then simplify into the following:

$$\Gamma_{IN} = S_{11}$$

$$\Gamma_{OUT} = S_{22}$$

Bilateral Case: When $S_{12} \neq 0$ and unilateral Figure of merit causes an unjustifiably high error in the gain equations, we are faced with the bilateral case where S_{12} can no longer be ignored. [1]

We know that from above equations, the maximum power transfer occurs when:

$$\Gamma_{IN} = \Gamma_{s}^{*} = S_{11} + \frac{S_{12} . S_{21} . \Gamma_L}{1 - S_{22} . \Gamma_L}$$

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

Γ_S should be selected equal to Γ_{MS} and Γ_L should be selected equal to Γ_{ML}.

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

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

Where,

$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2$ $B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2$

$C_1 = S_{11} - \Delta S^{*}_{22}$ $C_2 = S_{22} - \Delta S^{*}_{11}$

LNA can be designed by calculating Γ_{MS} and Γ_{ML} using the calculated values from the above formulas. [1]

After Γ_{MS} and Γ_{ML} calculations, the Smith Chart is used for matching networks of input and output.

4.5.2 Gain Calculations

Ideal Gain Calculation is given by $G_{MSG} = \frac{|S_{21}|}{|S_{12}|} \Rightarrow 10\log_{10}\frac{|S_{21}|}{|S_{12}|} = G$ dB

4.5.2.A Transducer Power Gain (G_T) is defined as "the ratio of power delivered to the load of the system (P_L) by the available power from the source under matched condition (P_{AVS})." [1]

$$G_T = \frac{P_L}{P_{AVS}} = G_S \cdot G_0 \cdot G_L = \left(\frac{1-|\Gamma_S|^2}{|1-\Gamma_{IN}\Gamma_S|^2}\right) \cdot |S_{21}|^2 \cdot \left(\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}\right)$$

$$= \left(\frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2}\right) \cdot |S_{21}|^2 \cdot \left(\frac{1-|\Gamma_L|^2}{|1-\Gamma_{out}\Gamma_L|^2}\right)$$

4.5.2.B Operating Power Gain or Power Gain (G_P) is defined as "the ratio of power delivered to the load of the system (P_L) by the power given as input to the transistor (P_{IN})." [1]

$$G_T = \frac{P_L}{P_{IN}} = G_S \cdot G_0 \cdot G_L = \left(\frac{1}{1-|\Gamma_{IN}|^2}\right) \cdot |S_{21}|^2 \cdot \left(\frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2}\right)$$

4.5.2.C Available Power Gain (G_A) is defined as "the ratio of power available from the source of the system (P_{AVN}) under matched condition by power available from the transistor (P_{AVS})." [1]

$$G_T = \frac{P_{AVN}}{P_{AVS}} = G_S \cdot G_0 \cdot G_L = \left(\frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2}\right) \cdot |S_{21}|^2 \cdot \left(\frac{1}{1-|\Gamma_{out}|^2}\right)$$

where,

$$\Gamma_{IN} = \Gamma_S^* = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L}$$

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

The Gain formula for special cases Unilateral and Bilateral is as mentioned below.

Unilateral Case Gain:

$$G_{SU} = \left(\frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2}\right) . |S_{21}|^2 . \left(\frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2}\right)$$

The maximum gain under Unilateral Case is given by,

$$G_{TU,MAX} = G_{S,MAX} . G_0 . G_{L,MAX} = \left(\frac{1}{1 - |S_{11}|^2}\right) . |S_{21}|^2 . \left(\frac{1}{1 - |S_{22}|^2}\right)$$

Bilateral Case (Maximum) Gain:

$$G_{T,MAX} = G_{S,MAX} . G_0 . G_{L,MAX} = \frac{1}{1 - |\Gamma_{MS}|^2} . |S_{21}|^2 . \frac{1 - |\Gamma_{ML}|^2}{|1 - S_{22}\Gamma_{ML}|^2} = \frac{|S_{21}|}{|S_{12}|}\left(K - \sqrt{K^2 - 1}\right)$$

4.5.3 Noise

Noise is an unwanted factor that gets added to the system due to many reason. It degrades the performance of the system and hence is not desirable. It gets added due to external or internal factors acting on the system. Ideally, it cannot be removed completely, however it can be reduced to the lowest possible value.

Electrical Noise (Noise): Definition: "Any unwanted electrical disturbance or spurious signal. These unwanted signals are random in nature and are generated either internally in the electronic components or externally through impinging electromagnetic radiation." [1]

Noise gets added due to the following reasons. [1]

i. Thermal vibrations of atoms, molecules, electrons etc.
ii. Flow of these electrons across a wire in the system.
iii. Emission of the charges from a cathode of a diode, electron etc.
iv. Propagation of the waves through the atmosphere.

Another method to characterize the noise of an amplifier is by using the concept of **Noise Figure.**

Noise Figure: Definition: "It is the ratio of the total available noise power at the output, to the output available noise power due to thermal noise coming only from the input resistor at the standard room temperature (T_O = 290 °K)." [1]

$$P_N = G_{\mathbf{A}} \,.\, k \,.\, T_{\mathbf{e}} \,.\, B$$

where k is Boltzmann Constant

T_e is Equivalent Noise Temperature B is the Bandwidth

$(P_O)_I = G_{\mathbf{A}} \,.\, P \,.\, N\, i = G_{\mathbf{A}} \,.\, k \,.\, B \,.\, T_O$ and $(P_O)_{tot} = P_{NO} = P_N + (P_O)_i$

For Analysis, we need to use formula mentioned below for the calculation of noise figure circles:

$$F = F_{\mathbf{min}} + \frac{\frac{4\,R_N}{Z_0}\left(\,\left|\Gamma_S - \Gamma_{opt}\right|^2\right)}{\left(\,1 - |\,\Gamma_S\,|^2\right).\left|\,1 + \Gamma_{opt}\right|^2}$$

$$N = \frac{F - F_{\mathbf{min}}}{\frac{4\,R_N}{Z_0}}\left|\,1 + \Gamma_{opt}\right|^2 = \frac{\left|\,\Gamma_S - \Gamma_{opt}\right|^2}{1 - |\,\Gamma_S\,|^2}$$

$$r_{\mathbf{n}} = \frac{R_N}{Z_0}$$

where N = Noise Figure parameter

F = required noise figure in ratio

F_{min} = Optimum noise figure in ratio

Γ_{opt} = Reflection coefficient to achieve optimum noise

R_N = Equivalent noise resistance of transistor

The center of noise figure circle is plotted in $\Gamma_{\mathbf{s}}$ plane:

$$C_F = \frac{\Gamma_{opt}}{1 + N}$$

Radius of the noise figure circle:

$$R_F = \frac{\sqrt{N^2 + N(1 - |\Gamma_{opt}|^2)}}{1 + N}$$

The values of F_{min}, R_N (or r_n) and Γ_{opt} are given in the manufacturer's data sheet.

4.5.4 Mismatch Factor

This value signifies the mismatch that is preset between the load and the matching network. It is important to know this value since ignoring this can lead to a drop in the efficiency of the system in the form of drop of gain or increase of noise.

Mismatch factor also known as Mismatch Loss is specified in dB and determines the power loss due to mismatch. [1]

Source Mismatch factor is given by $\mathbf{M_S} = \frac{P_{IN}}{P_{\mathbf{AVS}}} = \frac{(1-|\Gamma_S|^2).(1-|\Gamma_{IN}|^2)}{(|1-\Gamma_S \Gamma_{IN}|^2)}$

$$\mathbf{M_S\ (dB)} = P_{IN}\ \mathbf{(dBm)} - P_{AVS}\ \mathbf{(dBm)}, \quad \mathbf{M_S\ is\ less\ than\ 0}$$

Load Mismatch factor is given by $\mathbf{M_L} = \frac{P_L}{P_{\mathbf{AVN}}} = \frac{(1-|\Gamma_L|^2).(1-|\Gamma_{OUT}|^2)}{(|1-\Gamma_L \Gamma_{OUT}|^2)}$

$$\mathbf{M_L\ (dB)} = P_L\ \mathbf{(dBm)} - P_{AVN}\ \mathbf{(dBm)}, \quad \mathbf{M_L\ is\ less\ than\ 0}$$

Ideally, the value of M_S and M_L should be equal to 1, but it lies in the range of 0 to 1.

Input and Output VSWR: Input VSWR is the input power P_{IN} going into the input port of matching network M_1 using the input reflection coefficient Γ_a [1]

$$M_{\mathbf{S}} = 1 - |\Gamma_a|^2 \Rightarrow \Gamma_a = \sqrt{1 - M_{\mathbf{S}}} \qquad |\Gamma_a| = \frac{(Z_a - Z_0)}{(Z_a + Z_0)}$$

$$(\mathrm{VSWR})_{IN} = \frac{1 + |\Gamma_a|}{1 - |\Gamma_a|} = \frac{1 + \sqrt{1 - M_{\mathbf{S}}}}{1 - \sqrt{1 - M_{\mathbf{S}}}}$$

In the same way, the output VSWR can be defined as the output power P_{out} leaving the output port of matching network M_2 using the input reflection coefficient Γ_b

$M_{\mathbf{L}} = 1 - |\Gamma_b|^2 \Rightarrow \Gamma_b\sqrt{1 - M_{\mathbf{L}}}$ $\qquad |\Gamma_b| = \dfrac{(Z_b - Z_0)}{(Z_b + Z_0)}$

$$(\mathrm{VSWR})_{out} = \frac{1 + |\Gamma_b|}{1 - |\Gamma_b|} = \frac{1 + \sqrt{1 - M_{\mathbf{L}}}}{1 - \sqrt{1 - M_{\mathbf{L}}}}$$

4.6 Matching Networks

This is the sixth step. It describes the rules for Matching Networks.

For the LNA, it is required to determine its matching network at the input and output. [1]

4.6.1 Rules for Lumped Elements on Smith Chart [1]

Based on the discussion presented in the previous two sections, there are certain rules that if followed would simplify and even speed up the matching circuit design process. These rules can be summarized as follows: [1]

Rule #1 Always use a ZY Smith Chart.

Rule #2 Start off from the load end at all times. Move “towards the Generator” in order to avoid any uncertainty or misunderstanding about the starting position.

Rule #3 Always move on a constant-R or constant-G circle in such a way as to arrive eventually at the center of the Smith Chart.

Rule #4 Each motion along a constant-R or constant-G circle gives the value of a reactive element.

Rule #5 Moving on a constant-R circle yields series reactive elements, whereas moving on a constant-G circle yields shunt reactive elements.

Rule#6 The travel (or motion) direction on a constant-R or constant-G circle determines the type of element to be used, i.e. a capacitor or an inductor.

Rules 5 and 6 lead to the following additional two rules.

Rule #7 When the motion is upward, in most cases it corresponds to a series or a shunt inductor.

Rule #8 When the motion is downward, in most cases it corresponds to a series or a shunt capacitor.

4.6.2 Rules for Distributed Elements on Smith Chart [1]

At higher frequencies where the component or circuit size is comparable with wavelength, matching of the load to the transmission line is done with distributed components. The most common technique in this type of design is the use a transmission line (called a stub). This is for a single open circuited length or short circuited length. From the load, this is either a parallel or in series connection with the transmission feed line and is at a distance. [1]

In single stub matching networks, the two variable limitations are the distance "d" from load to the stub and the length "l" called as stub lengths which provides the value of stub susceptance or reactance. [1]

Selection of distance "d" is crucial for both shunt and series stub as explained below:

a. For the shunt stub, "d" should be chosen such that the input admittance Y_P is in the form of $Y_P = Y_O + jB$ with the susceptance of the stub selected as (-jB). This results in a matched condition. [1]

b. For the series stub, "d" should be chosen such that the Z_S which is the impedance is of the form $Z_S = Z_O + jX$ with the reactance of the stub selected as $-jX$. This also results in a matched condition. [1]

Choice of Short- or Open-Circuited Stubs - With a λ/4 difference in length between the two, a short or open transmission line with proper length can provide reactance or susceptance of any value for the needed design. [1]

Structural considerations behind choice of a short- versus an open stub are as follows:

a. Open Stubs: For micro strip and strip line technology use of open stubs are preferred. Use of short stubs require a hole which goes to the ground plane through the substrate, which adds extra work and can be eliminated through the use of open circuits.

b. Short Stubs: For coaxial line or waveguide as a transmission line media, use of short stubs are preferred because the open stubs may radiate causing power losses thus making the stub no longer a purely reactive element.

Stub Realization Using Micro Strip Lines - Series transmission lines and shunt stubs (short or open) can easily be realized using design steps for micro strip line technology. [1]

Given a dielectric constant (ε_r), its height (h) and a certain characteristic impedance value (Z_O), the width of the micro strip line (W) can be calculated. The values can be verified using various software's available. Two tools were used to verify the stub calculations. This is explained in the next chapters.

Chapter 5. Design Calculations for the Amplifier

5.1 S-Parameter

At 16 GHz, the S-matrix for the FET FHX13/14LG is

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} 0.564\angle 148.8^\circ & 0.058\angle -12.8^\circ \\ 2.244\angle -50.7^\circ & 0.677\angle 172.6^\circ \end{bmatrix}$$

The simulation of the S parameters is as shown below for the transistor used in the design.

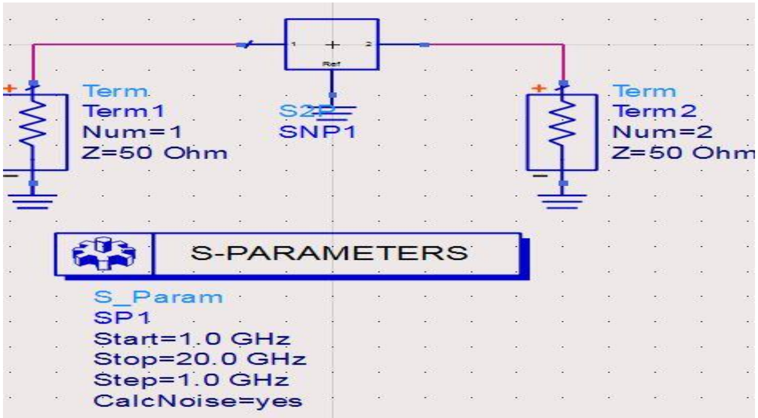


Figure 5.1: Simulation Circuit for S parameters

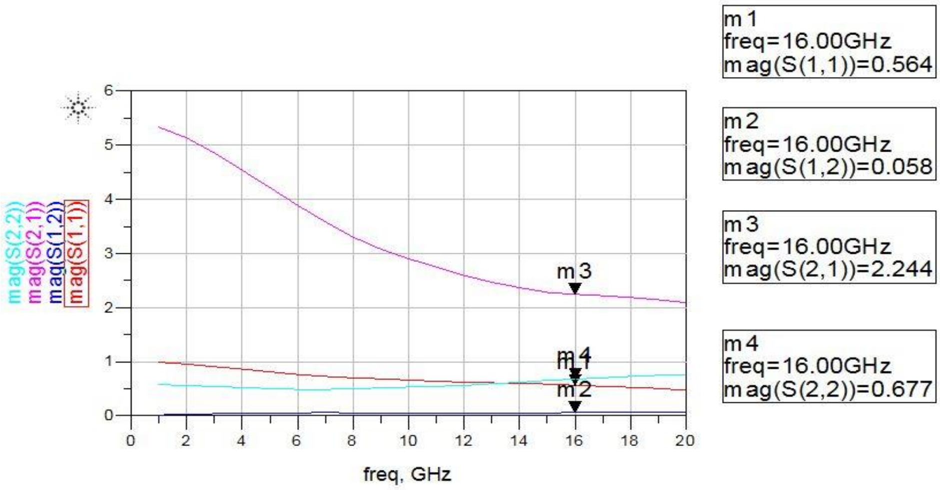


Figure 5.2: Simulation results for S parameters at 16 GHz

After simulating the circuit for the transistor, the S parameters can be seen for a wide frequency range. In this, we can check the values for our design as well which is at 16 GHz. This analysis are obtained by using the .S2P file of the transistor.

Specifications: Frequency = 16 GHz, Zo = 50Ω, ε_r = 10.7 (Rodgers Data Sheet), h = 0.127 mm

5.2 D.C. Biasing Calculations

The following values are considered from the transistor datasheet

V_t = - 0.7 V I_{ds} = 10 mA

V_{ds} = 2 V I_{dss} = 30 mA

First calculate V_{gs} using equation,

$I_{ds} = K (V_{gs} - V_t)^2$

$K = \frac{I_{dss}}{V_t^{\,2}} = \frac{30\,.\,10^{-3}}{(-0.7)^2} = 0.06122$

Substituting the value of K in the I_{ds} formula above, it can be seen that,

V_{gs} = - 0.3V and -1.1V

V_{gs} = -1.1V was not acceptable as it did not match the experimental value that can be observed for the graph of the transistor from the datasheet of -0.4V unlike the value of -0.3V

The value of Vgs from datasheet and above calculation can be concluded to -0.4V

Now, by Voltage Divider Rule,

$$-2.4 = \left(\frac{R_{G2}}{R_{G1} + R_{G2}}\right)(-7)$$

Assuming value of R_{G2} = 1MΩ, we get the value of R_{G1} = 1.91 MΩ

Now, $R_S = \frac{V}{I} = \frac{5}{10\,.\,10^{-3}} = 500\ \Omega$

RF Choke and D.C. Block Calculations:

RF Choke: Z_L is 10 times of $Z_{0,}$

$|Z_L| = 10 . Z_0 = 10 . 50\ \Omega = 500\ \Omega$ and $\omega = 2\pi F = 2\pi \times 16$ GHz $= 32\pi . 10^9$ rad/sec

$|Z_L| = \omega L \Rightarrow 500 = 32\pi . 10^9 . L \Rightarrow L = 4.976$ nH

D.C. Block: Z_C is 1/10 times of Z_0

$|Z_C| = \frac{1}{10} . Z_0 = \frac{1}{10} . 50 = 10 . 50\ \Omega = 5\ \Omega$ and $\omega = 32\pi . 10^9$ rad/sec

$|Z_C| = \frac{1}{\omega C} \Rightarrow 5\ \Omega = \frac{1}{32\pi . 10^9 . C}$ leading to C = 1.99 pF

In the circuit, the value of $|Z_C|$ was modified to a value of 2 Ω instead of 5 Ω leading to value of C = 4.97 pF. This modification is due to a requirement of the gap capacitance which should be 1/3rd of the width w.

When $|Z_C| = 5\ \Omega$, then C = 1.99 pF leading to a Gap Capacitance of S = 0.109 mm which is approximately same as width w = 0.1136 mm and hence is unacceptable.

Modifying Z_C to 2 Ω, it gave C = 4.97 pF leading to Gap Capacitance of S = 0.0409 mm which is approximately 1/3rd of width w = 0.1136 mm. This value is acceptable. The important condition here is that S (gap capacitance) is equal to 1/3rd of w (width).

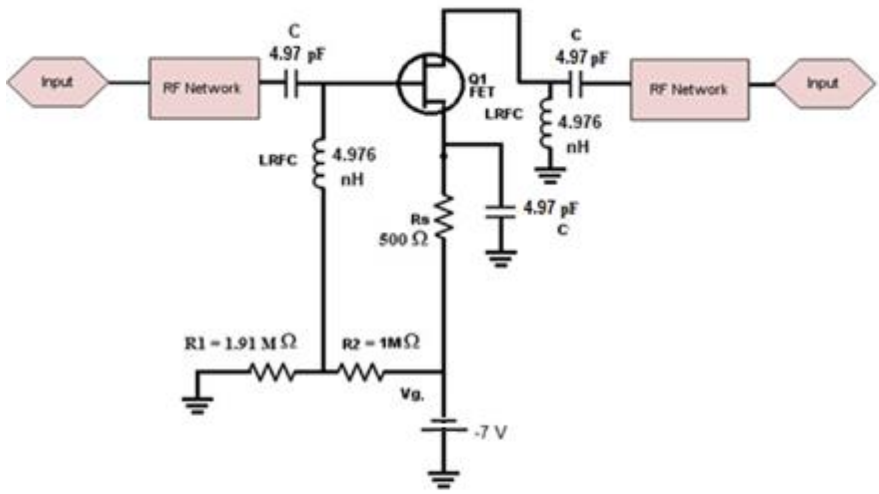


Figure 5.3 Final D.C. Biasing Circuit

5.3 Stability Test

$$\Delta = S_{11} S_{22} - S_{12} S_{21}$$
$$= (0.564\angle 148.8°)(0.677\angle 172.6°)$$
$$-(0.058\angle -12.8°)(2.244\angle -50.7°)$$

$$= 0.267\angle -26.5°$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + \Delta^2}{2\,|(S_{12})(S_{21})|} = \frac{1 - |0.564|^2 - |0.677|^2 + (0.267\angle -26.5°)^2}{2\,|(0.058)(2.244)|} = 1.134$$

5.4 Unilateral Plot Calculations

$$U = \frac{|S_{11}|\,|S_{21}|\,|S_{12}|\,|S_{22}|}{(1 - |S_{11}|^2)(1 - |S_{22}|^2)}$$
$$= \frac{|0.564|\,|2.244|\,|0.058|\,|0.677|}{(1 - |0.564|^2)(1 - |0.677|^2)} = 0.1345$$
$$> 0.6$$

$$\frac{1}{(1+U)^2} < R < \frac{1}{(1-U)^2} => 0.784 < R < 1.3349 => -1.05\text{ dB} < R < 1.25\text{ dB}$$

U > 0.6 and the error range > ± 0.5 dB, Hence it is a Bilateral Case.

