

A DISSERTATION ON  
**DESIGN OF LOW NOISE AMPLIFIER AT 2.4 GHz USING  
MICROSTRIP LINES FOR WIRELESS APPLICATIONS**

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award of the degree of

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**in**  
**Microwave & Optical Communication Engineering**

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# CERTIFICATE

This is to certify that the dissertation title “*Analysis and Design of Low Noise Amplifier at 2.4 GHz using Microstrip Lines for Wireless Applications*” is the authentic work of **Mr. Santosh Kumar Patel** under my guidance and supervision in the partial fulfillment of requirement towards the degree of Master of Technology in *Microwave and Optical Communication Engineering*, jointly run by the Department of Electronics and Communication Engineering and Department of Applied Physics at *Delhi Technological University, New Delhi*. To the best of my knowledge, the matter embodied in the thesis has not been submitted to any other University/ Institute for the award of any other degree.

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# DECLARATION

I hereby declare that all the information in these documents has been obtained and presented in accordance with academic rules and ethical conduct. It is being submitted for the degree of Master of Technology in Microwave and Optical Communication Engineering at Delhi Technological University. It has not been submitted before for any degree or examination in any other university.

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# ACKNOWLEDGEMENT

With all praises to the almighty and by His blessings I have finally completed this thesis.

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# ABSTRACT

Demand for WLAN system is increasing and keep improving with new technologies. One of the problems faced by WLAN system is noise caused by propagation through air, electron movement and interfacing signal that degrade the transceiver system performance. Low power wireless electronics is becoming more popular due to durability, portability and small dimension. Especially, electronic devices in instruments, scientific and medical (ISM) band is convenient from the spectrum regulations and technology availability point of view. In the communication engineering society, to make a robust transceiver is always a matter of challenges for the better performance.

Advanced Design System (ADS) tool was used for design and simulation, and each design was tuned to get the optimum value for noise figure, gain and input reflection coefficient. LNA stand-alone gives acceptable value of noise figure and gain but the bandwidth was too narrow compared to specification. Then, Low-Noise Amplifier was designed with Microstrip lines where the optimum value of noise figure and gain was found. The gain was almost flat over the whole band, i.e., 2.4-2.5 GHz compared to Low-Noise Amplifier designed with lumped components.

In this thesis work, a new approach of design and performance analysis of Low-Noise Amplifier is performed at 2.40 GHz. Performances of different designs are compared with respect to noise figure, gain, input and output reflection coefficient. In the design process, a single stage LNA using discrete component is designed with transistor, ATF-54143, and LNA with transistor ATF-54143 using Microstrip lines is designed with substrate material FR<sub>4</sub> having specification such as relative dielectric constant  $\epsilon_r = 4.4$ , thickness  $h = 0.8$  mm, permeability  $\mu_r = 1$ , metal conductivity  $Cond = 5.88e7$ , conductor thickness  $T = 0.035$  mm. After the simulation we got the simulated result of low noise amplifier as forward voltage gain( $S_{21}$ ) of 15.23dB, noise figure is 0.386dB, input reflection coefficient ( $S_{11}$ ) is -15dB, output reflection coefficient( $S_{22}$ ) is -10.26dB with discrete components and forward voltage gain( $S_{21}$ ) of 13.34dB, noise figure is 0.667dB, input reflection coefficient ( $S_{11}$ ) is -17dB, output reflection coefficient( $S_{22}$ ) is -13dB with Microstrip lines, stability factor ( $K_f$ ) for both circuit is more than 1 which indicates that amplifier is unconditionally stable.

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# LIST OF SYMBOLS

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$V_{out}$	Output Voltage
$V_{in}$	Input Voltage
$\Gamma_s$	Source Reflection Coefficient
$\Gamma_L$	Load Reflection Coefficient
$K$	Stability Factor
$\Delta f$	Bandwidth
$f_o$	Centre Frequency
$G_m$	Transconductance
$P_D$	Power Dissipation
$C$	Correlation coefficient
$Q$	Quality Factor
$L$	Chanel Length
$g_{do}$	Drain to Source Conductance
$\gamma$	Drain Noise Coefficient
$\delta$	Gate Noise Coefficient
$C_{gs}$	Gate to Source Capacitance
$W$	Width of MOSFET
$I_D$	Drain Current
$L$	Inductor
$C$	Capacitor
$G_T$	Transducer Gain
$G_P$	Operating Power Gain
$G_A$	Available Power Gain