5.5 Bilateral Case Calculations

$$\Gamma_{MS} = \frac{B_1 \pm \sqrt{B_1^2 - 4\,|C_1|^2}}{2C_1} = \frac{0.787 \pm \sqrt{(0.787)^2 - 4\,(0.38)^2}}{2(0.38\angle 143.3°)} = 0.832\angle -143.3°$$

$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 = 1 + (0.564)^2 - (0.677)^2 - (0.267)^2 = 0.787$

$C_1 = S_{11} - \Delta S_{22}^* = (0.564\angle 148.8°) - (0.267\angle 26.5°)(0.677\angle -172.6°) = 0.38\angle 143.3°$

$$\Gamma_{ML} = B_2 \pm \frac{\sqrt{B_2^2 - 4\,|C_2|^2}}{2C_2} = \frac{1.068 \pm \sqrt{(1.0677)^2 - 4\,(0.53)^2}}{2(0.53\angle 169.2°)} = 0.886\angle -169.2°$$

$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 = 1 + (0.677)^2 - (0.564)^2 - (0.267)^2 = 1.0677$

$C_2 = S_{22} - \Delta S_{11}^* = (0.677\angle 172.6°) - (0.267\angle 26.5°)(0.564\angle -148.8°) = 0.53\angle 169.2°$

5.6 Noise

F_{min} = 0.63 dB = 1.156, $\Gamma_{opt} = 0.38\angle -175$, r_n = 0.04 (From Datasheet)

F = 1 dB = 1.259

$$N = \frac{|\Gamma_s - \Gamma_{opt}|^2}{1-|\Gamma_s|^2} = \frac{F - F_{min}}{4 \cdot r_n} |1 + \Gamma_{opt}|^2 = \frac{1.259 - 1.156}{4\,(0.04)} |1 + 0.38\angle -175|^2 = 0.247$$

Center is at, $C_F = \frac{\Gamma_{opt}}{1+N} = \frac{0.38\angle -175}{1 + 0.247} = 0.305\angle -175$

Noise Figure circle radius is:

$$R_F = \frac{\sqrt{N^2 + N(1 - |\Gamma_{opt}|^2)}}{1 + N} = \frac{\sqrt{0.247^2 + 0.247(1 - |0.38|^2)}}{1 + 0.247} = \frac{0.411}{1.166} = 0.418$$

With these values, the Noise figure circle are plotted and the value of Γ_S and Γ_L are determined. With repetitive iterations on the Smith Chart, we try to get these values as much stable and constant as possible. By selecting a value on the Noise figure circle closer to Γ_{MS}, we try to select Γ_S as shown in Smith Chart below. We have selected to be $\Gamma_S = 0.58\angle -143.3$.

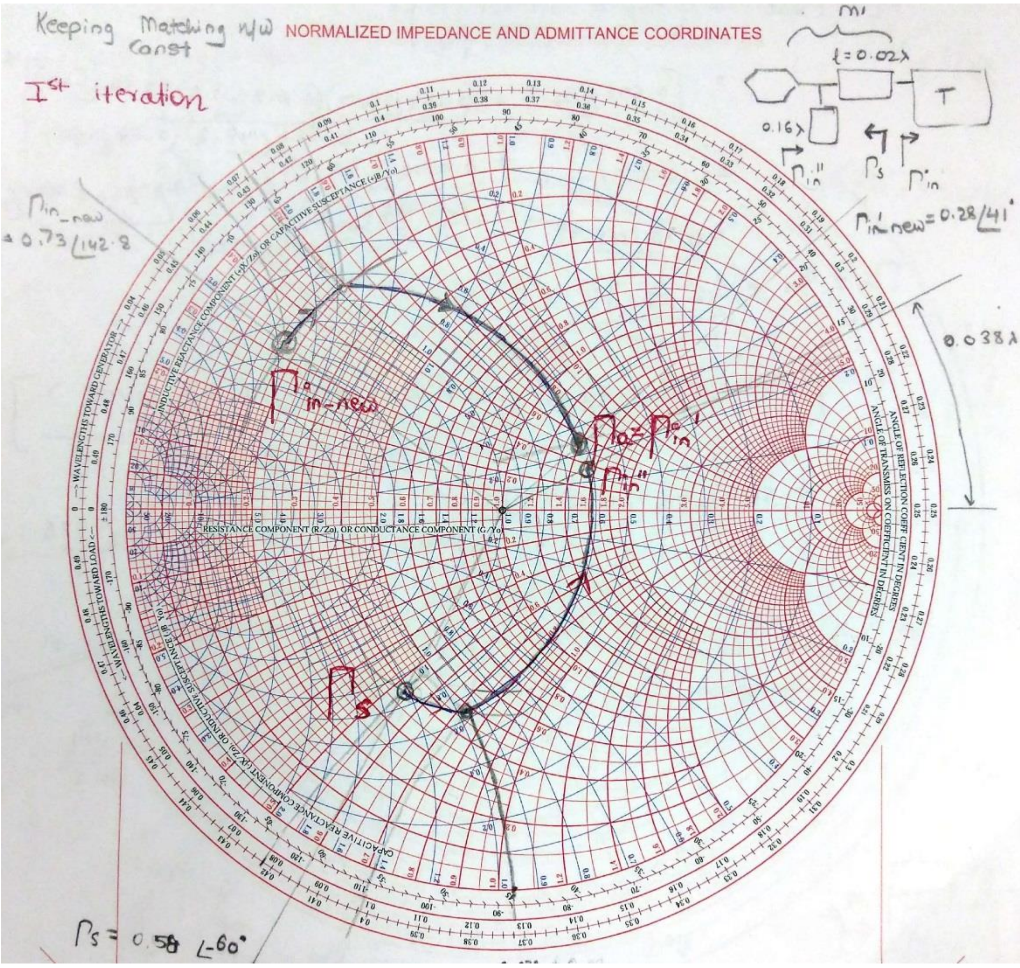


Figure 5.4: First iteration to calculate Γ_a, Noise Figure

Using this value, we can calculate Γ_L.

$$\Gamma_L = \Gamma_{out}^* = \left(S_{22} + \frac{S_{12}.S_{21}.\Gamma_S}{1 - S_{11}.\Gamma_S}\right)^*$$

$$= \left[(0.677\,\angle 172.6°) + \frac{(0.058\,\angle -12.8°)(2.244\,\angle -50.7°)\,.\,(0.58\angle -143.3)}{1 - (0.564\angle 148.8°)\,.\,(0.58\angle -143.3)}\right]^*$$

$$= 0.784\angle -170.26$$

$$\Gamma_{IN} = \left(S_{11} + \frac{S_{12}.S_{21}.\Gamma_L}{1 - S_{22}.\Gamma_L}\right) = 0.78\,\angle 143°$$

Using, $F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1\,.\,G_2}$

F_1, F_2 and F_3 can be calculated since they are the same stages.

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \,.\, G_2} => 1.259 = F_1 + \frac{F_1 - 1}{22.31} + \frac{F_1 - 1}{(22.31) \,.\, (22.31)}$$

$F_1 = 1.247 = 0.960$ dB

Repeating the noise process again using $F = F_1 = 1.247$,

$$N = \frac{F - F_{\mathbf{min}}}{4 \,.\, r_n} |1 + \Gamma_{opt}|^{\,2} = \frac{1.247 - 1.156}{4\,.(0.04)} |1 + 0.38\angle - 175|^{\,2} = 0.219$$

Center is at, $C_F = \frac{\Gamma_{opt}}{1 + N} = \frac{0.38\angle - 175}{1 + 0.219} = 0.312\angle - 175$

$$R_F = \frac{\sqrt{N^2 + N\,(1 - |\Gamma_{opt}|^{\,2})}}{1 + N} = \frac{\sqrt{0.219^2 + 0.219\,(1 - |0.38|^{\,2})}}{1 + 0.219} = \frac{0.485}{1.219}$$
$$= 0.398$$

Again, from the Smith Chart we choose Γ_S closest to Γ_{MS}. Hence we get $\Gamma_S = 0.62\angle - 143.3$

Using this value, we can calculate Γ_L again.

$$\Gamma_L = \Gamma_{out}{}^{*} = \left(S_{22} + \frac{S_{12}\,.S_{21}\,.\Gamma_S}{1 - S_{11}\,.\Gamma_S}\right)^{*}$$
$$= \left[(0.677\,\angle 172.6°) + \frac{(0.058\,\angle - 12.8°)(2.244\,\angle - 50.7°)\,.(0.62\angle - 143.3)}{1 - (0.564\angle 148.8°)\,.(0.62\angle - 143.3)}\right]^{*}$$

$$= 0.795\angle - 170.1$$

$$\Gamma_{IN} = \left(S_{11} + \frac{S_{12}\,.S_{21}\,.\Gamma_L}{1 - S_{22}\,.\Gamma_L}\right) = 0.732\,\angle 142.8°$$

As we can see, the values of Γ_S and Γ_L do not vary that much. Hence we can conclude to proceed with the first value to calculate the gain for the LNA.

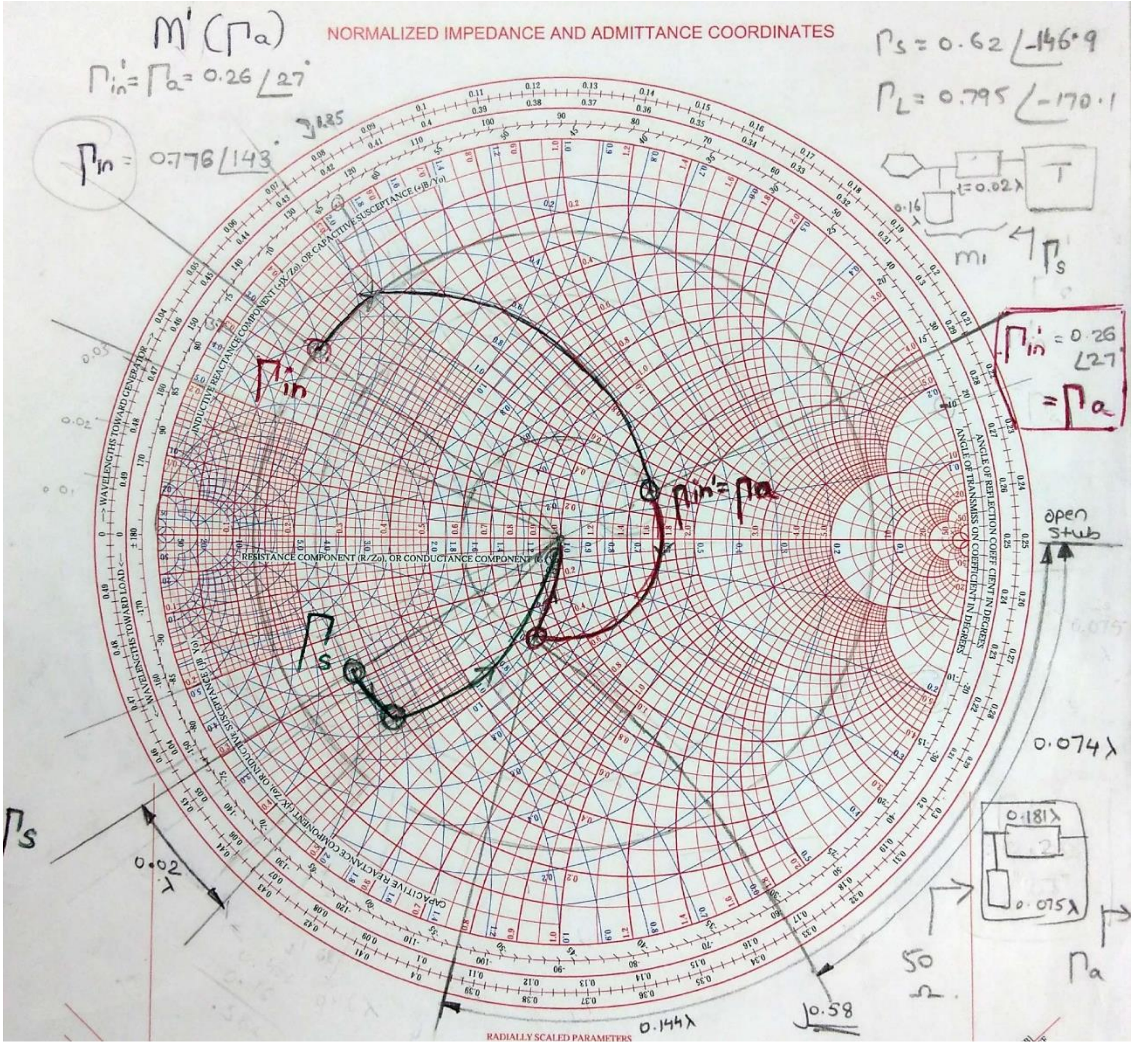


Figure 5.5: Second iteration to calculate Γ_a, Noise Figure

Using Gain of an LNA (calculation for LNA GAIN shown later), we can calculate F_1 again,

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \,.\, G_2}$$

we can calculate F_1, F_2 and F_3 since they are the same stages.

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \,.\, G_2} => 1.259 \;=\; F_1 + \frac{F_1 - 1}{19.68} + \frac{F_1 - 1}{(19.68)\,.\,(19.68)}$$

$F_1 = 1.2458 = 0.954$ dB

By repeating a third iteration for the above process, the new values came out to be,

Noise Parameter N = 0.199

Center C_F = 0.316

Radius R_F = 0.382

Using these values, the value for Γ_S and Γ_L came almost equal to the values that were obtained from the above calculation which concluded the iterations as well.

$$F_{MGA} = F_{\mathbf{min}} + \frac{\frac{4R_N}{Z_0}(\ |\Gamma_S - \Gamma_{opt}|^2)}{(1-|\Gamma_S|^2).|1+\Gamma_{opt}|^2} = \frac{4.\ (0.04)(\ |(0.832\angle-143.3)-(0.38\angle-175)|^2)}{(1-|(0.832|^2).(1+0.38\angle-175^2)} + 1.156$$

=1.272 => 10log1.272 = 1.04 dB

Noise Figure of Cascaded System,

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 .\ G_2} = 1.2458 + \frac{1.2458 - 1}{19.68} + \frac{1.272 - 1}{(19.68).\ (19.68)}$$

$$= 1.2458 + 0.0124 + 0.000702 = 1.2589 => 10\log 1.2598$$

$$= 0.9999 \text{ dB}$$

5.7 Gain Calculations

Ideal Maximum Gain Calculation

$$G = \frac{|S_{21}|}{|S_{12}|} = \frac{|2.244|}{|0.058|} = 38.689 \Rightarrow 10\log_{10} 45.534 = 15.87 \text{ dB}$$

LNA Gain Calculation

$$G_{\mathbf{T}} = G_{\mathbf{S}} . G_{\mathbf{0}} . G_{\mathbf{L}} = \left(\frac{1 - |\Gamma_{\mathbf{S}}|^2}{|1 - S_{\mathbf{11}}\ \Gamma_{\mathbf{S}}|^2}\right) . |S_{\mathbf{21}}|^2 . \left(\frac{1}{|1 - \Gamma_{\mathbf{out}}|^2}\right) =$$

$$\frac{1-(0.62)^2}{|1-(0.564)(0.62)|^2} . (2.244)^2 . \frac{1}{|1-(0.886)|^2} = 19.68 => 10\log 19.68 = 12.94 \text{ dB}$$

MGA Gain Calculation

$$G_{T,MAX} = G_{S,MAX} \cdot G_0 \cdot G_{L,MAX} = \frac{1}{1-|\Gamma_{MS}|^2} \cdot |S_{21}|^2 \cdot \frac{1-|\Gamma_{ML}|^2}{|1-S_{22}\,\Gamma_{ML}|^2} =$$

$$= \frac{1}{1-|0.832|^2} \cdot (2.244)^2 \cdot \frac{1-|0.886|^2}{|1-((0.677\angle 172.6°)\,(0.886\angle -169.2°))|^2} = 22.31 = 13.48 \text{ dB}$$

Gain of Cascaded System

LNA (dB) + LNA (dB) + MGA (dB) = 12.94 + 12.94 + 13.48 = 39.39 dB

Note: All the above results are verified on various software's. Below are the screen shots from the RF/MW E-Book Software* provided by the professor himself which is also available with his book "Advanced RF & Microwave Circuit Design". In these images, we can verify the accuracy of our hand calculations.

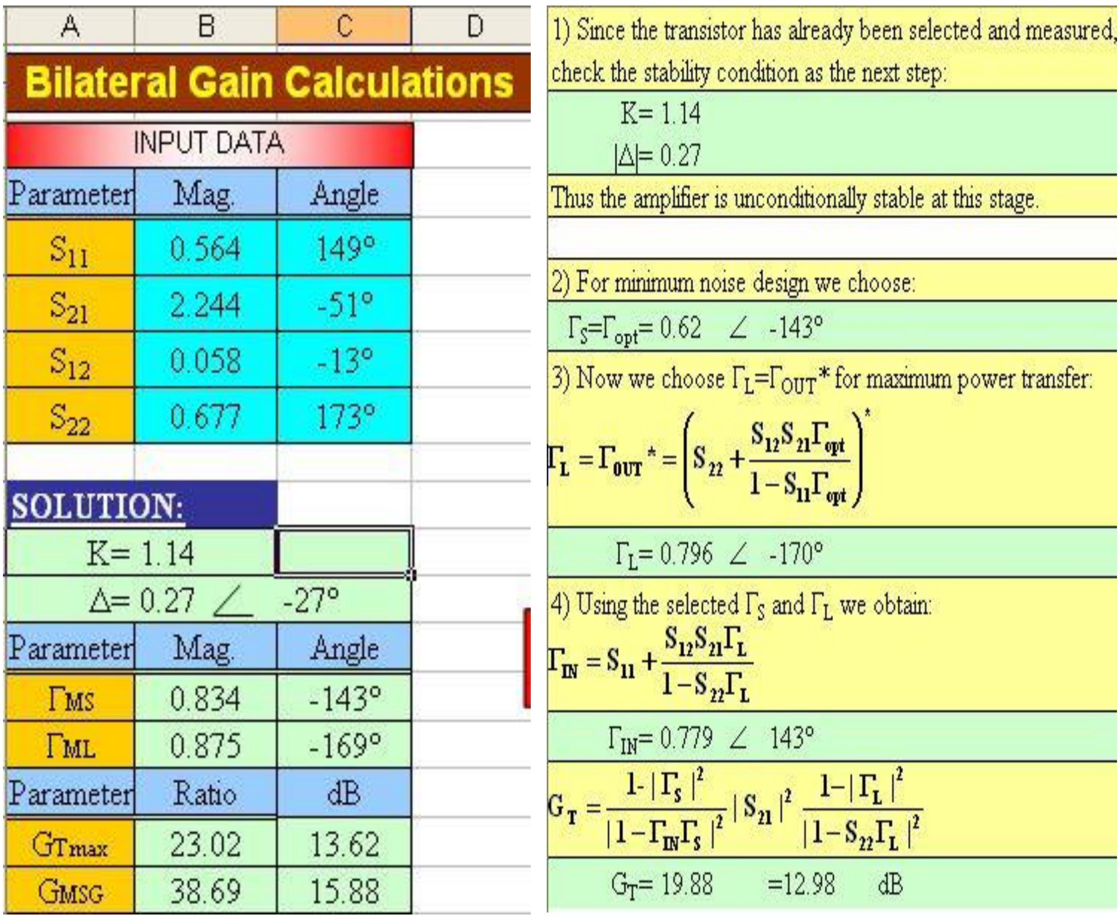


Figure 5.6: Result for Gain, Γ_{MS}, Γ_{ML} and Γ_{IN}

Quasi-TEM Transmission Line (microstrip)

INPUT DATA

Parameter	Value	Unit
f	16	Ghz
w	0.1136	mm
h	0.127	cm
ε_r	10.7	

SOLUTION:

OUTPUT DATA

Parameter	Value	Unit
ε_{ff}	7.13	
λ_o	1.875	cm
λ_{TEM}	0.57	cm
λ	0.702	cm
Vp	4.2E+09	cm/sec
fo	3.37	GHz

Figure 5.7: Result for transmission length

After comparing the above results with the software, it can be seen that the results match. Hence, it can be concluded that our hand calculations are correct. The same has been verified in the MATLAB and is mentioned in Appendix's A, B and C.

Chapter 6. Smith Charts (by Hand Calculation and Simulation)

6.1 Introduction

This chapter deals with the matching networks through Smith Chart. It also utilizes a software verification to check the accuracy of the Smith Charts done by hand.

We are using distributed elements in our design due to high frequency and as described in chapter 4, the steps help us to understand how we proceed in order to design our matching network.

After performing the design using both the open as well as short stubs for distributed elements, it was observed that short stubs gave a gain which was less when compared to the gain obtained by the open stubs primarily because of the long stub lengths. Though short stubs are preferred in real life as they do not allow any external interferences, it could not be used due to the reason mentioned above.

6.2 LNA - Using Unilateral Case

A) $\Gamma_S = 0.62\angle -143.3°$

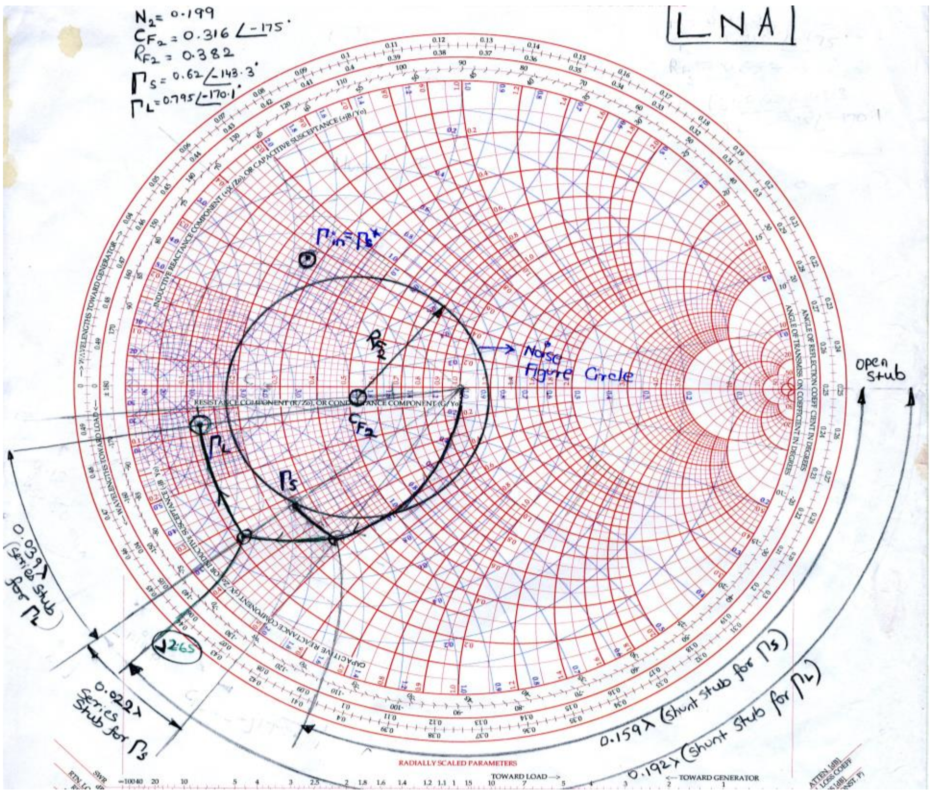


Figure 6.1: Smith Chart for matching network of LNA

This is the LNA matching network. M_1 is at the input side of the LNA transistor. Similarly, M_2 is at the output side. Below is the process to design an M_1 and M_2 respectively.

For the design of matching network, we plot Γ_S followed by Γ_L and move towards the center which has a load of 50 Ω.

The design process for both open and short stub is same except that they both lead to different stub lengths since they are measured from different points on the Smith Chart.

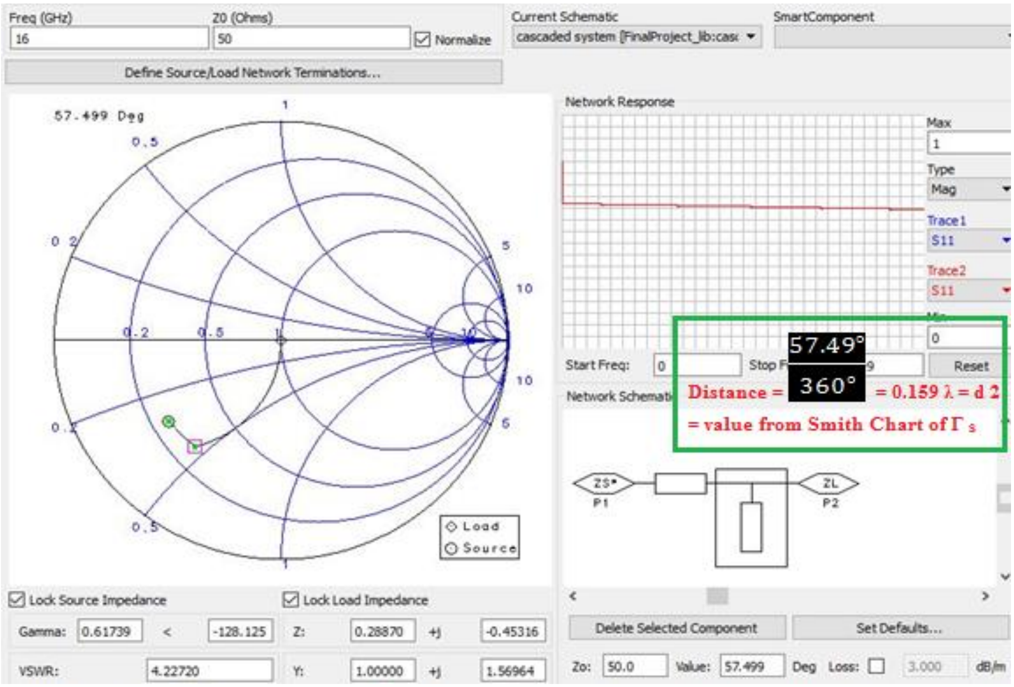


Figure 6.2: Simulation result for shunt stub of Γ_S for LNA

Here, we can see that the values from the Smith Chart and that from the simulation are almost equal. The calculation for the stub length on the software is mentioned in the image itself where the angle is divided by 360°. This division gives the length in terms of λ. By substituting the value of λ, we get the final answer in either mm, cm or m depending on the units of the λ. The same process is carried out for other matching networks and is shown ahead.

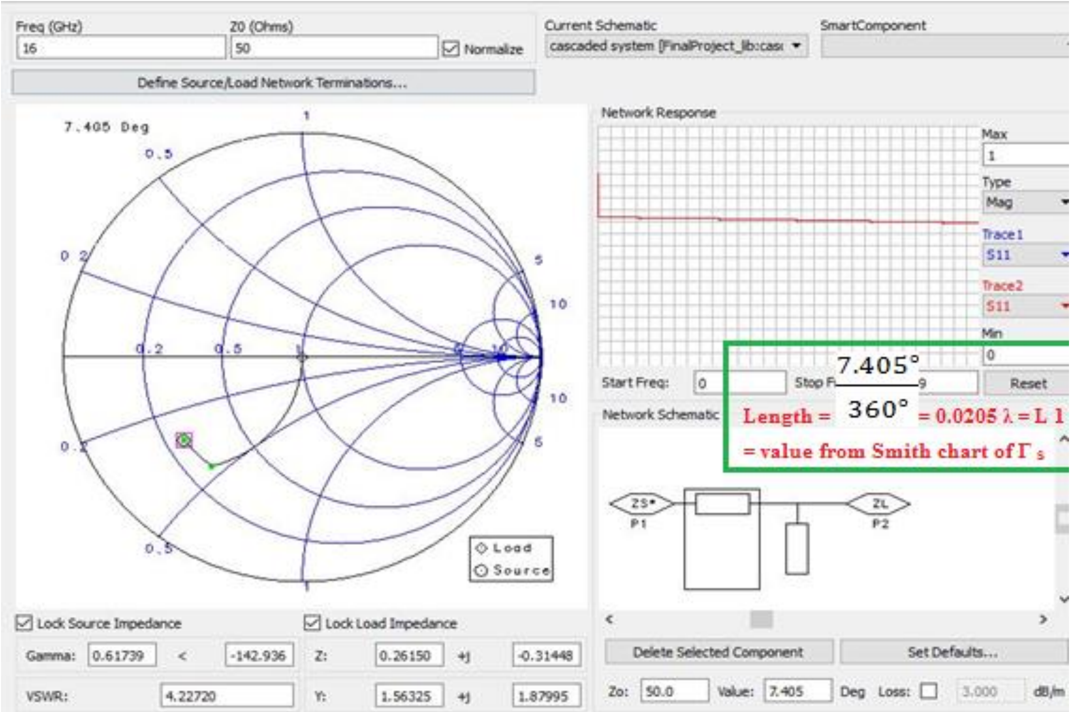


Figure 6.3: Simulation result for series stub of Γ_S for LNA

B) $\Gamma_L = 0.795 \angle -170.1$

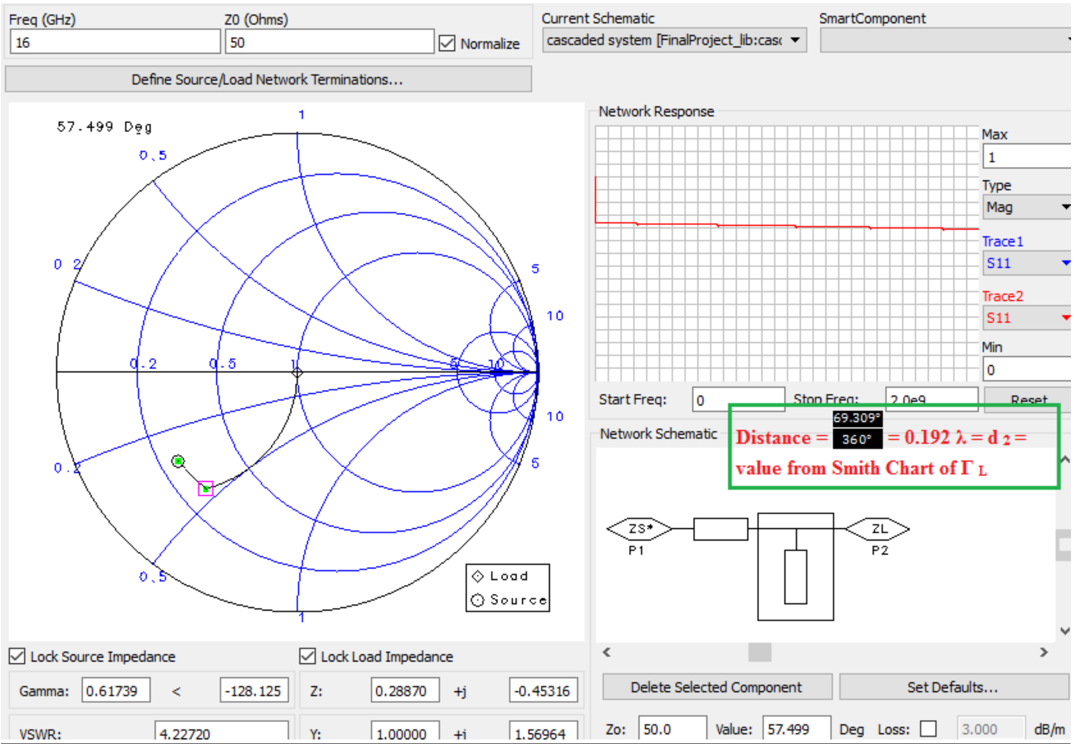


Figure 6.4: Simulation result for shunt stub of Γ_L for LNA

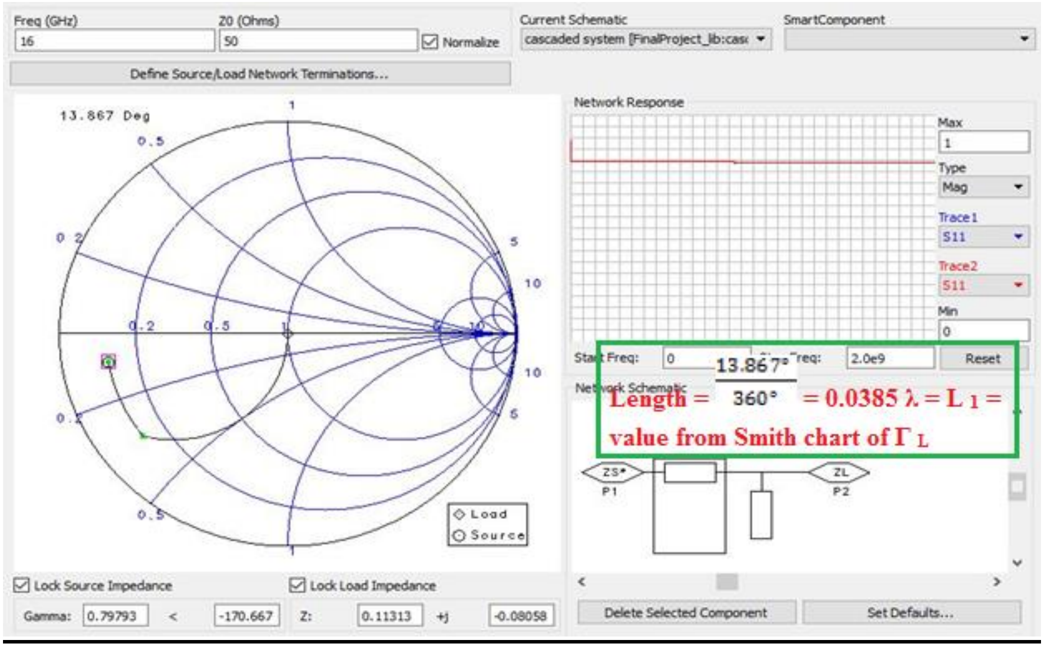


Figure 6.5: Simulation result for series stub of Γ_L for LNA

6.3 MGA - Using Bilateral Case

A) $\Gamma_{MS} = 0.815 \angle -143.3$

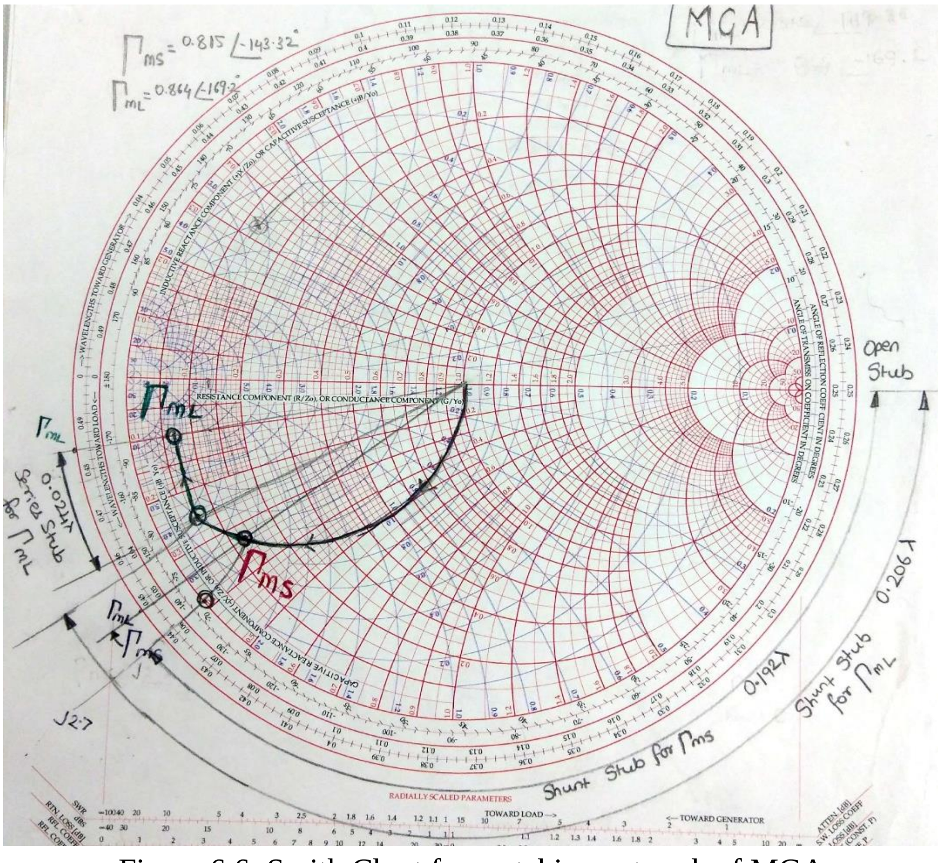


Figure 6.6: Smith Chart for matching network of MGA

The above Smith Chart is for matching networks M_1 and M_2 for a Maximum Gain Amplifier (MGA).

The figure above for the Smith Chart has both the matching networks for MGA, shown in it along with their stub length.

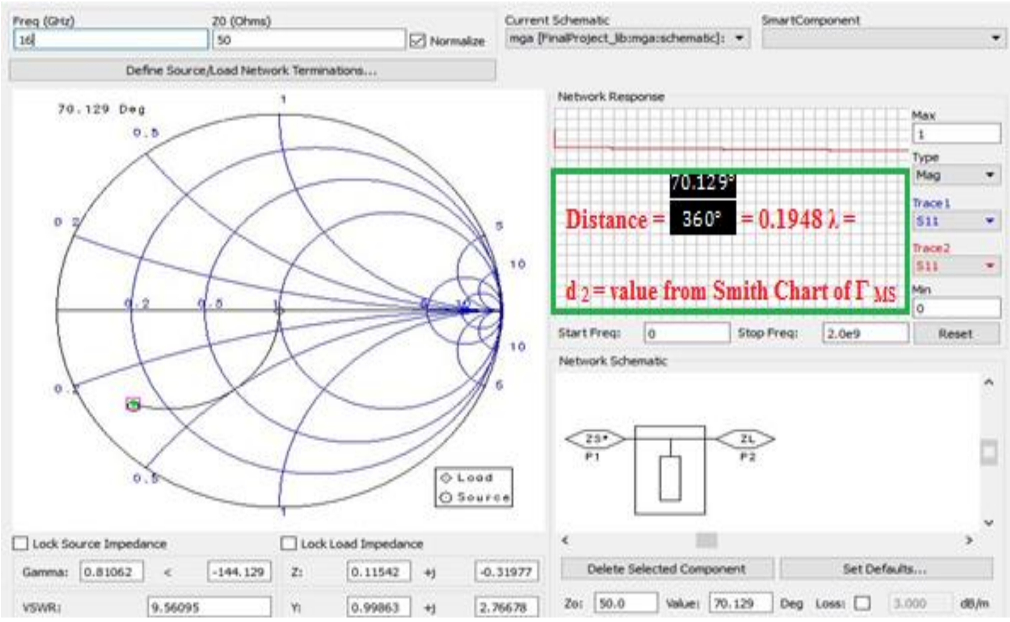


Figure 6.7: Simulation result for shunt stub of Γ_{MS} for MGA

B) $\Gamma_{ML} = 0.864 \angle -169.2$

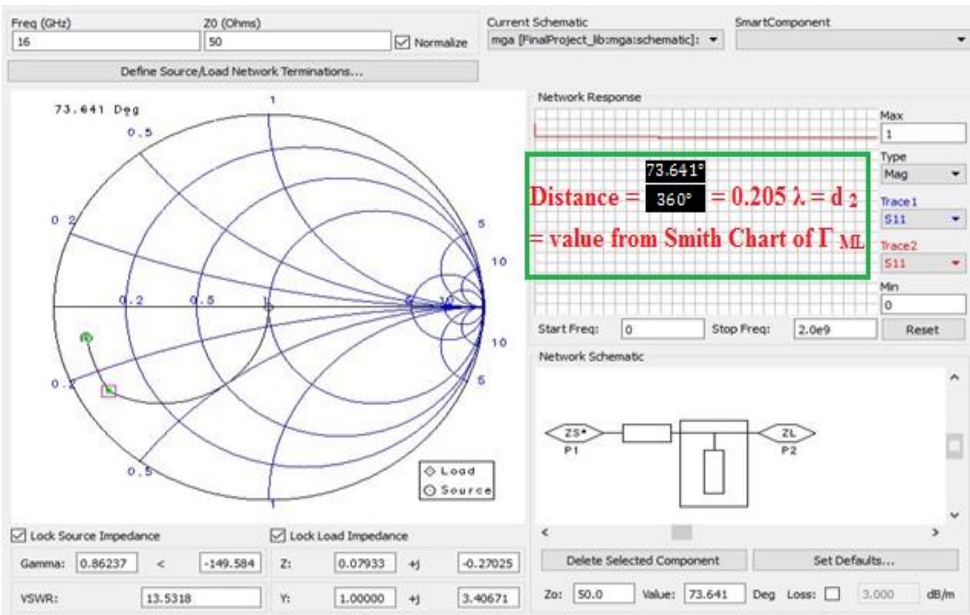


Figure 6.8: Simulation result for shunt stub of Γ_{ML} for MGA

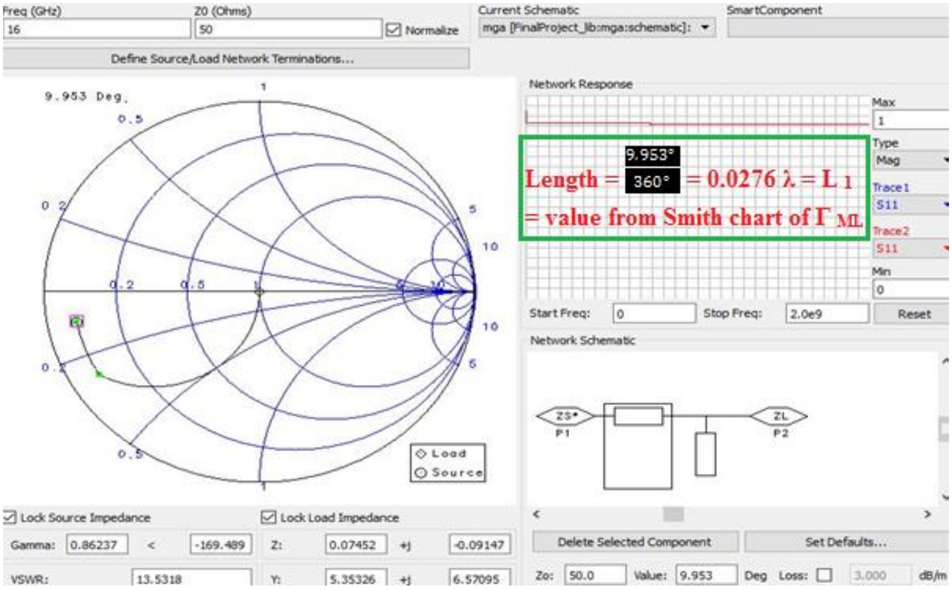


Figure 6.9: Simulation result for series stub of Γ_{ML} for MGA

The three figures above, Figure 6.7, 6.8 and 6.9 show the software verification result for a series and shunt stubs of matching networks M_1 and M_2 respectively for an MGA.

6.4 M-Prime (M ' Network)

M prime is the matching network to overcome the mismatch between the networks in order to improve the efficiency of the circuit. It is used between LNA's to overcome the mismatch factor.