# LIST OF ABBREVIATIONS

---

ISM	Instruments Scientific and Medical
WLAN	Wireless Local Area Network
LNA	Low-Noise-Amplifier
BPF	Band-Pass Filter
IMN	Input Matching Network
OMN	Output Matching Network
LPD433	Low Power Device, 433 MHz
PMR446	Private Mobile Radio, 446 MHz
ITU	International Telecommunication Union
ITU-R	The ITU Radio-communication Sector
FCC	Federal Communications Commission
WDCT	Digital Cordless Telecommunications
RFID	Radio Frequency Identification
HiperLAN	High Performance Radio LAN
Wi-Fi	Wireless Fidelity
PCB	Printed Circuit Board
PCS	Personal Communications Service
WCDMA	Wideband CDMA
ADS	Advanced Design System
WLL	Wireless Local Loop
SMD	Surface Mounted Device

# CHAPTER 1

## INTRODUCTION

### 1.1 OVERVIEW OF RF TECHNOLOGY

RF technology has changed quite a bit since the days of Marconi and Tesla-both the men who enabled radio communications. Modern radio frequency engineering is an exciting and dynamic field; due to the beneficial inter dependency between recent developments in electronic device technology and the increase in demand for voice, data, and video communication capacity. Prior to this revolution in communications, RF technology was the nearly exclusive domain of the defence industry but the recent increase in demand for communications systems with applications such as wireless paging, broadcast video, Bluetooth transceiver, Wi-Fi (WLAN), WiMax, CDMA, WCDMA, EGPRS, GSM and many more is revolutionizing the technology [1-2]. RF technology is important for these applications because these require high operational frequencies which allow both large numbers of independent channels as well as significant available bandwidth per channel for high speed communication. Figure 1.1 shows some disciplines that requires RF design.

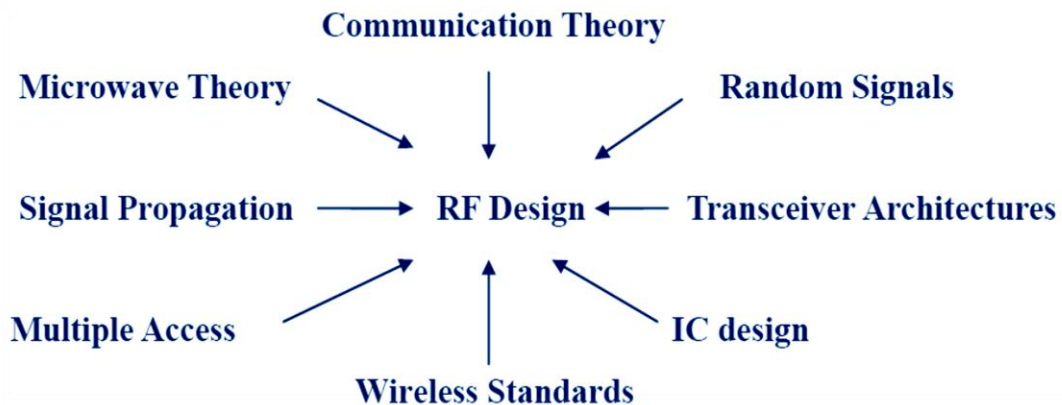


Figure 1.1: Disciplines requiring RF design [1]

The field of design in the electronics system that employs RF and Microwave Engineering includes the frequencies ranging from 300kHz to over 100GHz. RF

engineering refers to the circuits/device operating in the frequency range 300kHz to over 1GHz while the microwave engineering refers to the circuits/device operating in the frequency range between 300kHz or 1GHz to over 100GHz. Unfortunately, RFIC design is quite complex when tested results are obtained that differ drastically from the simulation result. The reasons for this disparity may normally be traced to one of the following [2]:

- The frequency of operation is such that the circuit's elements display complex behaviour, not represented by the pure element definitions utilized during the design.
- The circuit layout includes coupling paths not accounted for in the design.
- The ratio of transverse dimensions of transmission lines to wavelength is non-negligible thus, additional unwanted stored modes become available.
- The package that houses the circuit becomes an energy storage cavity, thus absorbing some of the energy propagating through it.
- The perfect (ideally) dc bias source is not adequately decoupled from the circuit.
- The degree of impedance match among interconnected circuits is not good enough, so that large voltage standing wave ratio (VSWR) is present, which give rise to inefficient power transfer and to ripples in the frequency response.