$\Gamma_S = 0.62 \angle -146.9$ and $\Gamma_L = 0.795 \angle -170.1$ leading to

$\Gamma_{in}' = \Gamma_a = 0.26 \angle 27$

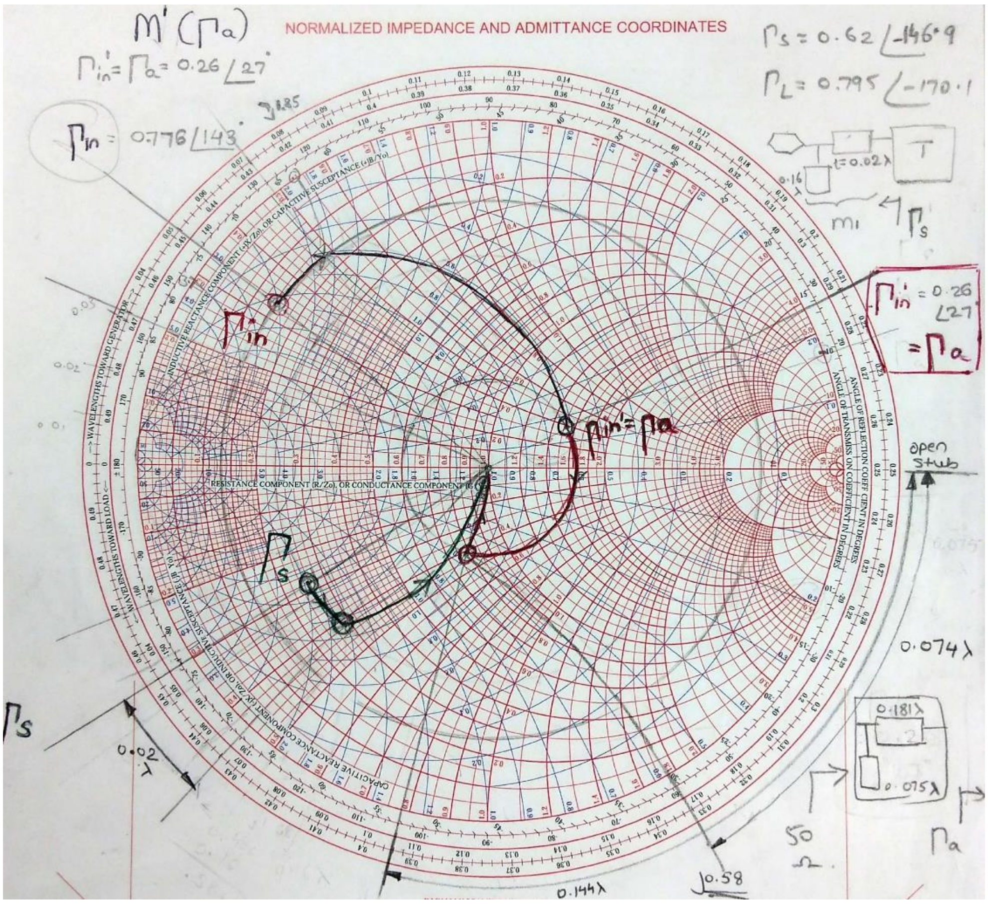


Figure 6.10: Smith Chart for matching network of M prime

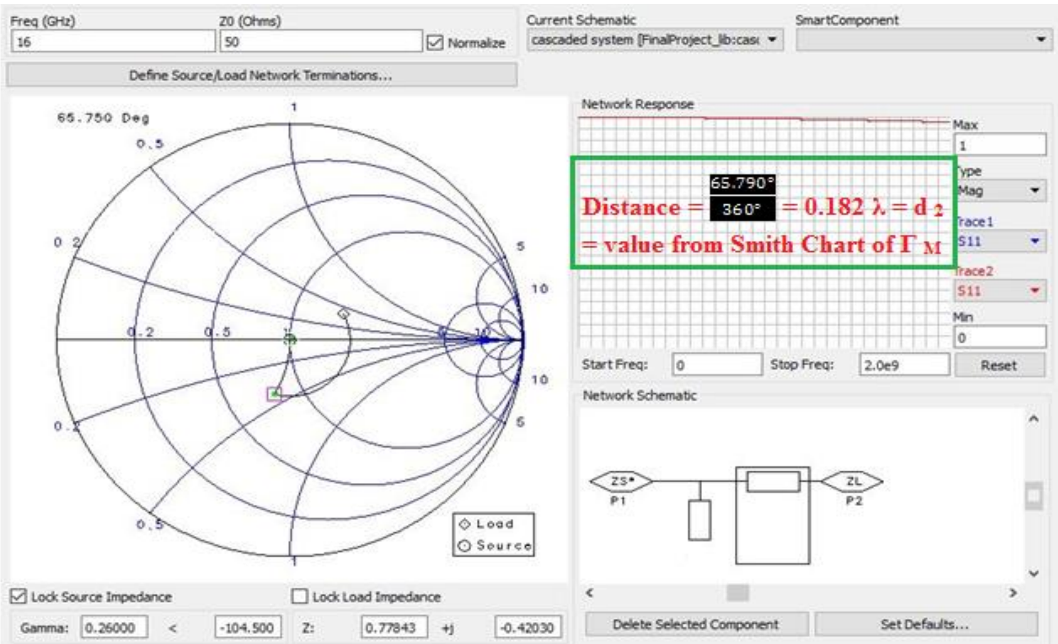


Figure 6.11: Simulation result for series stub for M Prime

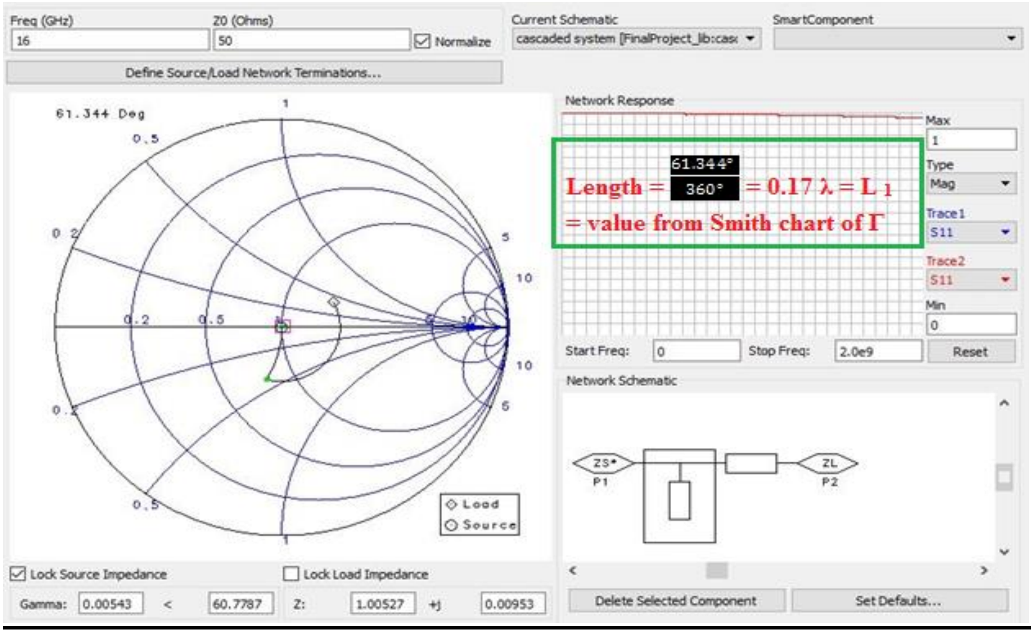


Figure 6.12: Simulation result for shunt stub for M Prime

6.5 Combining LNA-M_2 to M ’, also combining LNA-M_2 to MGA-M $_1$

Ideally, when there are multiple shunt stubs, it is not a good practice to keep shunt stubs separate. If there are consecutive shunt stubs, they can be combined to make a single shunt stub. The Smith Chart below shows two such combinations for two different shunt stubs.

In this Smith Chart, we have combined the shunt stub of matching network M_2 of the LNA with the shunt stub of M ’.

Also, in the same Chart we have combined the shunt stub of matching network M_2 of the LNA with the shunt stub of matching network M_1 of the MGA. In order to combine the shunt stubs, we need to calculate the equivalent susceptance value of the stub from the Smith Chart, add them together and find the equivalent open stub length of this calculated susceptance value.

$\Gamma_{in} = 0.776\ \angle 143$, $\Gamma_S = 0.62\ \angle -146.9$, $\Gamma_L = 0.795\ \angle -170.1$

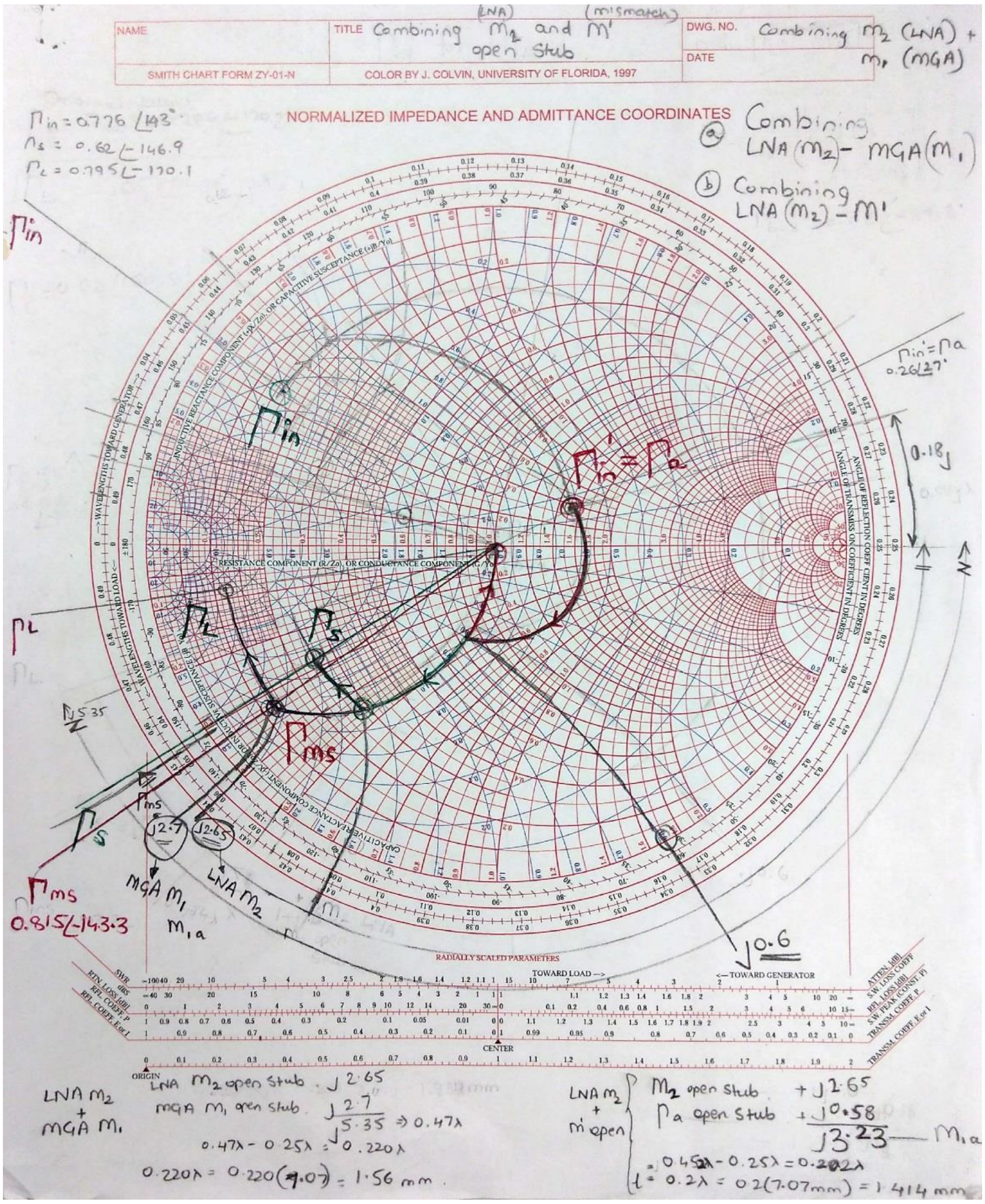


Figure 6.13: Combing the stubs LNA-M2 to M'and LNA-M2 to MGA-M 1

Combining **LNA-M2 and MGA-M1** (All are Open stubs)

LNA-M 2 + MGA-M 1 = j 2.65 + j 2.7 = j 5.35 => 0.47 λ

For open stub, 0.47 λ – 0.25 λ = 0.22 λ = 0.22 (0.707 mm) = 1.56 mm

Combining **LNA-M2 and M '** (All are Open stubs)

LNA-M 2 + MGA-M 1 = j 2.65 + j 0.6 = j 3.23 => 0.45 λ

For open stub, 0.45 λ – 0.25 λ = 0.202 λ = 0.22 (0.707 mm) = 1.414 mm

Chapter 7. Micro-strip (w by h) Calculations

7.1 For Zo = 50Ω

Frequency = 16 GHz, ε $_r$ = 10.7 (from Rodgers Data Sheet), h = 0.127 mm

Approximate Method:

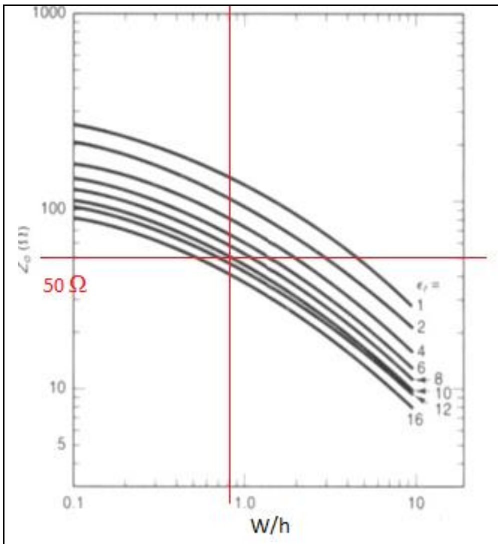


Figure 7.1: Plots of Zo vs. W/h with ε $_r$ as a parameter (Zo = 50 Ω).

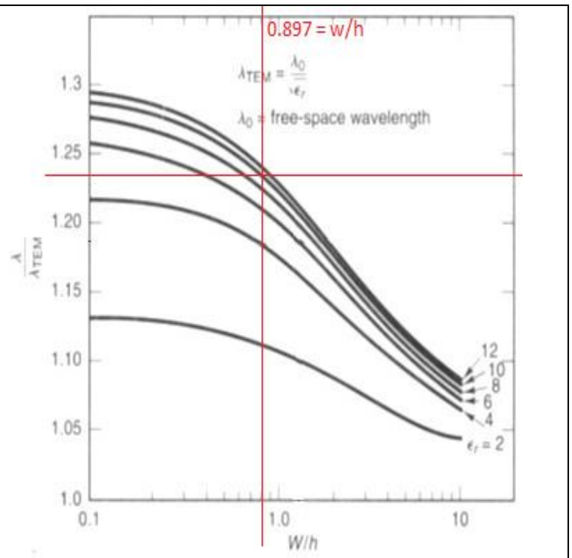


Figure 7.2: Plots of Zo vs. W/h with ε_r as a parameter (Zo = 50 Ω).

Formulas used in Micro-strip calculation - Exact Method:

W/h Formula- Assuming (ε $_{ff}$) and Zo are given, then the micro-strip dimensions (W/h) can be found as follows:

For W/h ≤ 2:

$$\frac{W}{h} = \frac{8e^{A}}{e^{2A} - 2}$$

For W/h ≥ 2:

$$\frac{W}{h} = \frac{2}{\pi}\left[B - 1 - \ln(2B-1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left\{\ln(B-1) + 0.39 - \frac{0.61}{\varepsilon_r}\right\}\right]$$

Where

$$A = \frac{Z_O}{60}\sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right) \quad \text{and} \quad B = \frac{377\pi}{2Z_O\sqrt{\varepsilon_r}}$$

For w / h < 2

$$A = \frac{Z_0}{60}\left(\frac{\varepsilon_r + 1}{2}\right)^{1/2} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right)$$

$$= \frac{50}{60}\left(\frac{10.7 + 1}{2}\right)^{1/2} + \frac{10.7 - 1}{10.7 + 1}\left(0.23 + \frac{0.11}{10.7}\right)$$

$$= 0.833\ (2.418) + 0.829\ (0.23 + 0.0102) = \mathbf{2.2133}$$

$$\frac{w}{h} = \frac{8 * e^{A}}{e^{2A} - 2} = \frac{8 * e^{2.2133}}{e^{2\,(2.2133)} - 2} = \frac{8 * 9.136}{83.46 - 2} = \mathbf{0.897}$$

From Rodger's sheet, we have considered h = 0.127 mm

This leads to a value of w = 0.1136 mm

Effective dielectric constant –

The effective dielectric constant (ε_{ff}) is given by [assuming that the dimensions of the micro-strip line (W, h) are known]:

For W/h ≤ 1

$$\varepsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}} + 0.04\left(1 - \frac{W}{h}\right)^2\right]$$

For W/h ≥ 1

$$\varepsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}}\right]$$

By using the equations provided, we can find $\varepsilon_{\mathbf{eff}}$ and length

$$\varepsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}} + 0.04\left(1 - \frac{W}{h}\right)^2\right] = \frac{10.7+1}{2} + \frac{10.7-1}{2}\left[\left(1 + 12\left(\frac{1}{0.897}\right)\right)^{-\frac{1}{2}} + 0.04(1 - 0.897)^2\right] = 5.85 + 4.85[0.2634 + 4.41 \text{ x } 10^{-4}] = 7.130$$

λ Formula-

The wavelength in the micro-strip line (λ) is given by:

Speed of light, $c = 3 \,.\, 10^{8}$ m/s = $3 \,.\, 10^{11}$ mm/s, and frequency f = 16 GHz

For W/h < 0.6:

$$\boldsymbol{\lambda} = \frac{\boldsymbol{\lambda}_O}{\sqrt{\boldsymbol{\varepsilon}_r}}\left[\frac{\boldsymbol{\varepsilon}_r}{1+0.6(\boldsymbol{\varepsilon}_r-1)(W/h)^{0.0297}}\right]^{1/2}$$

For W/h ≥ 0.6:

$$\boldsymbol{\lambda} = \frac{\boldsymbol{\lambda}_O}{\sqrt{\boldsymbol{\varepsilon}_r}}\left[\frac{\boldsymbol{\varepsilon}_r}{1+0.63(\boldsymbol{\varepsilon}_r-1)(W/h)^{0.1255}}\right]^{1/2}$$

$$\lambda_0 = \frac{c}{f} = \frac{3\,.\,10^{11}}{16\,.\,10^{9}} = 18.75 \text{ mm}$$

Using w/h ≥ 0.6 formula, we get

$$\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}}\left[\left(\frac{\varepsilon_r}{1+0.63\,(\varepsilon_r-1)(\frac{W}{h})^{0.1255}}\right)^{\frac{1}{2}}\right]$$

$$= \frac{18.75}{\sqrt{10.7}}\left[\left(\frac{10.7}{1+0.63\,(10.7-1)(0.897)^{0.1255}}\right)^{\frac{1}{2}}\right] = 7.07 \text{ mm}$$

$$\lambda_{TEM} = \frac{\lambda_0}{\sqrt{\varepsilon_r}} = \frac{18.75}{\sqrt{10.7}} = 5.73 \text{ mm}$$

$$\frac{\lambda}{\lambda_{TEM}} = \frac{7.07}{5.73} = 1.23\,,$$

Also, by graph, $\frac{\lambda}{\lambda_{TEM}} = 1.23$ approx.

7.2 For Zo = 120 Ω

Frequency = 16 GHz, ε_r = 10.7 (from Rodgers Data Sheet), h = 0.127 mm

Approximate Method:

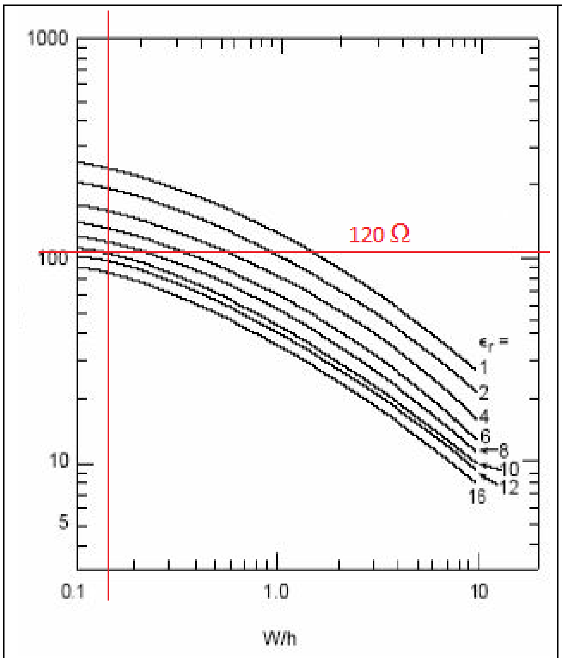


Figure 7.3: Plot of Zo vs. W/h with ε_r as a parameter (Zo = 120 Ω).

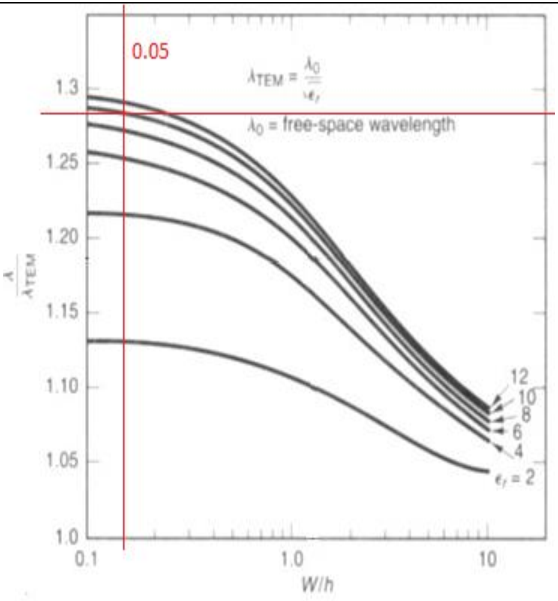


Figure 7.4: Plots of Zo vs. W/h with εr as a parameter (Zo = 120 Ω).

Formulas used in Micro-strip calculation - Exact Method:

W/h Formula-

Assuming (ε_{ff}) and Zo are given, then the micro-strip dimensions (W/h) can be found as follows:

For W/h ≤ 2:

$$\frac{W}{h} = \frac{8e^{A}}{e^{2A} - 2}$$

For W/h ≥ 2:

$$\frac{W}{h} = \frac{2}{\pi}\left[B - 1 - \ln(2B - 1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left\{\ln(B - 1) + 0.39 - \frac{0.61}{\varepsilon_r}\right\}\right]$$

Where

$$A = \frac{Z_O}{60}\sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right) \quad \text{and} \quad B = \frac{377\pi}{2Z_O\sqrt{\varepsilon_r}}$$

For w / h < 2

$$A = \frac{Z_0}{60}\left(\frac{\varepsilon_r + 1}{2}\right)^{1/2} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right)$$

$$= \frac{120}{60}\left(\frac{10.7 + 1}{2}\right)^{1/2} + \frac{10.7 - 1}{10.7 + 1}\left(0.23 + \frac{0.11}{10.7}\right)$$

$$= 2\,(2.418) + 0.829\,(0.23 + 0.0102) = \mathbf{5.03}$$

$$\frac{w}{h} = \frac{8 * e^{A}}{e^{2A} - 2} = \frac{8 * e^{5.03}}{e^{2\,(5.03)} - 2} = \frac{8 * 152.9}{23388.5 - 2} = 0.052$$

From Rodger's sheet, we have considered h = 0.127 mm

This leads to a value of w = 0.0066 mm

Effective dielectric constant –

The effective dielectric constant (ε_{ff}) is given by [assuming that the dimensions of the micro-strip line (W, h) are known]:

For W/h ≤ 1

$$\varepsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}} + 0.04\left(1 - \frac{W}{h}\right)^2\right]$$

For W/h ≥ 1

$$\varepsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}}\right]$$

By using the equations provided, we can find $\varepsilon_{\mathbf{eff}}$ and length

$$\epsilon_{ff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{h}{w}\right)\right)^{-\frac{1}{2}} + 0.04\left(1 - \frac{W}{h}\right)^2\right] = \frac{10.7+1}{2} + \frac{10.7-1}{2}\left[\left(1 + 12\left(\frac{1}{0.052}\right)\right)^{-\frac{1}{2}} + 0.04(1 - 0.052)^2\right] = 5.85 + 4.85[0.066 + 0.036] = 6.34$$

λ Formula-

The wavelength in the micro-strip line (λ) is given by:

Speed of light, c = 3 . 10^8 m/s = 3 . 10^{11} mm/s, and frequency f = 16 GHz

For W/h < 0.6:

$$\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}}\left[\frac{\varepsilon_r}{1+0.6(\varepsilon_r-1)(W/h)^{0.0297}}\right]^{1/2}$$

For W/h ≥ 0.6:

$$\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}}\left[\frac{\varepsilon_r}{1+0.63(\varepsilon_r-1)(W/h)^{0.1255}}\right]^{1/2}$$

$$\lambda_0 = \frac{c}{f} = \frac{3\,.\,10^{11}}{16\,.\,10^{9}} = 18.75 \text{ mm}$$

Using w/h ≤ 0.6 formula, we get

$$\lambda = \frac{\lambda_0}{\sqrt{\varepsilon_r}}\left[\left(\frac{\varepsilon_r}{1+0.6\,(\varepsilon_r-1)(\frac{W}{h})^{0.0297}}\right)^{\frac{1}{2}}\right]$$

$$= \frac{18.75}{\sqrt{10.7}}\left[\left(\frac{10.7}{1+0.6\,(10.7-1)(0.052)^{\,0.0297}}\right)^{\frac{1}{2}}\right] = 7.45 \text{ mm}$$

$$\lambda_{TEM} = \frac{\lambda_0}{\sqrt{\varepsilon_r}} = \frac{18.75}{\sqrt{10.7}} = 5.73 \text{ mm}$$

$$\frac{\lambda}{\lambda_{TEM}} = \frac{7.45}{5.73} = 1.30$$

Also, by graph, $\frac{\lambda}{\lambda_{TEM}} = 1.29$ approx.