## **1.2 MOTIVATION**

Being the first block of the receiver chain LNA has to be able to receive the maximum signal without reflecting any part of it at the same time it has to ensure that it adds less noise to the signal component. The input impedance of the LNA should be  $50\Omega$  to match the source impedance, also the output stage should be able to give less noise, and more gain. Since any receiving block's performance depends mainly upon the LNA performance. It is the most important block for any communication system. So a LNA circuit with high gain and very low noise figure is highly needed for any modern receiver system, at the same time it should be less power consuming. In this work my aim to design a low noise amplifier at 2.4 GHz operating frequency for wireless

communication because low noise amplifier is first block of receiver circuit and it is important to give more concentration to achieving high gain and low noise figure.

### **1.3 OBJECTIVE OF RESEARCH**

The main objective of the research presented in this project is to design a low noise amplifier to be used in wireless communications operation. The goal of project is to understand the concepts of RF design such as gain, noise, matching networks, and linearity that are essential for amplifiers.

In the communication system, Low noise amplifier is the second element after antenna. LNA is used to boost up the received low energy and noisy signal to a desired energy level with noise suppression. So, noise figure (NF) is the key issue of concern in this design [3-5]. The receiver is the most power hungry block and the power consumption should be as low as possible. So, noise figure and power consumption are not less important issues than gain [6-7]. With the small power consumption, the LNA should amplify the weak receiving signal to the level suitable for processing and provide gain to overcome the noise of subsequent stages while adding small amount of noise as possible. Matching of each block of the receiver is also an important issue in order to provide maximum power transfer at a particular frequency the matching is required [8-9]. Furthermore, input and output matching to the source and load can maximize the gain. Input and output impedance matching is characterized by the input and output return loss. Since, it can affect the performance of the device. The gain should be large enough and the same time noise should be as less as possible [10]. The LNA should present specific impedance at the input, e.g.  $50\Omega$  to interface with the filter or antenna. Finding the delicate balance in all issues or parameters becomes the challenge more often than simply maximizing a single key parameter. The most recognizable trade-off is between LNA gain and noise figure (NF). Linearity is also an important design issue. Third order intercept point,  $IP_3$  (Third order intercept point) has emerged as an important parameter in LNA design. The easiest way to improve the  $IP_3$  performance for a given frequency is to increase the current density or current drawn of LNA. Input and output return losses,  $S_{11}$  and  $S_{22}$ , are also available for trading off to achieve improved gain and NF performance. Typically, the input and output matches are designed to afford good gain and NF performance. It is a balance of performance



parameters. The approach for achieving this target is to design a low noise amplifier for wireless application at 2.4 GHz operating frequency range.

## **1.4 THESIS STRUCTURE**

The thesis report is divided into six chapters, each having ample information for comprehending the concept of this project.

*Chapter 1* introduces the background history of RF technology and objective of research of low noise amplifier.

*Chapter 2* discussed the Low Noise Amplifier also deals the details about the research work carried out so far in the design of Low Noise Amplifier for different domain of RF.

*Chapter 3* provides the details about the different basic parameters of RF circuits design i.e. different type of Gain, noise in MOSFET, Noise Figure, Stability, linearity, Bandwidth etc.

*Chapters 4* presents different topology of Low Noise amplifier Design, based upon the Gain, Noise figure, and also discusses which topology is best suitable for Low Noise Amplifier.

*Chapter 5* discusses about the design, simulation, and analysis result of Low Noise amplifier, and some basic step for calculating the components parameters.

*Chapter 6* concludes the thesis with comparison of result to some reported works and then suggestion for further research has discussed.

# CHAPTER 2

## LOW NOISE AMPLIFIER AND LITERATURE REVIEW

### 2.1 DEFINITION OF LOW NOISE AMPLIFIER

The growth of wireless service and other telecom application has pushed the semiconductor industry towards complete system on chip solutions. Wireless system comprise of a front-end and a back end section. The front end section process analog signals in the high radio frequency range, while the back end section process analog and digital signals in the baseband low frequency range.