Calculation for Gap Capacitance and cutoff frequency

Width (W_1 and W_2) = 0.1136 mm, Cs = 4.97pF, h = 0.127 mm (Appendix E)

$Q_1 = 0.04598\ (0.03 + \left(\frac{W_1}{h}\right)^{Q_5}\ .\ (0.272 + 0.07\ \varepsilon_r)$

$= 0.04598\ (0.03 + \left(\frac{0.1136}{0.127}\right)^{Q_5}\ .\ (0.272 + 0.07(10.7)) = 0.0423$

$$Q_5 = \frac{1.23}{1+0.12\left(\frac{W_2}{W_1}-1\right)^{0.9}} = \frac{1.23}{1+0.12\left(\frac{0.1136}{0.1136}-1\right)^{0.9}} = 1.23$$

$$Cs\ [pF] = 500 \,.\, h \,.\, \exp\left((-1.86)\,.\,\frac{s}{h}\right) Q_1 \,.\, \left(1 + 4.19\left(1 - \exp\left((-0.785)\sqrt{\frac{h}{W_1}}\,.\,\frac{W_2}{W_1}\right)\right)\right)$$

$$\exp\left((-1.86)\,.\,\frac{s}{h}\right) = \frac{Cs\ [pF]}{500\,.h\,.\ Q1\,.(1 + 4.19(1 - \exp\left((-0.785)\sqrt{\frac{h}{W_1}}.\frac{W_2}{W_1}\right)))}$$

$$= \frac{1.99}{500\ (0.127).(0.0423).(1 + 4.19(1 - \exp\left((-0.785)\sqrt{\frac{0.127}{0.1136}}.\frac{0.1136}{0.1136}\right)))}$$

From the above equation, we get the value of Gap Capacitance S equal to, **S = 0.0409 mm**

Cut Off frequency f_0 (GHz) = $(0.3).\sqrt{\frac{Z_0}{h\,.\sqrt{\varepsilon_r - 1}}} = (0.3)\,.\sqrt{\frac{50}{(0.0127)\,.\sqrt{10.7-1}}}$ = 10.66 GHz

In the above equation, height is in cm.

Since cut off frequency is low with respect to the design frequency, now we need to check our impedance Z_0. To calculate Z_0 we use the formula mentioned below.

ε_{ff} = 7.130 (from above calculations), f ' = 16 GHz

$$F = \frac{4\,.h\,.f\,.\sqrt{\varepsilon_r - 1}}{c}\left\{0.5 + \left[1 + 2\log\left(1 + \frac{w}{h}\right)\right]^2\right\} = \frac{4\,.(0.127)\,.\,(16.10^9).\ \sqrt{10.7-1}}{3\,.\,10^{11}}\{0.5 + [1 + 2\log(1 + 0.8948)]^2\} = 0.2463$$

$$\varepsilon_{ff}(f) = \left[\frac{\sqrt{\varepsilon_r} - \sqrt{\varepsilon_{ff}(0)}}{1 + 4\,F^{-1.5}} + \sqrt{\varepsilon_{ff}(0)}\,\right]^2 = \left[\frac{\sqrt{10.7} - \sqrt{7.13}}{1 + 4(0.26)^{-1.5}} + \sqrt{7.13}\right]^2 = 7.225$$

$$Z_0(f) = Z_0(0).\frac{\varepsilon_{ff}(0) - 1}{\varepsilon_{ff}(f) + 1}\sqrt{\frac{\varepsilon_{ff}(0)}{\varepsilon_{ff}(f)}} = 50\,.\left(\frac{7.13 - 1}{7.23 + 1}\right)\sqrt{\frac{7.13}{7.23}} = 50.43\ \Omega$$

The above calculations were again performed at f ' = 16.8GHz which gave Z_0 =50.48Ω

All the above calculations are verified on various softwares and the results are attached in the appendix's respectively.

Chapter 8. Final Circuit Diagram

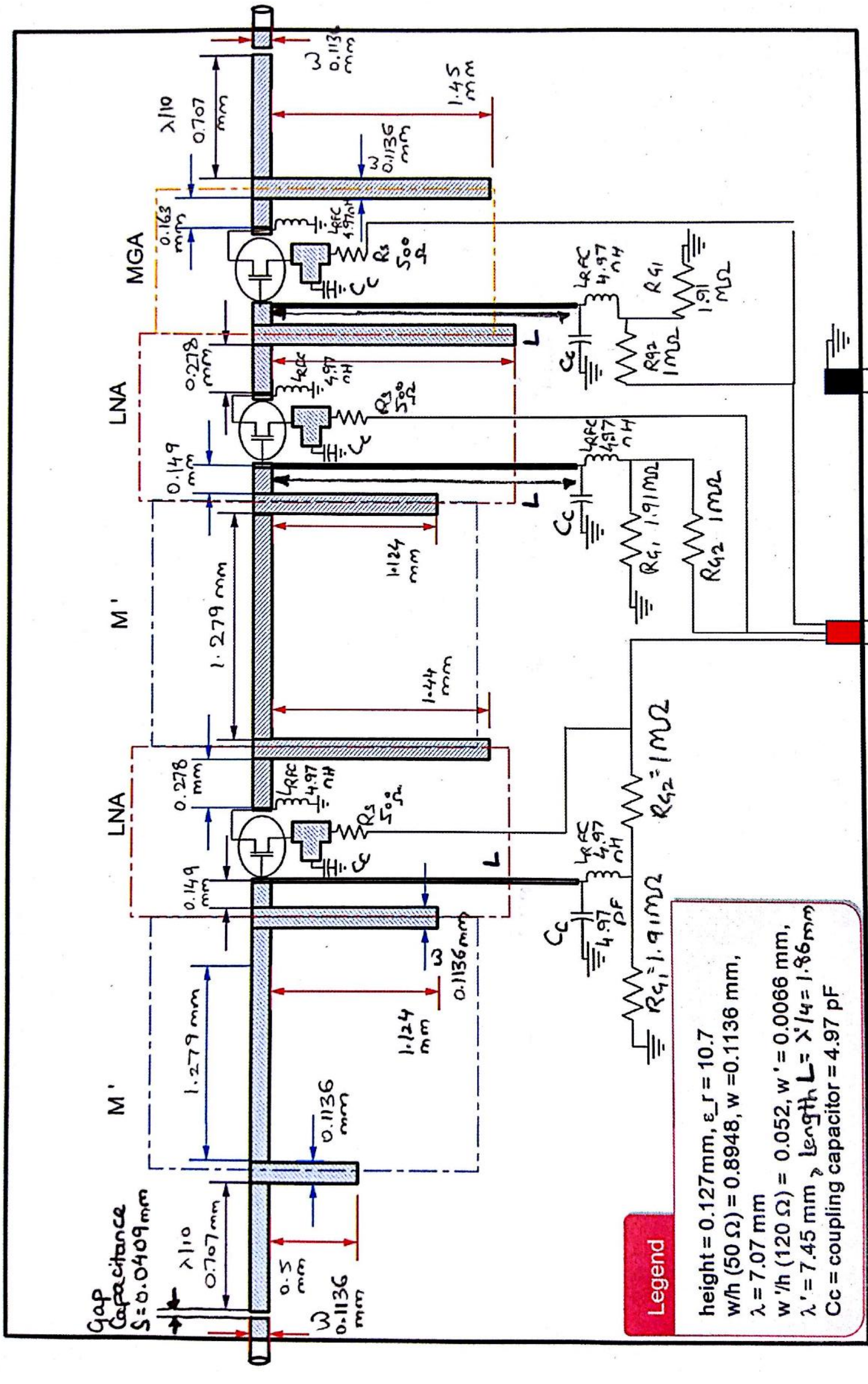


Figure 8.1: Final Layout by combining all the stages

Chapter 9. Circuit and Simulation Results for Single and Multiple Stages

The simulation result for the single stage Low Noise Amplifier confirms that the gain is 12.988 dB and noise figure is approximately 0.865 dB. Similarly, the gain for a single stage MGA is 13.584 dB and the noise figure is around 1.017 dB.

9.1 Single stage LNA result

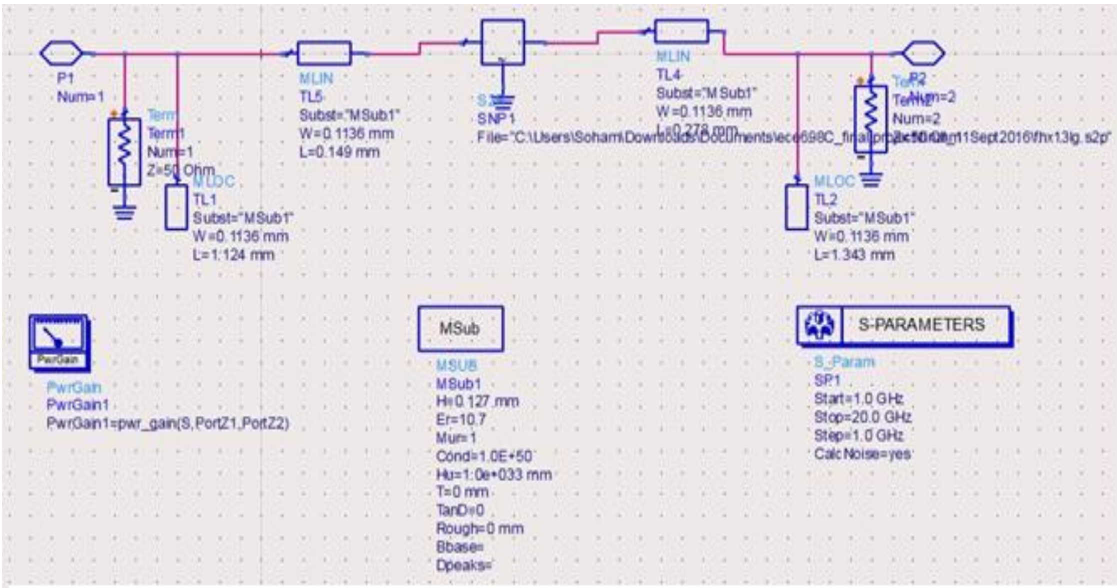


Figure 9.1: Circuit for Single Stage LNA

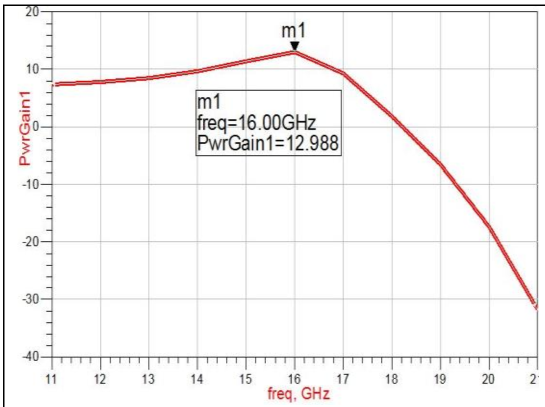


Figure 9.2: Gain for Single Stage LNA

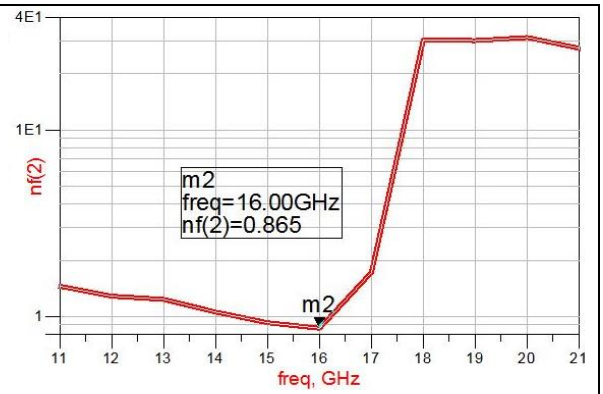


Figure 9.3: Noise Figure for Single Stage LNA

The simulation graphs helps us to verify the performance of our design. ADS was used to perform these simulations. The .S2P file of the transistor is required to load in the block so that it performs like the selected transistor. The calculations of the stub lengths for the respective matching network was discussed previously. Using this, the circuit was simulated to plot the graph for the gain and noise figure.

Similarly, the analysis for single stage MGA was performed followed by the design of the entire system after combining the LNA and MGA to form three stages with the mismatched M Prime (M ’) network and was simulated to obtain the values of gain and noise figure. The same analysis was performed on another simulation tool in order to verify the results and check the accuracy of the design on multiple platforms. Both the software results can be seen below and in appendix F.

9.2 Single stage MGA result

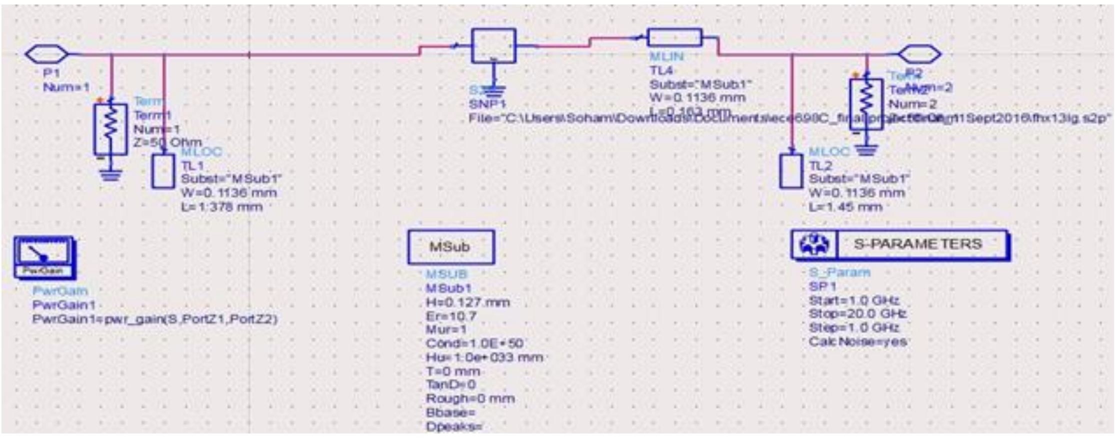


Figure 9.4: Circuit for Single Stage MGA

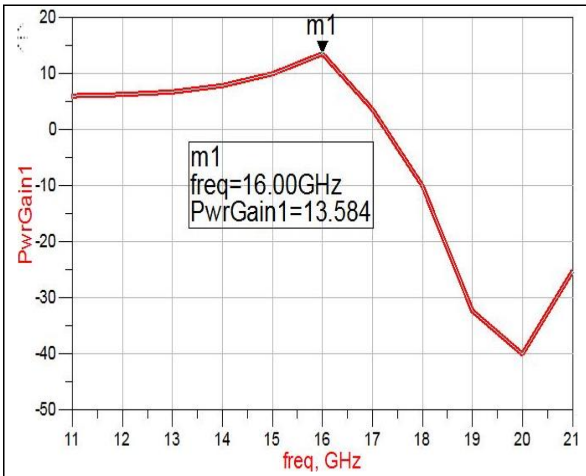


Figure 9.5: Gain for Single Stage MGA

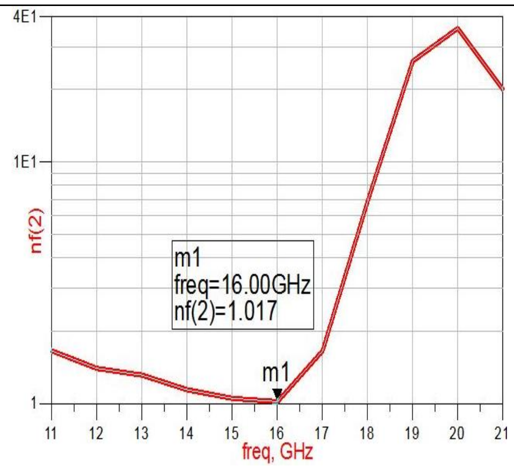


Figure 9.6: Noise Figure for Single Stage MGA

This satisfies the goal to design a multi stage Low Noise Amplifier with low noise figure.

9.3 Cascaded Stages

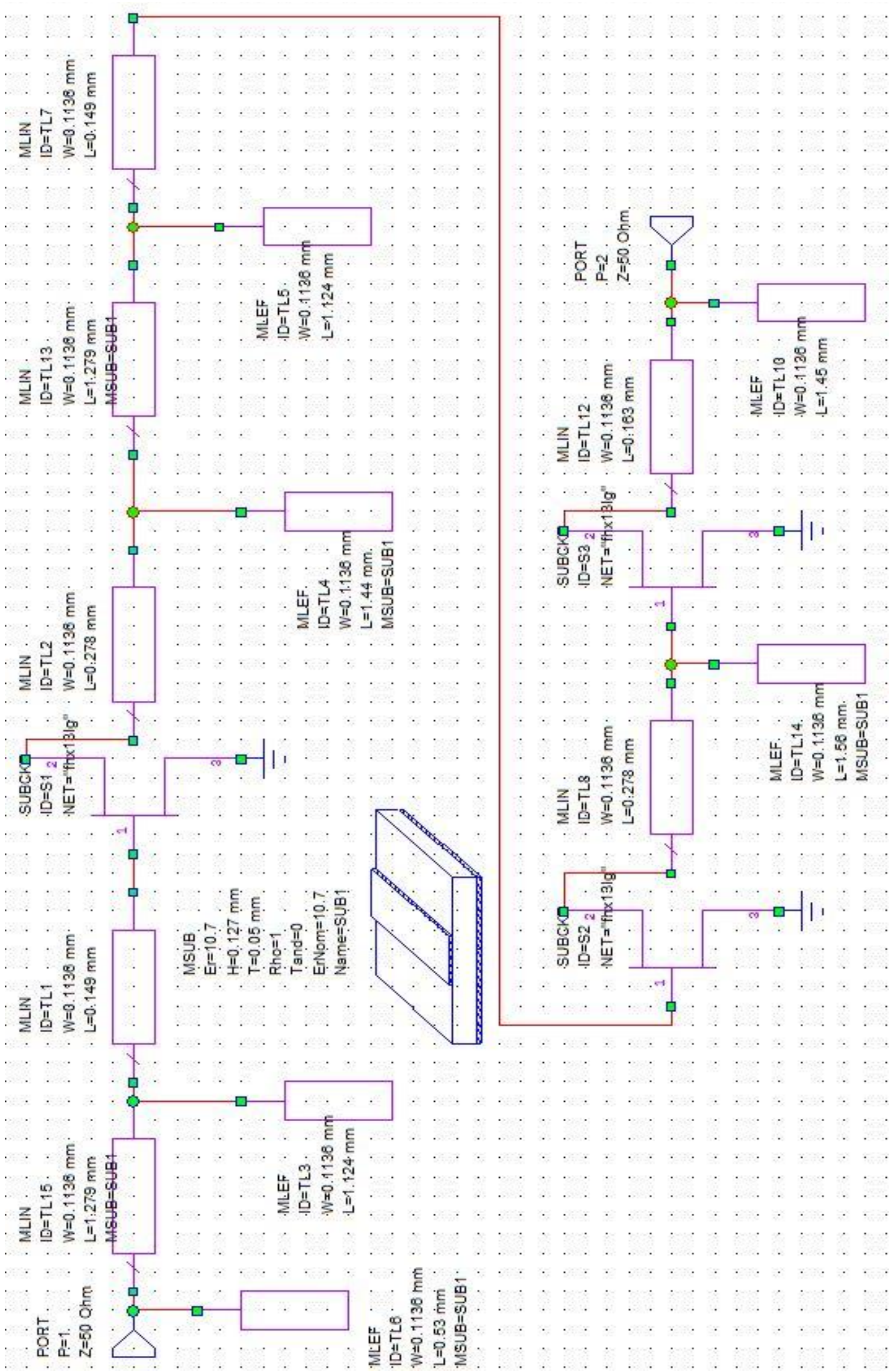


Figure 9.7: Final designed circuit (Microwave Office)

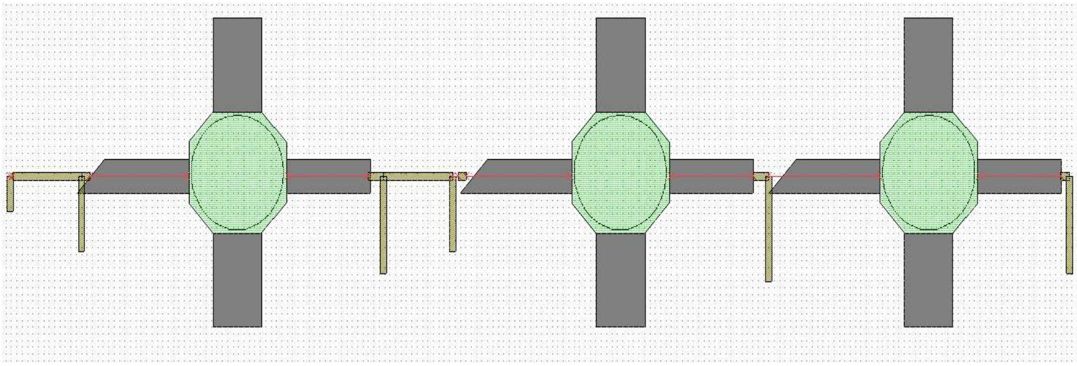


Figure 9.8: 2-D View – Scaled Dimensions for visibility (Microwave Office)

Figure 9.8 is the 2-D presentation of the multistage amplifier. It gives us an understanding of how the system would appear.

Similarly, below is the 3-D presentation of the same. It is not just for visual presentation, but for better understanding. Having an image is as good as overcoming one of the study barrier for better understanding.