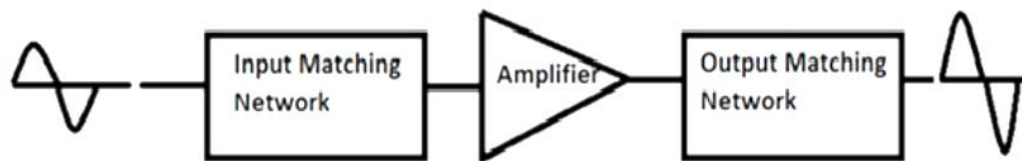


Figure 2.1: Basic low noise amplifier

The radio frequency signal received at the antenna is weak. Therefore an amplifier with high gain and good noise performance is needed to amplify this signal before it can be fed to others part of the receiver. Such an amplifier referred to as a low noise amplifier as shown in Figure 2.1. The first stage of the receiver is typically a low noise amplifier, whose main function is to provide enough gain to overcome the noise of subsequent stages (such as mixer). Aside from providing this gain while adding as little noise as possible, an LNA should accommodate large signals without distortion, and frequently must also present specific impedance. The general topology of low noise amplifier can be broken down into the three stages: input matching, amplifier and the output matching. Low noise amplifier used in various applications like ISM radio, cellular/PCS handsets, GPS receiver, cordless phone, wireless LAN, wireless data, satellite communication etc.

## 2.2 LITERATURE SURVEY

Lot of research is being currently done in designing low noise amplifier. Some of the important papers related to this thesis work have been divided into four sub groups:

1. Application Based LNAs
2. High Gain LNAs
3. Different Matching Techniques for LNAs
4. Low Noise Techniques for LNAs

The discussions on the literature belonging to the above subgroup are given below:

### 2.2.1 APPLICATION BASED LNAs

A low noise amplifier intended for use in a Global Positioning System (GPS) Receiver discussed by Derek K. Shaeffer et al. [11]. Inductive source degeneration topology has used for design of LNA.

A low noise amplifier, intended for use in a DECT (Digital Enhanced Cordless Communications) Receiver is presented by Jerome Le nyet *al.* [12], uses differential configuration with inductive source degeneration and constant  $g_m$  bias for design of LNA.

The concept of Concurrent Multiband Low Noise Amplifiers is presented by Hussein Hashemi et al. [13]. They discuss the advantages of field effect transistors over bipolar junction transistor and concurrent dual band low noise amplifier in communication systems.

A low-noise amplifier for use in Ultra wideband wireless receiver is introduced by Andrea Bevilacqua et al. [14]. In the design it is discussed why the low noise amplifier is essential for UWB receiver. In their design they used multi-section Chebyshev filter at the input network and buffer stage at the output stage for design of LNA.

A low noise amplifier, intended for use in Ultra wideband wireless receiver is discussed by Che-Cheng Huang<sup>1</sup> et al. [15] in which two stage cascode amplifiers with shunt-peaked load, two 2nd order notch filters, and an output buffer has been used for design of LNA.

A low noise amplifier intended for use in Ultra wideband wireless receiver front end is presented by Qiang Li et al. [16]. In which current-reuse topology, capacitive peaking is used for design of LNA. Inductors were not used in the design.

Zhe-Yang Huang et al. [17] has discussed low noise amplifier has been designed by using wideband input impedance matching network, a cascode amplifier with shunt-peaked load, a RLC impedance feedback loop. In another design Meng-Ping Chen et al. [18] used three LC-tank cascode amplifier and one output buffer for design of LNA.

A low-noise amplifier intended for use in WiMax system is discussed by Yu-Lin Wang [20]. They used differential, cascode configuration RC shunt feedback and shunt peak load for design of LNA.

A low noise amplifier intended for use in Ultra wideband Wireless Receiver is also provided by Shila Shamsadini et al. [21] with two-stage, common-gate in cascade with cascode for design of LNA. Table 2.1 shows the summary of application based LNA's.

**Table 2.1: Summary of Application Based LNA's**

Parameters	[11]	[12]	[14]	[15]	[16]	[18]	[20]	[21]
Year	1997	2002	2004	2007	2007	2008	2009	2010
Technology(um)	0.18	0.25	0.18	0.18	0.13		0.18	0.18
$S_{11}$ (dB)			-10	-6.4	-8.3	-11	-11	-10
$S_{22}$ (dB)				-10.1			-10.9	
$S_{21}$ (dB)	22	15	9.3	13.3	11	24.4	14.4	
Noise Figure(dB)	3.5	2	4	2.5	4.8	2.4	3.8	2.9
Power consumed(mw)	30	25	9	21.9	19	9.2	14.8	10

### 2.2.2 HIGH GAIN LNAs

Bevin G.Perumana et al. [19] had used inductorless resistive feedback with current reuse topology and tuned resistive feedback for increase of gain in low noise amplifier for multiband applications.

Shila Shamsadini et al. [21] presented two-stage, common-gate in cascade with cascode for increase in gain low noise amplifier for Ultra Wideband applications. T. HuiTeo et al. [22] used fully integrated differential configuration for increase of gain in low noise amplifier for short-range radio in biomedical devices.

Jenn-Tzer et al. [23] used differential configuration and high-Q active inductor for design of low noise amplifier for multiband application. Table 2.2 shows the summary of high gain LNA's

**Table 2.2: Summary of High Gain LNA's**

Parameters	[19]	[22]	[21]
Year	2008	2008	2010
Technology(um)	0.09	0.18	0.18
$S_{21}$ (dB)	22	26	25

### 2.2.3 DIFFERENT MATCHING TECHNIQUES FOR LNAs

Derek K. Shaeffer et al. [11], discussed the different matching topologies at the input. It includes resistive termination, 1/gm termination, shunt-series feedback, inductive degeneration. For designing LNA inductive degeneration is used because of its advantage of providing low noise figure, high linearity, and high gain.

Andrea Bevilacqua et al. [14] used an inductively degenerated common-source amplifier by embedding the input network of the amplifying device in a multi-section reactive network so that the overall input reactance is resonated over a wider bandwidth.

Zhe-Yang Huang et al. [17], used wideband input impedance matching network, a cascade amplifier with shunt-peaked load, a RLC impedance feedback loop and an output buffer for input and output matching.

Meng-Ping Chen et al. [18], used input impedance matching network that included the conventional source degeneration input matching and an inductor shunted in a shunted resistor in RF signal path. The shunted resistor in RF path saved the three inductors used for design of Butterworth filter required earlier for the design in specified application. Table 3 shows the summary of different matching techniques for LNA's.

Michael T. Reiha et al. [24], presented modified cascode in a negative feedback configuration. Negative feedback offer tailor port impedances for noise and impedance matching.

**Table 2.3: Summary of Different Matching Techniques for LNA's**

<b>Parameters</b>	<b>[11]</b>	<b>[14]</b>	<b>[17]</b>	<b>[18]</b>	<b>[24]</b>
Year	1997	2004	2007	2008	2007
Technology(um)	0.6	0.18	0.18	0.18	0.13
$S_{11}$ (dB)		-10	-10.1	-6.8	-44
$S_{22}$ (dB)		-7	-9.76	-9.5	-12

#### **2.2.4 LOW NOISE TECHNIQUES FOR LNAs**

Derek K. Shaeffer et al. [11], had discussed the different matching topologies at the input. It includes resistive termination,  $1/g_m$  termination, shunt-series feedback, inductive degeneration. For designing they used inductive degeneration because of its advantage of providing low noise figure, high linearity, and high gain.

Michael T. Reiha et al. [24], presents the reactive (negative) feedback for design of low noise amplifier, reactive feedback reduces the noise figure by reducing the input insertion loss, flat group delay and lowering input referred noise.

Andries J. Scholten et al. [25], provided the noise modelling for RF CMOS circuits that included effect of drain current noise and gate noise for short channel MOSFETs. The results was modelled with a non-quasi- static RF model, based on channel segmentation, which is capable of predicting both drain and gate current noise accurately. Two additional noise mechanisms: avalanche noise associated with the avalanche current from drain to bulk and shot noise in the direct-tunnelling gate leakage current is also discussed.

Xiao-dong Wang et al. [26], added the capacitor between the gate and the source of the input MOSFET in cascode low noise inductive degeneration topology for reducing the noise figure in amplifiers.