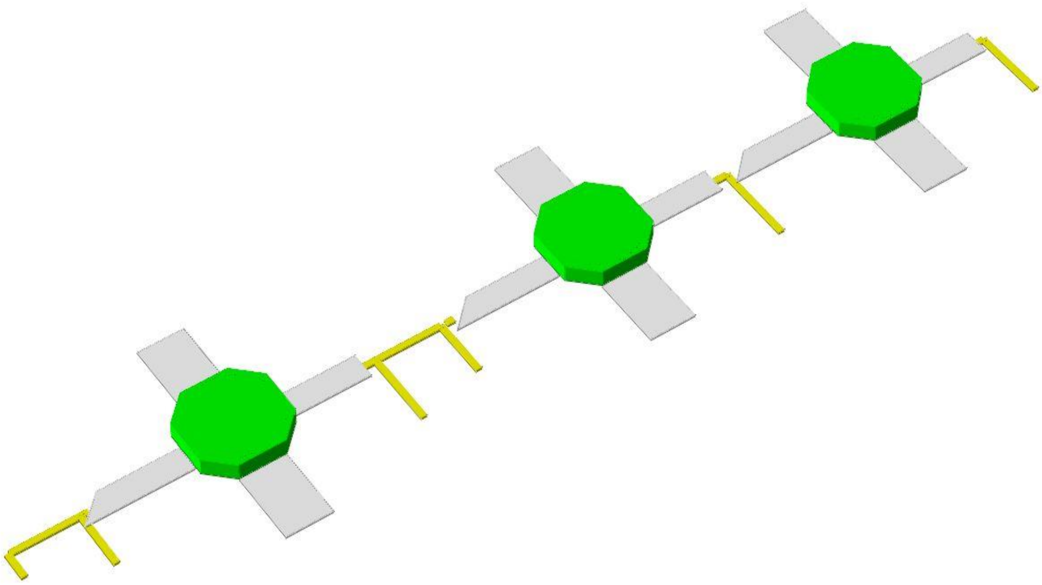


Figure 9.9: 3-D View – Scaled Dimensions for visibility (Microwave Office)

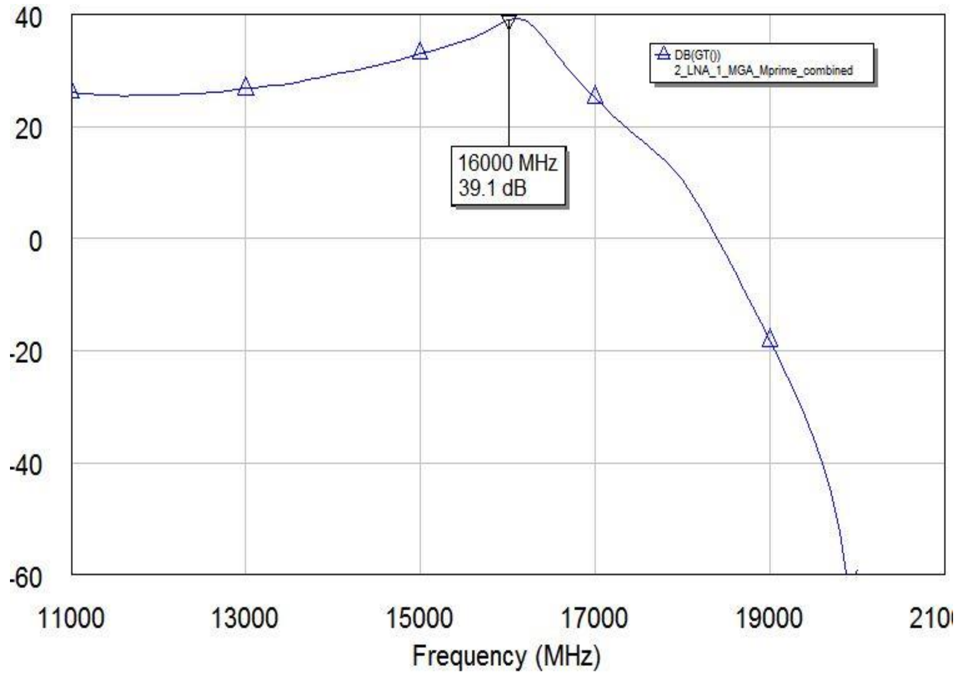


Figure 9.10: Gain of the final design (Microwave Office)

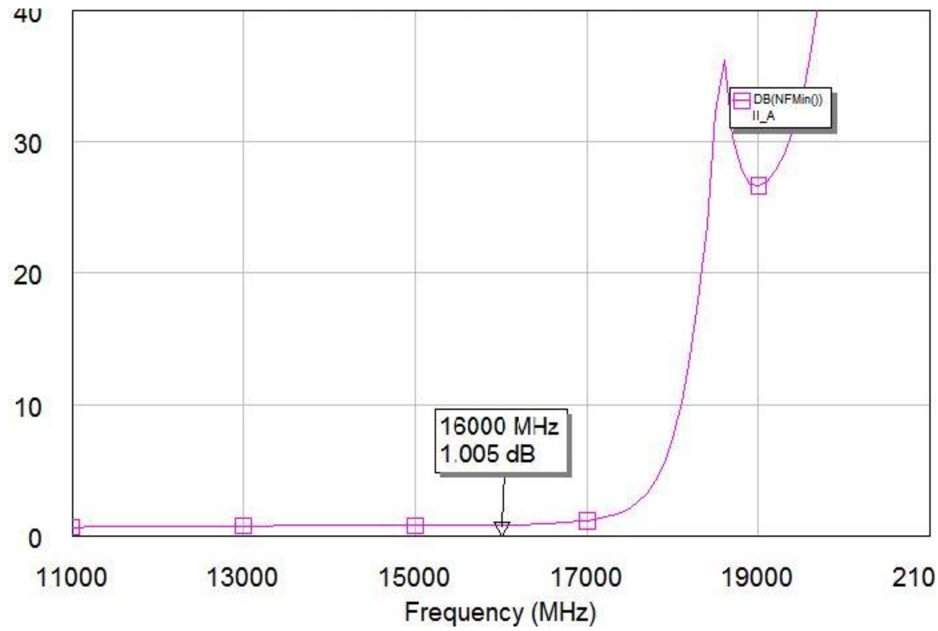


Figure 9.11: Noise Figure of the final design (Microwave Office)

From the simulation graphs, it can be seen that the designed multi stage amplifier is performing as per our expectation and giving results as per our calculations as well. The analysis has been done on multiple simulation software to verify the analysis.

9.4 Modifications and Future Scope

This is a narrow band amplifier design which means there is 10% bandwidth corresponding to a frequency range of 15.2 GHz to 16.8 GHz. Since the operation gain drops almost immediately after 16 GHz, there is a need to modify the design as follows:

Step 1. Using S-Parameters at 17 GHz from the datasheet, we obtain S = $\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$ =

$\begin{bmatrix} 0.543\angle 138.2° & 0.061\angle -17.6° \\ 2.217\angle -63.6° & 0.701\angle 163.4° \end{bmatrix}$

Step 2. Using this S Matrix, we get;

K = 1.043

Δ = $0.26\angle -46.83°$ (signifying Unconditional Stability)

Step 3. Two single-stage LNA, each with a gain of 12.73 dB and one MGA stage with a gain of 14.33 dB providing a total gain of 39.8 dB after combining all the three stages.

Step 4. Two single-stage LNA, each with a noise figure of 0.96 dB, and one MGA stage with a noise figure of 1.10, providing a total noise figure of 1.0085 dB after combining all the three stages.

Step 5. The simulated graphs for gain are shown below for comparison.

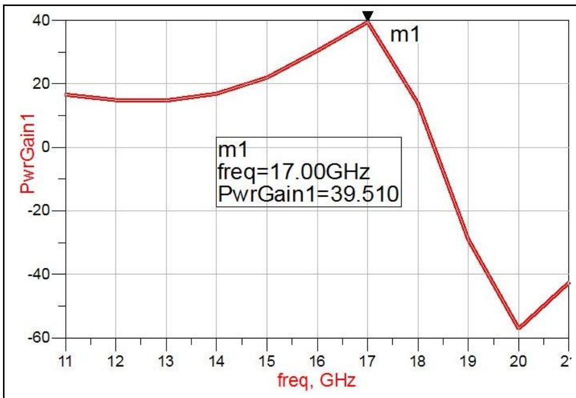


Figure 9.12: Three System Gain without M' network at 17 GHz

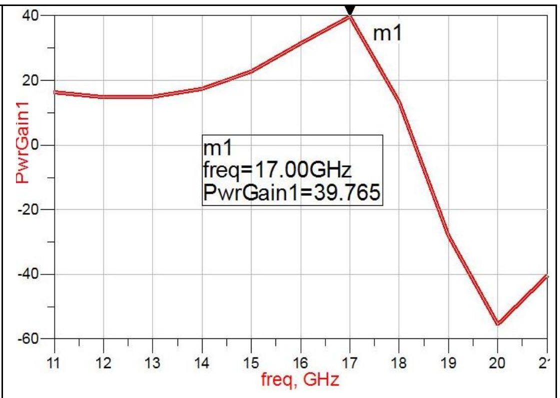


Figure 9.13: Three System Gain with M' network at 17 GHz

Hence, it is easy to conclude that since the gain dropped almost immediately after 17 GHz, this modification is an effective method for a design at 16 GHz which can provide a more stable and flatter gain. Therefore, we can conclude that to have a stable bandwidth of operation, we can use a slightly higher frequency for the S-parameters to avoid sharp decline in gain at the desired frequency.

Chapter 10. Conclusion

10.1 Introduction

Using the standard design procedure for an amplifier design, the amplification was achieved by obtaining the required gain within the specified noise figure of 1 dB at 16 GHz. The design uses the transistor (FHX13/14LG) as an active three port device and characterizes it in terms of S-parameters at the operating frequency.
The D.C. operating point was also established for the design of the D.C. biasing circuit. Followed by the stability check and after verifying the results on MATLAB simulation, the observed values were comparable. This was further verified on other simulation softwares such as Microwave office and ADS. A fourth software used in the verification was provided by the professor himself and is also available with his published textbook "Advanced RF & Microwave Circuit Design".

Each chapter in this work has had a significant role in order to complete the design. While the initial chapters explain the design process, the later chapters provide the procedure to calculate the values which would be used in the design process. Similarly, the later chapters utilize the Smith Chart and use software simulation to verify the results. The design has been challenging due to the strict adherence to the desired specifications. This project has helped me to learn various tools and methodologies which would prove invaluable in my future career in RF and Microwave engineering.

10.2 Comparison Tables for calculated values

A) Amplifier Design Values for Reflection Coefficient

Parameter	Hand Calculation	MATLAB
Γ_{MS}	0.81 ∠ -143.3 = -0.65 - j0.487	0.83 ∠-143.3 = -0.66 - j0.498
Γ_{ML}	0.86 ∠-169.2 = -0.85 - j0.160	0.87∠-169.2 = -0.86 - j0.163
Γ_S	0.62 ∠-143.3 = -0.49 - j0.37	0.62 ∠ -143.3 = -0.49 - j0.37
Γ_L	0.79∠-170 = -0.78 - j0.14	0.79∠-170 = -0.70 - j0.206

Table 10.1: Comparison of Amplifier Design Values

B) Amplifier Design Values

Parameter	**Hand Calculation**	**MATLAB**	**RF/MW E-Book software***	**AWR**
Δ	$0.267\angle{-26.5}$	$0.269\angle-26.4$	$0.27\angle{-27}$	-
K	1.134	1.1377	1.14	-
U	0.1345	0.1345	-	-
U Range in dB	$1.25<U<-1.09$	$1.25<U<-1.09$	-	-
G_{LNA} (1 Stage)	12.94 dB	12.95 dB	12.98 dB	12.98 dB
G_{MGA} (1 Stage)	13.48 dB	13.622 dB	13.62 dB	13.584 dB
G_{Total} (3 Stages)	39.39 dB	39.54 dB	-	39.1 dB
F_{LNA}	0.96 dB	0.96 dB	-	0.865 dB
F_{MGA}	1.04 dB	0.8677 dB	-	1.017 dB
F_{Total} (3 stages)	1.00 dB	1.0079 dB	-	1.005 dB

Table 10.2: Comparison for w by h Values for Z = 50 Ω

*RF/Microwave E-Book software from "Advanced RF & Microwave Circuit Design" written by Dr. Matthew Radmanesh and published by AuthorHouse in 2009.

C) W/h Calculations for Design Values in 50 Ω System

Parameter	Hand Calculation	MATLAB	RF/MW E-Book Software*
Z_0	50 Ω	50 Ω	-
ε_r	1.134	1.1377	10.7
ε_{ff}	7.130	7.128	7.13
w / h	0.895	0.8948	-
Width (w)	0.1136 mm	0.1136 mm	0.1136 mm
Height (h)	0.127 mm	0.127 mm	0.127 mm
λ	7.07 mm	7.07 mm	7.02 mm
λ_0	18.75 mm	18.75 mm	18.75 mm
Microstrip gap	0.0407 mm	0.0409 mm	-

Table 10.3: Comparison for w / h values for Z = 50 Ω

This table helps us to compare results and understand that as long as the procedure is correct, the answer will always be the same irrespective of the method. It is good to know that there are softwares available to verify the hand calculations, however, having an excellent command over the design methodology provides a good foundation as it enables us to learn from our mistakes in the design process.

*RF/Microwave E-Book software from "Advanced RF & Microwave Circuit Design" written by Dr. Matthew Radmanesh and published by AuthorHouse in 2009.

Bibliography

[1] Radmanesh, Matthew M., “RF & Microwave Design Essentials”, Bloomington, Ind.: AuthorHouse, 2007.

[2] Radmanesh, Matthew M., “Advanced RF & Microwave Circuit Design”, Bloomington, Ind.: AuthorHouse, 2009.

[3] David M Pozar, “Microwave Engineering” 4th Edition, John Wiley & Sons, Inc., 2012.

[4] Jahn Stefan, Microstrip Gap, 2007.

[5] RT/duroid 6006/6010LM Datasheet by Roger’s Corporation

Appendix A: MATLAB code for Amplifier Design values

```
%%Given values - ECE698C - 3 stage LNA by Soham S Jagalpure
f=17*10^(9);
Zo=50;

%% INPUT S-PARAMETER
S11 = 0.543*exp (j*138.2*pi*(1/180));
S12 = 0.061*exp ((-1)*j*17.6*pi*(1/180));
S21 = 2.217*exp ((-1)*j*63.6*pi*(1/180));
S22 = 0.701*exp (j*163.4*pi*(1/180));

Fmin=0.63;
Gamma_opt=0.38*exp(-j*175*pi/180);
rn=0.06;

S11_ABS=abs(S11);
S12_ABS=abs(S12);
S21_ABS=abs(S21);
S22_ABS=abs(S22);
%% STABILITY CHECKING
 display ('STABILITY CHECKING')
 Delta_polar = ((S11*S22)-(S12*S21))
 Delta_mag=abs (Delta_polar)
 Delta_mag_Angle=(angle(Delta_polar))*(180/pi);
 K = (1-(S11_ABS^2)-(S22_ABS^2)+(Delta_mag^2))/(2*(S12_ABS*S21_ABS))

if abs(K) > 1
   if Delta_mag < 1
      display (' UNCONDITIONALLY STABLE TRANSISTOR ')
   else
      display (' CONDITIONALLY STABLE TRANSISTOR ')
   end
else
   display (' CONDITIONALLY STABLE TRANSISTOR ')
end

%% Unilateral figure calculation
U=(S12_ABS*S21_ABS*S11_ABS*S22_ABS)/((1-(S11_ABS^2))*(1-(S22_ABS^2)))
U_lower_limit_ratio=(1/((1+U)^2))
U_upper_limit_ratio=(1/((1-U)^2))
U_lower_limit_dB=10*log10(U_lower_limit_ratio)
U_upper_limit_dB=10*log10(U_upper_limit_ratio)

%% Gms and Gml claculation
B1=1+(S11_ABS^2)-(S22_ABS^2)-(Delta_mag^2);
C1=S11-(Delta_polar*(conj(S22)));
Gamma_ms= (B1 - sqrt((B1^2)-(4* abs((C1)^2))))/(2*C1);
```

```
Gamma_ms_Mag=abs(Gamma_ms);
Gamma_ms_Angle=(angle(Gamma_ms))*(180/pi);
B2=1+(S22_ABS^2)-(S11_ABS^2)-(Delta_mag^2);
C2=S22-(Delta_polar*(conj(S11)));
Gamma_ml=(B2-sqrt((B2^2)-(4*(abs(C2))^2)))/(2*C2);
Gamma_ml_Mag=abs(Gamma_ml);
Gamma_ml_Angle=(angle(Gamma_ml))*(180/pi);
Gamma_S=0.58*exp ((-1)*j*132.4*pi*(1/180));
Gamma_S_Mag=abs(Gamma_S);
Gamma_out=0.795*exp (j*170.1*pi*(1/180));
Gamma_out_Mag=abs(Gamma_out);
Gamma_L=conj(S22+(S12*S21*Gamma_S/(1-(S11*Gamma_S))))
Gamma_L=conj(S22+(S12*S21*Gamma_S/(1-(S11*Gamma_S))));
Gamma_L_Mag=abs(Gamma_L)
Gamma_L_Angle=(angle(Gamma_L))*(180/pi);
Gamma_IN=(S11+(S12*S21*Gamma_L/(1-(S22*Gamma_L))));
Gamma_IN_Mag=abs(Gamma_IN)
Gamma_IN_Angle=(angle(Gamma_IN))*(180/pi);
%% maximum power transfer gain calculation
Gain_ideal_dB=10*log10(S21_ABS/S12_ABS);
%Gain_ratio_MGA=(S21_ABS/S12_ABS)*((abs(K))-(sqrt(((abs(K))^2)-1)))
Gain_ratio_MGA=(1/(1-(Gamma_ms_Mag^2)))*(S21_ABS^2)*((1-
(Gamma_ml_Mag^2))/(abs(1-(S22*Gamma_ml)))^2)
Gain_MGA_dB=10*log10(abs(Gain_ratio_MGA))
Gain_ratio_LNA=((1-(Gamma_S_Mag^2))/(abs(1-
(S11*Gamma_S))^2))*(S21_ABS^2)*(1/(1-(Gamma_out_Mag^2)))
Gain_LNA_dB=10*log10(abs(Gain_ratio_LNA))
Gain_cascaded_dB=2*Gain_LNA_dB+Gain_MGA_dB
%% NOISE FIGURE
NF_MGA_num=4*rn*abs((Gamma_ms-Gamma_opt)^2);
NF_MGA_den=(1-(Gamma_ms_Mag)^2)*abs(1+Gamma_opt^2);
NF_MGA_dB=NF_MGA_num/NF_MGA_den+Fmin
NF_MGA_ratio=10^(0.1*NF_MGA_dB)
NF_LNA_dB=0.96
NF_LNA_ratio=10^(0.1*NF_LNA_dB)
NF_cascaded_ratio=NF_LNA_ratio+((NF_LNA_ratio -
1)/(Gain_ratio_LNA))+((NF_MGA_ratio - 1)/(Gain_ratio_LNA^2))
NF_cascaded_dB=10*log10(NF_cascaded_ratio)
```

Result

Name	Value
B1	0.7352
B2	1.1283
C1	-0.2465 + 0.2697i
C2	-0.5305 + 0.1878i
Delta_mag	0.2613
Delta_mag_Angle	-46.8295
Delta_polar	0.1788 - 0.1906i
f	1.7000e+10

Name	Value
Fmin	0.6300
Gain_cascaded_dB	39.8037
Gain_ideal_dB	15.6044
Gain_LNA_dB	12.7319
Gain_MGA_dB	14.3399
Gain_ratio_LNA	18.7581
Gain_ratio_MGA	27.1640
Gamma_IN	-0.5321 + 0.5822i
Gamma_IN_Angle	132.4254
Gamma_IN_Mag	0.7888
Gamma_L	-0.7696 - 0.2589i
Gamma_L_Angle	-161.4047
Gamma_L_Mag	0.8120
Gamma_ml	-0.8781 - 0.3109i
Gamma_ml_Angle	-160.5014
Gamma_ml_Mag	0.9316
Gamma_ms	-0.6050 - 0.6618i
Gamma_ms_Angle	-132.4300
Gamma_ms_Mag	0.8967
Gamma_opt	-0.3786 - 0.0331i
Gamma_out	-0.7832 + 0.1367i
Gamma_out_Mag	0.7950
Gamma_S	-0.3911 - 0.4283i
Gamma_S_Mag	0.5800
K	1.0427
NF_cascaded_dB	1.0085
NF_cascaded_ratio	1.2614
NF_LNA_dB	0.9600
NF_LNA_ratio	1.2474
NF_MGA_dB	1.1087
NF_MGA_den	0.2239
NF_MGA_num	0.1072
NF_MGA_ratio	1.2908
rn	0.0600
S11	-0.4048 + 0.3619i
S11_ABS	0.5430
S12	0.0581 - 0.0184i
S12_ABS	0.0610
S21	0.9858 - 1.9858i
S21_ABS	2.2170
S22	-0.6718 + 0.2003i
S22_ABS	0.7010
U	0.1435
U_lower_limit_dB	-1.1650
U_lower_limit_ratio	0.7647
U_upper_limit_dB	1.3458
U_upper_limit_ratio	1.3633
Zo	50