Jenn-Tzer Yang et al. [17], used high-Q active inductor and differential configuration for lowering the noise figure. Table 4 shows the summary of low noise techniques for LNA's.

**Table 2.4: Summary of Low Noise Techniques for LNA's**

<b>Parameters</b>	<b>[11]</b>	<b>[24]</b>	<b>[26]</b>	<b>[17]</b>
Year	1997	2007	2008	2007
Technology(um)	0.18	0.13	0.18	0.18
NF(dB)	3.5	.43	.655	0.55 to 0.758

### **2.3 PROBLEM DEFINITION**

Design of Low Noise Amplifier for wireless communication with frequency range 2.4 GHz and some specific target has to be achieving i.e.

- (i) Gain should be greater than 15dB.
- (ii) Noise Figure should be lower than 2.0dB.
- (iii) Stability factor must be greater than 1.
- (iv) Power consumption should be low.

# CHAPTER 3

## BASIC PARAMETERS OF LNA DESIGN

### 3.1 TWO PORT NETWORK

The abstraction of the analog block that is to be designed into a two port network is always useful in designing. So the characteristics and parameters of two port representation will be discussed. In a two port network, two port and four terminals are used to represent the characteristics of the circuit, especially in terms of current and voltage relations. The condition for any particular two terminals to be considered as a port is that current entering through one terminal must be equal to the current leaving through other terminal. And each port, two variables will be there out of them one will be independent and one will be dependent. At low frequency the two variables are voltage and current and the common two port representation are impedance matrix (Z parameter) and admittance matrix(Y parameter).

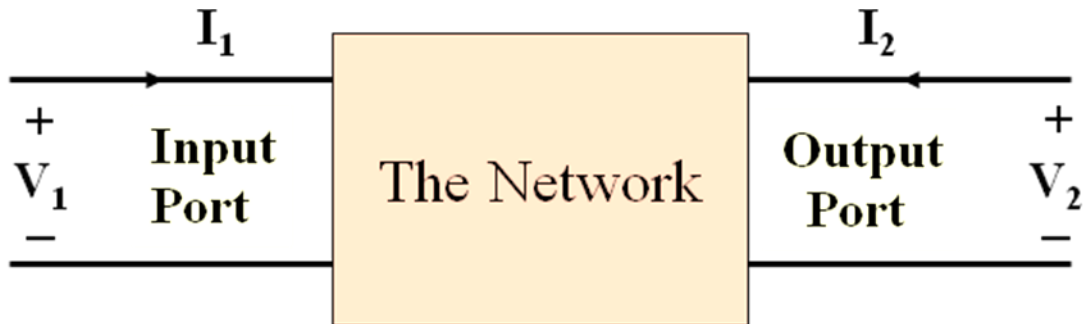


Figure 3.1: Two port network representation

The above figure shows a typical 2-port network, the impedance and admittance matrix are defined as shown in equations (3.1).and (3.2).

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} z_{11} & z_{21} \\ z_{21} & z_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ i_2 \end{bmatrix} \quad (3.1)$$

$$\begin{bmatrix} i_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{21} \\ y_{21} & y_{22} \end{bmatrix} \begin{bmatrix} v_1 \\ v_2 \end{bmatrix} \quad (3.2)$$



But it is to be noted that the concept of Y and Z parameters are useful only at low frequency because at low frequency these parameters can be determined by applying either current or voltage at one port and measuring other variable by shorting and opening other port respectively, but at very high frequency it will be very difficult to open or short any port because it needs a broadband match at high frequency and also an active two port network might oscillate if one of its port is open or short circuited. That is why a new parameter has to be defined at RF, this parameter is called scattering parameter.

### 3.1.1 SCATTERING PARAMETERS

Since the main problem with RF was that one has to rely on open and short circuiting the other port if one has to define Y and Z parameters for RF circuits therefore scattering parameter has the advantage that they can be measured by matching source and load impedance to the reference impedance. To represent a two-port network at microwave frequencies, scattering parameters(S -parameters) can be used. S-parameters themselves ( $S_{11}$ ,  $S_{12}$ ,  $S_{21}$ , and  $S_{22}$ ) represent reflection and transmission coefficients of the two-port under certain “matched” conditions.  $S_{11}$  is the reflection coefficient and  $S_{21}$  is the transmission coefficient at port 1 when port 2 is terminated in a load whose impedance is equal to that of the transmission line characteristic impedance. Likewise  $S_{22}$  is the reflection coefficient and  $S_{12}$  is the transmission coefficient at port 2 when port 1 is terminated in a matched load. The S parameter representation for two port network is as shown in Figure (3.2).