Appendix B: MATLAB code for Microstrip w/h and Length Calculation

```
c=3e11;     %speed of light mm/sec     3e8 m/s
%Now to find w/h
disp('Case 1 - Find w_by_h and Lambda ');
disp('_______________________________');
e_r_case1_given=input('enter value of e_r:  ');     %dielectric constant for quartz = 5
Z_0_case1_given=input('enter value of Z_0:  ');     %characteristics impedance Z_0 =
50,75
%woverh=0.2739;
A=(Z_0_case1_given/60)*sqrt((e_r_case1_given+1)/2)+((e_r_case1_given-
1)/(e_r_case1_given+1)*(0.23+0.11/e_r_case1_given));
freq_0=input('enter value of frequency:  ');     %height in cm => 1cm = 10mm
lambda_0=(c/freq_0);     %lamda0 in cm
lambda_TEM=lambda_0/sqrt(e_r_case1_given);
disp('lambda_0 calculated in mm : ');
disp(lambda_0);
w_by_h_calculated = (8*exp(A))/(exp(2*A)-2);
disp('_______________________________');
disp('w_by_h calculated is: ');
disp(w_by_h_calculated);
disp('_______________________________');
disp('calculation for Lambda in mm: ')
reply = input ('Is W/h <0.6 ? yes or no: ', 's');
if strcmp(reply,'yes')
lambda=(lambda_0/sqrt(e_r_case1_given))*(e_r_case1_given/(1+0.6*(e_r_case1_give
n-1)*(w_by_h_calculated)^0.0297))^0.5
else
lambda=(lambda_0/sqrt(e_r_case1_given))*(e_r_case1_given/(1+0.63*(e_r_case1_giv
en-1)*(w_by_h_calculated)^0.1255))^0.5
end;
lambda_by_lambda_tem=lambda/lambda_TEM;
disp('****************************');
disp('CASE 2 - Calculation of Lambda and Z_0 : ')
disp('****************************');
e_r_case2_given=input('enter value of e_r:  ');
w_by_h_case2_given=input('enter value of w_by_h:  ');
disp('_______________________________');
disp('calculation for e_ff : ')
reply = input ('Is W/h <=2? yes or no: ', 's');
if strcmp(reply,'yes')
   e_ff=((e_r_case2_given+1)/2)+((e_r_case2_given-
1)/2)*((1/(sqrt(1+(12*(1/w_by_h_case2_given))))))+(0.04*(1-
w_by_h_case2_given)^2)
else
   e_ff=((e_r_case2_given+1/2)+(e_r_case2_given-
1/2)*((1+12*(1/w_by_h_case2_given))^(-0.5)))
end;
```

```
disp('_________________________________');
disp('Calculation for Z_0 : ')
reply = input ('Is W/h <=1? yes or no: ', 's');
if strcmp(reply,'yes')
   Z0_calculated
=(60/sqrt(e_ff))*(log(8*(1/w_by_h_case2_given)+(w_by_h_case2_given/4)))
else
   Z0_calculated
=(120*pi/(sqrt(e_ff)*(w_by_h_case2_given+1.393+0.667*(log(w_by_h_case2_given+
1.444)))))
end
disp('_________________________________');
disp('Calculation for Lambda in mm: ')
reply = input ('Is W/h <0.6 ? yes or no: ', 's');
if strcmp(reply,'yes')
lambda=(lambda_0/sqrt(e_r_case2_given))*(e_r_case2_given/(1+0.6*(e_r_case2_give
n-1)*(w_by_h_case2_given)^0.0297))^0.5
else
lambda=(lambda_0/sqrt(e_r_case2_given))*(e_r_case2_given/(1+0.63*(e_r_case2_giv
en-1)*(w_by_h_case2_given)^0.1255))^0.5
end;
disp('*****************************');
disp('*****************************');
disp('Calculations for Width W, Length l, and thickness t in mm');
height=input('enter the value of h in mm: ');
%f=input('enter the value of freq: ');
disp('Frequency taken into Calculation is freq_0 = ');
disp(freq_0);
width_w=height*w_by_h_calculated;
length_l=c/(4*freq_0*(sqrt(e_ff)));
deno=exp((Z_0_case1_given)*((sqrt(e_r_case1_given+1.41))/87));
thickness_t=13.4*(((5.98*height)/deno)-0.8*width_w);
disp('height h in mm is: '); disp(height);
disp('_________________________________');
disp('calculated width w in mm is: '); disp(width_w);
disp('calculated thickness t in mm is: '); disp(thickness_t);
disp('calculated length l in mm is: '); disp(length_l);
```

RESULT for Z0 = 50 Ω

Name	Value
A	2.2148
c	3.0000e+11
deno	7.3888
e_ff	7.1281
e_r_case1_given	10.7000
e_r_case2_given	10.7000
freq_0	1.6000e+10
height	0.1270
height_calculated	0.0564
lambda	7.0735
lambda_0	18.7500

Name	Value
lambda_by_lambda_tem	1.2340
lambda_TEM	5.7320
length_l	1.7557
reply	'no'
thickness_t	0.1592
w_by_h_calculated	0.8948
w_by_h_case2_given	0.8948
width_w	0.1136
Z0_calculated	49.7852
Z_0_case1_given	50

RESULT for Z0 = 120 Ω

Name	Value
A	5.0366
c	3.0000e+11
deno	121.5008
e_ff	6.2045
e_r_case1_given	10.7000
e_r_case2_given	10.7000
freq_0	1.6000e+10
height	0.1270
lambda	7.4520
lambda_0	18.7500

Name	Value
lambda_by_lambda_tem	1.3001
lambda_TEM	5.7320
length_l	1.8819
reply	'yes'
thickness_t	0.0130
w_by_h_calculated	0.0520
w_by_h_case2_given	0.0520
width_w	0.0066
Z0_calculated	121.3070
Z_0_case1_given	120

Appendix C: MATLAB code for Impedance Z_0

```
c=3e11;     %speed of light cm/sec     3e8 m/s
%Now to find w/h
e_r_case1_given=input('enter value of e_r:  ');     %dielectric constant for quartz = 5
Z_0_case1_given=input('enter value of Z_0:  ');     %characteristics impedance Z_0 =
50,75
height_mm=input('enter value of height_mm:  ');
A=(Z_0_case1_given/60)*sqrt((e_r_case1_given+1)/2)+((e_r_case1_given-
1)/(e_r_case1_given+1)*(0.23+0.11/e_r_case1_given));
freq=input('enter value of frequency:  ');     %height in cm => 1cm = 10mm
w_by_h_calculated = (8*exp(A))/(exp(2*A)-2)
%% Calculate e_eff(f)
disp('Case 1 - Find e_eff_f at 16 GHz ');disp('_______________________________');
e_ff_0=((e_r_case1_given+1)/2)+((e_r_case1_given-
1)/2)*((1/(sqrt(1+(12*(1/w_by_h_calculated))))))+(0.04*(1-w_by_h_calculated)^2)
F_num_A=4*height_mm*freq*sqrt(e_r_case1_given-1)/c
F_num_B=0.5+(1+2*log10(1+w_by_h_calculated))^2
F=F_num_A*F_num_B
e_eff_freq=((sqrt(e_r_case1_given)-sqrt(e_ff_0))/(1+4*(F^-1.5))+sqrt(e_ff_0))^2
%% Calculate Impedance Z_0
disp('Case 2 - Find e_eff_f at freq: ');disp('_______________________________');
Z_0_f=Z_0_case1_given*((e_eff_freq-1)/(e_ff_0-1))*sqrt(e_ff_0/e_eff_freq)
```

RESULT

For Frequency f = 16 GHz

Name	Value
A	2.2148
ans	3.1145
c	3.0000e+11
deno	7.3888
e_eff_freq	7.2235
e_ff	7.1280
e_ff_0	7.1280
e_r_case1_given	10.7000
e_r_case2_given	10.7000
F	0.2463
F_num_A	0.0844
F_num_B	2.9184
freq	1.6000e+10
freq_0	1.6000e+10
height	0.1270
height_calculated	0.0564
height_mm	0.1270
lambda	7.0735
lambda_0	18.7500

For Frequency f = 16.8 GHz

Name	Value
A	2.2148
ans	3.1145
c	3.0000e+11
deno	7.3888
e_eff_freq	7.2306
e_ff	7.1280
e_ff_0	7.1280
e_r_case1_given	10.7000
e_r_case2_given	10.7000
F	0.2586
F_num_A	0.0886
F_num_B	2.9184
freq	1.6800e+10
freq_0	1.6000e+10
height	0.1270
height_calculated	0.0564
height_mm	0.1270
lambda	7.0735
lambda_0	18.7500

For Frequency f = 16 GHz

lambda_by_lambda_tem	1.2340
lambda_TEM	5.7320
length_l	1.7557
reply	'no'
thickness_t	0.1592
w_by_h_calculated	0.8948
w_by_h_case2_given	0.8948
width_w	0.1136
Z0_calculated	49.7852
Z_0_case1_given	50
Z_0_f	50.4424

Result log

enter value of e_r: 10.7

enter value of Z_0: 50

enter value of height_mm: 0.127

enter value of frequency: 16e9

w_by_h_calculated =

0.8948

Case 1 - Find e_eff_f at 16 GHz

e_ff_0 =   7.1280

F_num_A =   0.0844

F_num_B =   2.9184

F =   0.2463

e_eff_freq =   7.2235

Case 2 - Find e_eff_f at 16 GHz

Z_0_f =   50.4424

For Frequency f = 16.8 GHz

lambda_by_lambda_tem	1.2340
lambda_TEM	5.7320
length_l	1.7557
reply	'no'
thickness_t	0.1592
w_by_h_calculated	0.8948
w_by_h_case2_given	0.8948
width_w	0.1136
Z0_calculated	49.7852
Z_0_case1_given	50
Z_0_f	50.4749

Result log

enter value of e_r: 10.7

enter value of Z_0: 50

enter value of height_mm: 0.127

enter value of frequency: 16.8e9

w_by_h_calculated =

0.8948

Case 1 - Find e_eff_f at 16 GHz

e_ff_0 =   7.1280

F_num_A =   0.0886

F_num_B =   2.9184

F =   0.2586

e_eff_freq =   7.2306

Case 2 - Find e_eff_f at 16 GHz

Z_0_f =   50.4749

Appendix D: Datasheet of FHX13/14LG

FHX13LG, FHX14LG

Super Low Noise HEMT

FEATURES

- Low Noise Figure: 0.45dB (Typ.)@f=12GHz (FHX13)
- High Associated Gain: 13.0dB (Typ.)@f=12GHz
- Lg ≤ 0.15μm, Wg = 200μm
- Gold Gate Metallization for High Reliability
- Cost Effective Ceramic Microstrip (SMT) Package
- Tape and Reel Packaging Available

DESCRIPTION

The FHX13LG, FHX14LG is a Super High Electron Mobility Transistor(SuperHEMT™) intended for general purpose, ultra-low noise and high gain amplifiers in the 2-18GHz frequency range. The devices are packaged in cost effective, low parasitic, hermetically sealed metal-ceramic package for high volume telecommunication, TVRO, VSAT or other low noise applications.

Eudyna stringent Quality Assurance Program assures the highest reliability and consistent performance.

ABSOLUTE MAXIMUM RATING (Ambient Temperature Ta=25°C)

Item	Symbol	Rating	Unit
Drain-Source Voltage	V_{DS}	3.5	V
Gate-Source Voltage	V_{GS}	-3.0	V
Total Power Dissipation	P_t*	180	mW
Storage Temperature	T_{stg}	-65 to +175	°C
Channel Temperature	T_{ch}	175	°C

***Note:** Mounted on Al_2O_3 board (30 x 30 x 0.65mm)

Eudyna recommends the following conditions for the reliable operation of GaAs FETs:

1. The drain-source operating voltage (V_{DS}) should not exceed 2 volts.
2. The forward and reverse gate currents should not exceed 0.2 and -0.05 mA respectively with gate resistance of 4000Ω.
3. The operating channel temperature (T_{ch}) should not exceed 80°C.

ELECTRICAL CHARACTERISTICS (Ambient Temperature Ta=25°C)

Item		Symbol	Condition	Limit Min.	Limit Typ.	Limit Max.	Unit
Saturated Drain Current		I_{DSS}	V_{DS} = 2V, V_{GS} =0V	10	30	60	mA
Transconductance		g_m	V_{DS} = 2V, I_{DS} =10mA	35	50	-	mS
Pinch-off Voltage		V_p	V_{DS} = 2V, I_{DS} =1mA	-0.1	-0.7	-1.5	V
Gate Source Breakdown Voltage		V_{GSO}	I_{GS} = -10μA	-3.0	-	-	V
Noise Figure	FHX13LG	NF	V_{DS} = 2V, I_{DS} = 10mA, f = 12GHz	-	0.45	0.50	dB
Associated Gain	FHX13LG	G_{as}		11.0	13.0	-	dB
Noise Figure	FHX14LG	NF		-	0.55	0.60	dB
Associated Gain	FHX14LG	G_{as}		11.0	13.0	-	dB
Thermal Resistance		R_{th}	Channel to Case	-	300	400	°C/W

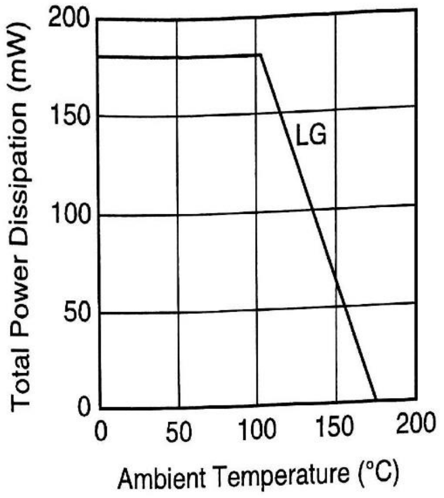
AVAILABLE CASE STYLES: LG

Note: RF parameters for LG devices are measured on a sample basis as follows:

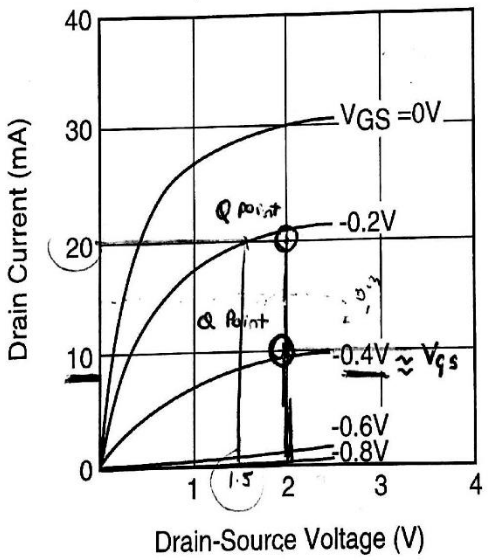
Lot qty.			Sample qty.	Accept/Reject
1200	or	less	125	(0,1)
1201	to	3200	200	(0,1)
3201	to	10000	315	(1,2)
10001	or	over	500	(1,2)

Eudyna

POWER DERATING CURVE



DRAIN CURRENT vs. DRAIN-SOURCE VOLTAGE



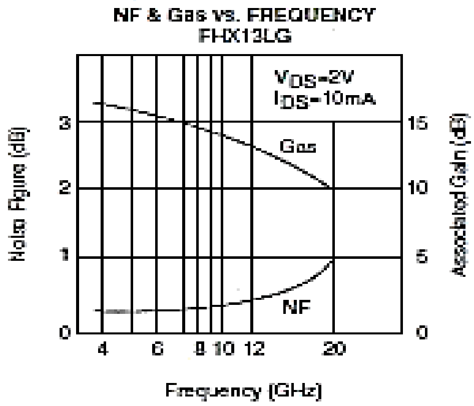
① RFC choke betn DC - RF/mw ckt

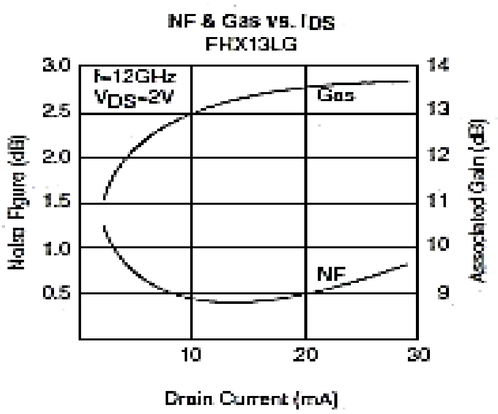
② quarter wave transformer betn DC - RF/mw
Z_0 transformer ↑↑↑ $Z_0 >> 1$

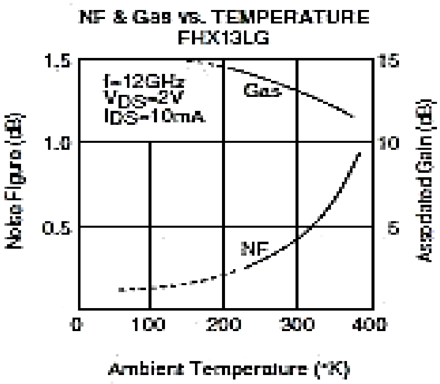
③ C ↑↑ to QW transformer (creates open ckt at i/p)

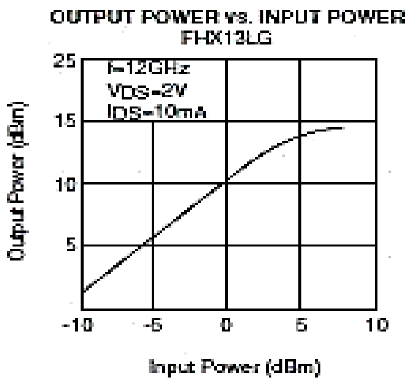
④ Symm Stubs for better i/p VSWR

$\boxed{V_{GS} \approx -0.4\,V}$


NF & Gas vs. FREQUENCY
FHX13LG
VDS=2V
IDS=10mA
Gas
NF
Noise Figure (dB)
Associated Gain (dB)
Frequency (GHz)

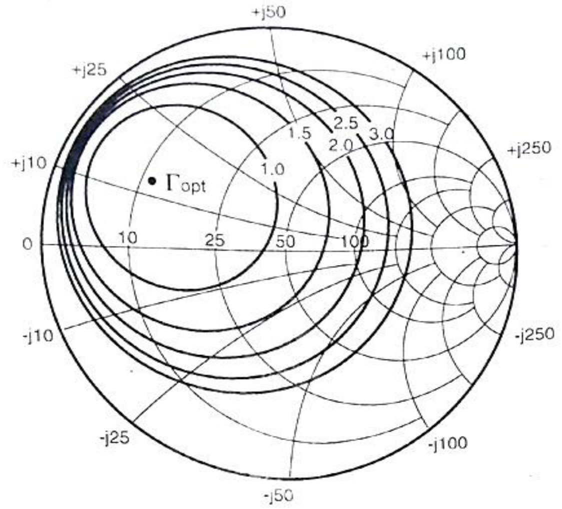

NF & Gas vs. IDS
FHX13LG
f=12GHz
VDS=2V
Gas
NF
Noise Figure (dB)
Associated Gain (dB)
Drain Current (mA)


NF & Gas vs. TEMPERATURE
FHX13LG
f=12GHz
VDS=2V
IDS=10mA
Gas
NF
Noise Figure (dB)
Associated Gain (dB)
Ambient Temperature (°K)


OUTPUT POWER vs. INPUT POWER
FHX13LG
f=12GHz
VDS=2V
IDS=10mA
Output Power (dBm)
Input Power (dBm)

TYPICAL NOISE FIGURE CIRCLE

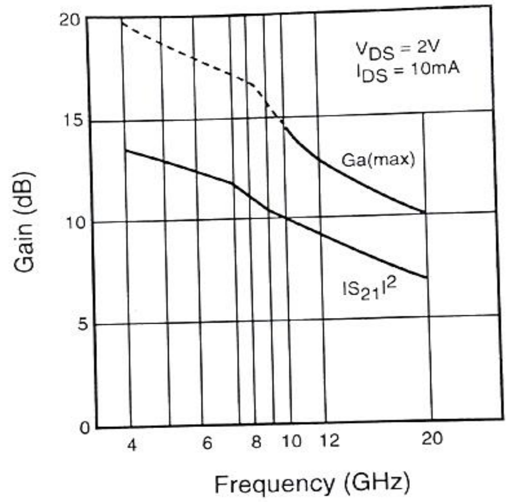
FHX13LG



f = 12 GHz
V_{DS} = 2V
I_{DS} = 10mA

Γopt = 0.61∠150°
Rn/50 = 0.04
NFmin = 0.45dB

Ga(max) & $|S_{21}|^2$ vs. FREQUENCY

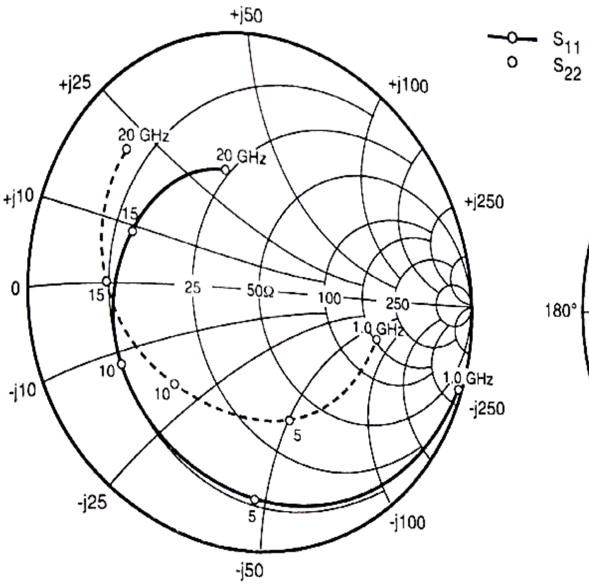


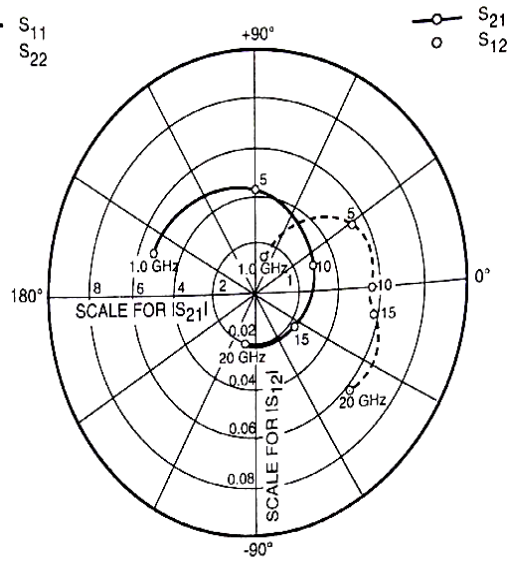
NOISE PARAMETERS

FHX13LG

V_{DS}=2V, I_{DS}=10mA

Freq. (GHz)	Γopt (MAG)	Γopt (ANG)	NFmin (dB)	r_n Rn/50
2	0.96	29	0.33	0.22
4	0.92	57	0.34	0.20
6	0.86	83	0.35	0.15
8	0.79	107	0.37	0.11
10	0.71	129	0.40	0.07
12	0.61	150	0.45	0.04
14	0.50	168	0.53	0.04
16	0.38	-175	0.63	0.06
18	0.24	-161	0.83	0.10



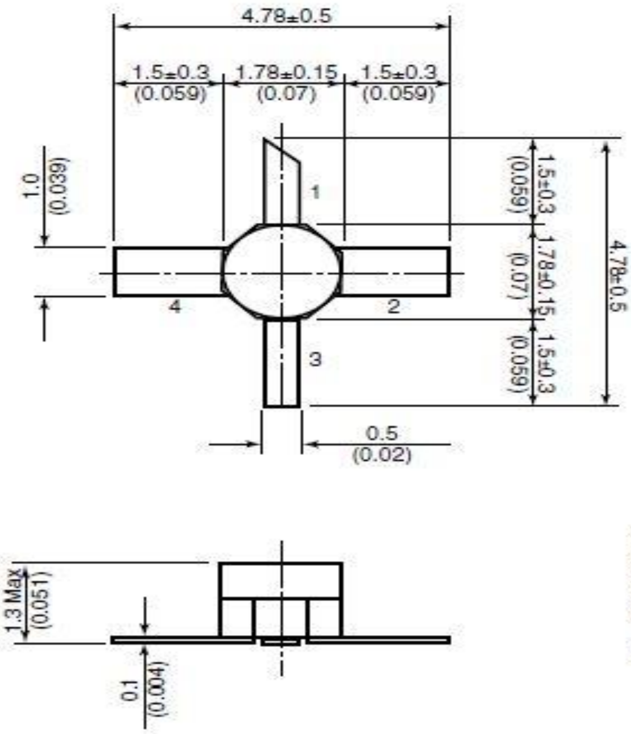


S-PARAMETERS
FHX13/14LG
V_{DS} = 2V, I_{DS} = 10mA

FREQUENCY (MHZ)	S11 MAG	S11 ANG	S21 MAG	S21 ANG	S12 MAG	S12 ANG	S22 MAG	S22 ANG
1000	0.988	-20.0	5.327	160.1	0.015	75.7	0.574	-16.3
2000	0.956	-39.5	5.133	141.0	0.028	63.3	0.560	-32.1
3000	0.908	-58.1	4.851	123.0	0.039	50.1	0.539	-47.3
4000	0.862	-75.5	4.534	105.9	0.048	39.0	0.522	-62.0
5000	0.811	-91.6	4.213	89.7	0.053	29.3	0.502	-75.6
6000	0.763	-107.1	3.886	74.4	0.056	21.0	0.488	-89.6
7000	0.727	-121.1	3.582	60.0	0.057	13.2	0.487	-103.0
8000	0.701	-133.3	3.300	46.4	0.056	7.9	0.498	-114.9
9000	0.682	-144.1	3.078	33.8	0.055	3.5	0.515	-125.0
10000	0.659	-154.2	2.899	21.4	0.055	-0.0	0.531	-134.4
11000	0.636	-164.4	2.748	9.3	0.054	-2.6	0.544	-144.0
12000	0.618	-175.4	2.593	-3.3	0.054	-5.2	0.561	-155.1
13000	0.608	175.5	2.466	-14.8	0.054	-5.7	0.590	-164.0
14000	0.596	166.6	2.366	-26.6	0.055	-7.8	0.619	-172.4
15000	0.585	158.3	2.279	-38.3	0.056	-9.7	0.654	-179.7
16000	0.564	148.8	2.244	-50.7	0.058	-12.8	0.677	172.6
17000	0.543	138.2	2.217	-63.6	0.061	-17.6	0.701	163.4
18000	0.525	127.3	2.185	-77.1	0.063	-24.7	0.727	154.1
19000	0.506	116.2	2.143	-91.4	0.063	-33.1	0.748	143.6
20000	0.470	106.5	2.089	-105.4	0.061	-43.7	0.763	137.2

Case Style "LG"

Metal-Ceramic Hermetic Package



1. **Gate**
2. **Source**
3. **Drain**
4. **Source**

Unit: mm(inches)

Appendix E: Microstrip Gap Capacitance and its MATLAB code

Microstrip gap

A symmetrical microstrip gap can be modeled by two open ends with a capacitive series coupling between the two ends. The physical layout is shown in fig. 11.6.

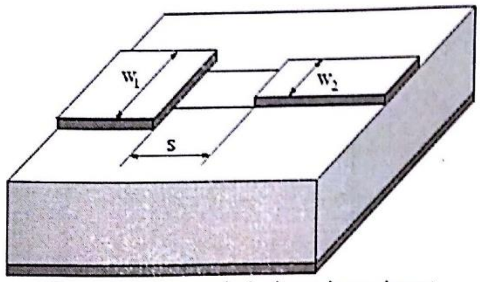


Figure 11.6: symmetrical microstrip gap layout

The equivalent π-network of a microstrip gap is shown in figure 11.7. The values of the components are according to [27] and [30].

$$C_S\ [\mathrm{pF}] = 500 \cdot h \cdot \exp\left(-1.86 \cdot \frac{s}{h}\right) \cdot Q_1 \cdot \left(1 + 4.19\left(1 - \exp\left(-0.785 \cdot \sqrt{\frac{h}{W_1} \cdot \frac{W_2}{W_1}}\right)\right)\right) \tag{11.193}$$

$$C_{P1} = C_1 \cdot \frac{Q_2 + Q_3}{Q_2 + 1} \tag{11.194}$$

$$C_{P2} = C_2 \cdot \frac{Q_2 + Q_4}{Q_2 + 1} \tag{11.195}$$

with

$$Q_1 = 0.04598 \cdot \left(0.03 + \left(\frac{W_1}{h}\right)^{Q_5}\right) \cdot (0.272 + 0.07 \cdot \varepsilon_r) \tag{11.196}$$

$$Q_2 = 0.107 \cdot \left(\frac{W_1}{h} + 9\right) \cdot \left(\frac{s}{h}\right)^{3.23} + 2.09 \cdot \left(\frac{s}{h}\right)^{1.05} \cdot \frac{1.5 + 0.3 \cdot W_1/h}{1 + 0.6 \cdot W_1/h} \tag{11.197}$$

$$Q_3 = \exp\left(-0.5978 \cdot \left(\frac{W_2}{W_1}\right)^{1.35}\right) - 0.55 \tag{11.198}$$

$$Q_4 = \exp\left(-0.5978 \cdot \left(\frac{W_1}{W_2}\right)^{1.35}\right) - 0.55 \tag{11.199}$$

$$Q_5 = \frac{1.23}{1 + 0.12 \cdot (W_2/W_1 - 1)^{0.9}} \tag{11.200}$$

with C_1 and C_2 being the open end capacitances of a microstrip line (see eq. (11.192)). The numerical error of the capacitive admittances is less than $0.1\,\mathrm{mS}$ for

$$
\begin{aligned}
0.1 \le W_1/h &\le 3 \\
0.1 \le W_2/h &\le 3 \\
1 \le W_2/W_1 &\le 3 \\
6 \le \epsilon_r &\le 13 \\
0.2 \le s/h &\le \infty \\
0.2\mathrm{GHz} \le f &\le 18\mathrm{GHz}
\end{aligned}
$$

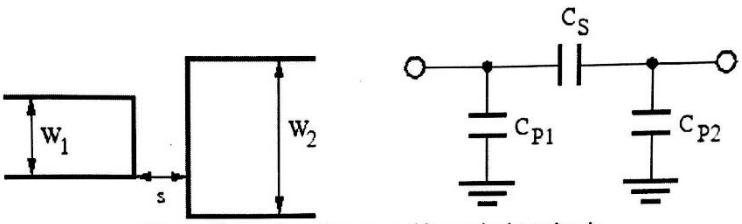


Figure 11.7: microstrip gap and its equivalent circuit

The Y-parameters for the given equivalent small signal circuit can be written as stated in eq. (11.201) and are easy to convert to scattering parameters.

$$
Y = \begin{bmatrix} j\omega \cdot (C_{P1} + C_S) & -j\omega C_S \\ -j\omega C_S & j\omega \cdot (C_{P2} + C_S) \end{bmatrix} \tag{11.201}
$$

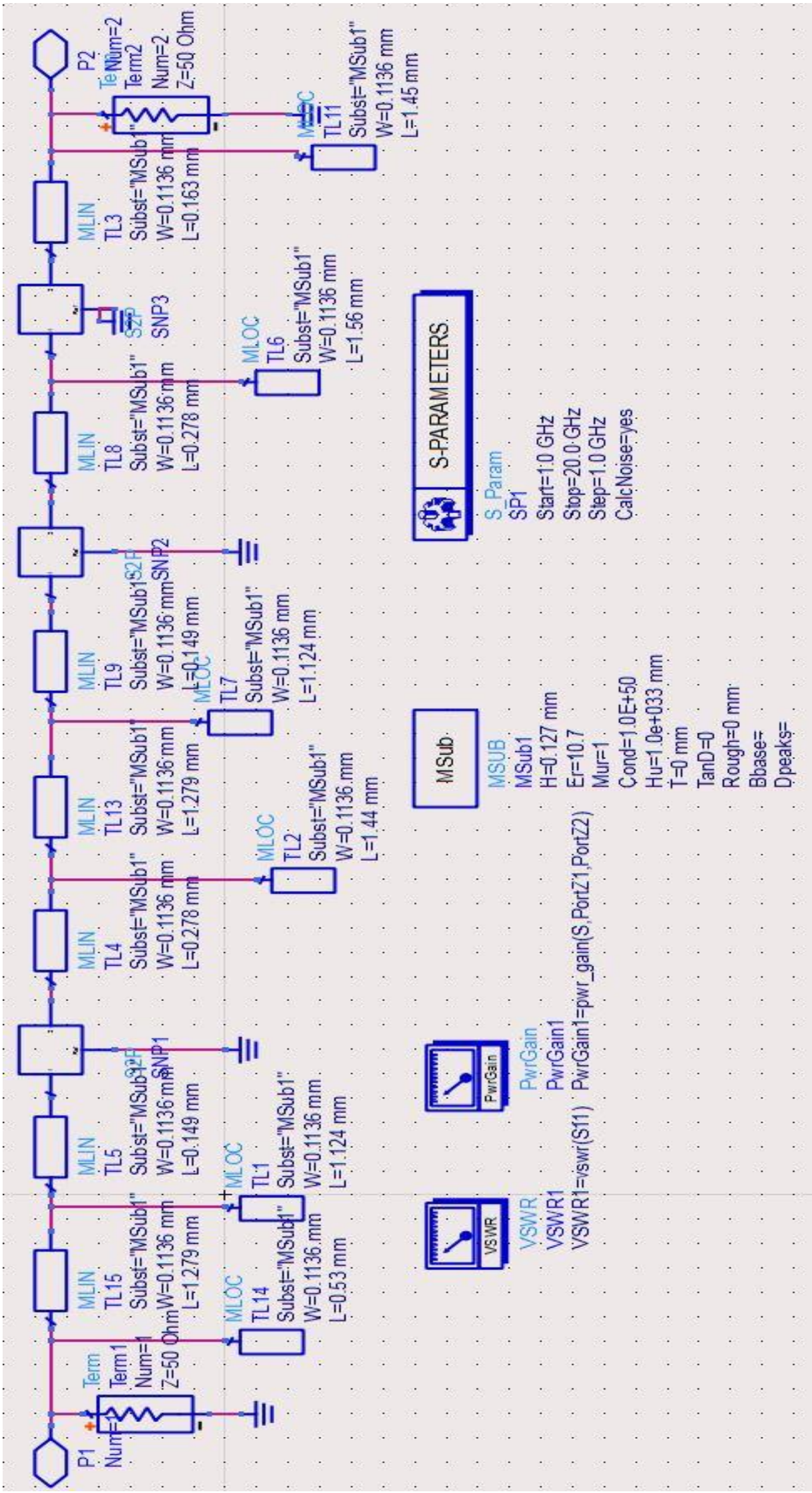
MATLAB code

```
%% 3-Stage LNA at 16GHz by SohamSJ %%
disp('Find Gap Capacitance value ');
disp('_________________________________');
height_mm=input('enter value of h:  ');        %height in mm
width_W1_or_W2=input('enter value of W:  ');     %width in mm
e_r=input('enter value of e_r:  ');   %dielectric constant
capacitance_pF=input('enter value of capacitance in pF:  ');   %capacitance
Q5 = 1.23/(1+(0.12*((width_W1_or_W2/width_W1_or_W2)-1)^0.9))
Q1 = 0.04598 *(0.03 + ((width_W1_or_W2/height_mm )^(Q5)))*(0.272 + 0.07*e_r)
a=Q1*(1+4.19*(1-exp(-0.785*sqrt(height_mm/width_W1_or_W2))))
b=capacitance_pF/(500*height_mm)
s=(height_mm*log(b/a))/(-1.86)
```

RESULT :

Name	Value
a	0.1424
b	0.0783
capacitance_pF	4.9700
e_r	10.7000
height_mm	0.1270
Q1	0.0423
Q5	1.2300
s	0.0409
width_W1_or_W2	0.1136

Appendix F: Circuit and Simulation Result on ADS



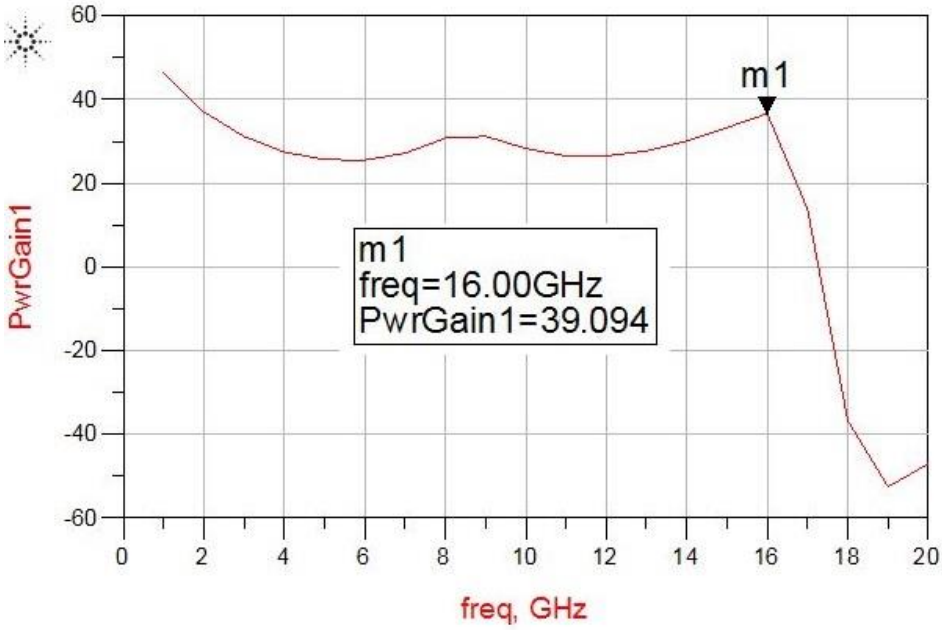


Figure Appendix F.1: Gain (ADS)

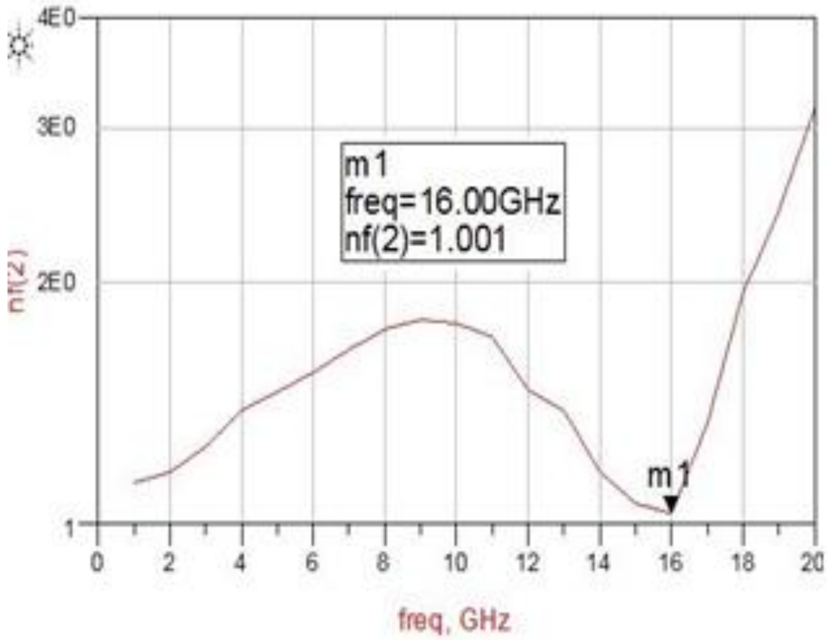


Figure Appendix F.2: Noise (ADS)

Appendix G: RT/duroid 6006/6010LM Datasheet for w/h calculations

ROGERS CORPORATION

Data Sheet

RT/duroid® 6006/6010LM
High Frequency Laminates

RT/duroid® 6006/6010LM microwave laminates are ceramic-PTFE composites designed for electronic and microwave circuit applications requiring a high dielectric constant. RT/duroid 6006 laminate is available with a dielectric constant value of 6.15 and RT/duroid 6010LM laminate has a dielectric constant of 10.2.

RT/duroid 6006/6010LM microwave laminates feature ease of fabrication and stability in use. They have tight dielectric constant and thickness control, low moisture absorption, and good thermal mechanical stability.

RT/duroid 6006/6010LM laminates are supplied clad both sides with 1/2 oz. to 2 oz./ft^2 (18 to 70 µm) standard and reverse treated electrodeposited copper foil. Thick aluminum, brass, or copper plate on one side may be specified.

Standard tolerance dielectric thicknesses of 0.010", 0.025", 0.050", 0.075", and 0.100" (0.254, 0.635, 1.270, 1.905, 2.54 mm) are available. When ordering RT/duroid 6006 and RT/duroid 6010LM laminates, it is important to specify dielectric thickness and weight of copper foil required.

Features and benefits:

- High dielectric constant for circuit size reduction
- Low loss. Ideal for operating at X-band or below
- Low Z-axis expansion for RT/duroid 6010LM. Provides reliable plated through holes in multilayer boards
- Low moisture absorption for RT/duroid 6010LM. Reduces effects of moisture on electrical loss
- Tight ε_r and thickness control for repeatable circuit performance

Some Typical Applications:

- Patch Antennas
- Satellite Communications Systems
- Power Amplifiers
- Aircraft Collision Avoidance Systems
- Ground Radar Warning Systems

100 S. Roosevelt Avenue, Chandler, AZ 85226
Tel: 480-961-1382 Fax: 480-961-4533 www.rogerscorp.com

PROPERTY	TYPICAL VALUES RT/duroid 6006	TYPICAL VALUES RT/duroid 6010.2LM	DIRECTION	UNITS	CONDITIONS	TEST METHOD
[2]Dielectric Constant ε_r Process	6.15± 0.15	10.2 ± 0.25	Z		10 GHz 23°C	IPC-TM-650 2.5.5.5 Clamped stripline
[3]Dielectric Constant ε_r Design	6.45	10.7	Z		8 GHz - 40 GHz	Differential Phase Length Method
Dissipation Factor, tan δ	0.0027	0.0023	Z		10 GHz/A	IPC-TM-650 2.5.5.5
Thermal Coefficient of ε_r	-410	-425	Z	ppm/°C	-50 to 170°C	IPC-TM-650 2.5.5.5
Surface Resistivity	$7X10^7$	$5X10^6$		Mohm	A	IPC 2.5.17.1
Volume Resistivity	$2X10^7$	$5X10^5$		Mohm•cm	A	IPC 2.5.17.1
Youngs' Modulus						ASTM D638 (0.1/min. strain rate)
under tension	627 (91) 517 (75)	931 (135) 559 (81)	X Y	MPa (kpsi)	A	
ultimate stress	20 (2.8) 17 (2.5)	17 (2.4) 13 (1.9)	X Y	MPa (kpsi)	A	
ultimate strain	12 to 13 4 to 6	9 to 15 7 to 14	X Y	%	A	
Youngs' Modulus						ASTM D695 (0.05/min. strain rate)
under compression	1069 (155)	2144 (311)	Z	MPa (kpsi)	A	
ultimate stress	54 (7.9)	47 (6.9)	Z	MPa (kpsi)	A	
ultimate strain	33	25	Z	%		
Flexural Modulus	2634 (382) 1951 (283)	4364 (633) 3751 (544)	X	MPa (kpsi)	A	ASTM D790
ultimate stress	38 (5.5)	36 (5.2) 32 (4.4)	X Y	MPa (kpsi)	A	
Deformation under load	0.33 2.10	0.26 1.37	Z Z	%	24 hr/ 50°C/ 7MPa 24 hr/ 150°C/ 7 MPa	ASTM D261
Moisture Absorption	0.05	0.01		%	D48/50°C, 0.050" (1.27mm) thick	IPC-TM-650, 2.6.2.1
Density	2.7	3.1		g/cm³		ASTM D792
Thermal Conductivity	0.49	0.86		W/m/°K	80°C	ASTM C518
Thermal Expansion	47 34, 117	24 24,47	X Y,Z	ppm/°C	0 to 100°C	ASTM 3386 (5K/min)
Td	500	500		°C TGA		ASTM D3850
Specific Heat	0.97 (0.231)	1.00 (0.239)		J/g/K (BTU/lb/°F)		Calculated
Copper Peel	14.3 (2.5)	12.3 (2.1)		pli (N/mm)	after solder float	IPC-TM-650 2.4.8
Flammability Rating	V-0	V-0				UL94
Lead-Free Process Compatible	Yes	Yes				

[1] SI unit given first with other frequently used units in parentheses.
[2] Dielectric constant is based on .025 dielectric thickness, one ounce electrodeposited copper on two sides.
[3] The design Dk is an average number from several different tested lots of material and on the most common thickness/s. If more detailed information is required, please contact Rogers Corporation.
Refer to Rogers' technical paper "Dielectric Properties of High Frequency Materials" available at http://www.rogerscorp.com/acm.

Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

STANDARD THICKNESS	STANDARD PANEL SIZE	STANDARD COPPER CLADDING
0.005" (0.127mm) 0.010" (0.254mm) 0.025" (0.635mm) 0.050" (1.27mm) 0.075" (1.90mm) 0.100" (2.50mm) Non-standard thicknesses available	10" X 10" (254 X 254mm) 10" X 20" (254 X 508mm) *20" X 20" (508 X 508mm) - non-standard 18" X 12" (457 X 305 mm) *18" X 24" (457 X 610 mm) - non-standard (*note: the above 2 panel sizes are available in >0.025" only)	½ oz. (18 µm), 1 oz. (35µm), 2 oz. (70µm) electrodeposited & reverse treated EDC copper foil. Heavy metal claddings are available, based on dielectric thickness. Contact Rogers' Customer Service.

The information in this data sheet is intended to assist you in designing with Rogers' circuit materials. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit materials for each application.

These commodities, technology and software are exported from the United States in accordance with the Export Administration regulations. Diversion contrary to U.S. law prohibited.

Revised 1111 040915 Publication: #92-105