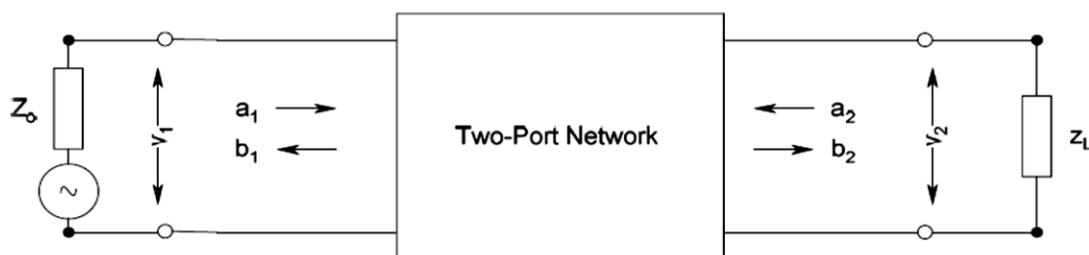


Figure 3.2: Two port network showing incident and reflected wave

Here in the above figure  $a_1$  is the normalized incident wave at port and  $b_1$  is the normalized reflected wave at the same port,  $v_1$  and  $i_1$  are voltage at port 1 and current

entering into port 1. And the  $a_1$ ,  $b_1$ , the terminal current and terminal voltage relation is expressed in Equations (3.3) and (3.4)

$$a_i = \frac{v_i + z_0 i_i}{2\sqrt{z_0}} \quad (3.3)$$

$$b_i = \frac{v_i - z_0 i_i}{2\sqrt{z_0}} \quad (3.4)$$

Here  $Z_0$  is the reference impedance and it is taken as  $50 \Omega$  for all practical application. The overall scattering parameter in the matrix form can expressed as shown in Equation (3.5)

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \quad (3.5)$$

But it is to be noted that the concept of Y and Z parameters are useful only at low frequency because at low frequency these parameters can be determined by applying either current or voltage at one port and measuring other variable by shorting and opening other port respectively, but at very high frequency it will be very difficult to open or short any port because it needs a broadband match at high frequency and also an active two port network might oscillate if one of its port is open or short circuited. That is why a new parameter has to be defined at RF, this parameter is called scattering parameter.

By expanding the scattering matrix the value of  $S_{11}$ ,  $S_{12}$ ,  $S_{21}$  and  $S_{22}$  can be expressed as shown below:

$$s_{11} = \frac{b_1}{a_1} \quad (3.6)$$

$$s_{12} = \frac{b_1}{a_2} \quad (3.7)$$

$$s_{21} = \frac{b_2}{a_1} \quad (3.8)$$

$$s_{22} = \frac{b_2}{a_2} \quad (3.9)$$

$S_{11}$  is input reflection coefficient,  $S_{12}$  is reverse transmission coefficient,  $S_{21}$  is forward transmission coefficient,  $S_{22}$  output reflection coefficient. Clearly  $S_{11}$  is the ratio of reflected wave to the incident wave at port 1 when port 2 is perfectly matched. Perfectly matched means the load impedance is equal to the value of characteristic impedance. Similarly  $S_{21}$  is the ratio of reflected wave at port 2 to the incident wave at port 1 when port 2 is properly matched, in other way it can be said that the ratio of value of voltage received at port 2 to the voltage on port 1 is  $S_{21}$ , so it is the Gain of the circuit.

### 3.2 GAIN OF TWO PORT NETWORKS

For RF circuits designers prefer to express Gain in term of power, for any circuit the purpose is to maximize Gain and in order to get maximum gain the most important concept is that there should not be any power reflection means to say that all incident power should be applied to the circuit. The basic principle of this concept is based on maximum power transfer theorem, so it needs to be explained.

#### 3.2.1 MAXIMUM POWER TRANSFER THEOREM

It is known that in order to the maximum power delivered by source going into the load the load impedance should be equal to the source impedance i.e.  $R_s=R_L$ . For the case of complex load it can be explained as follows –

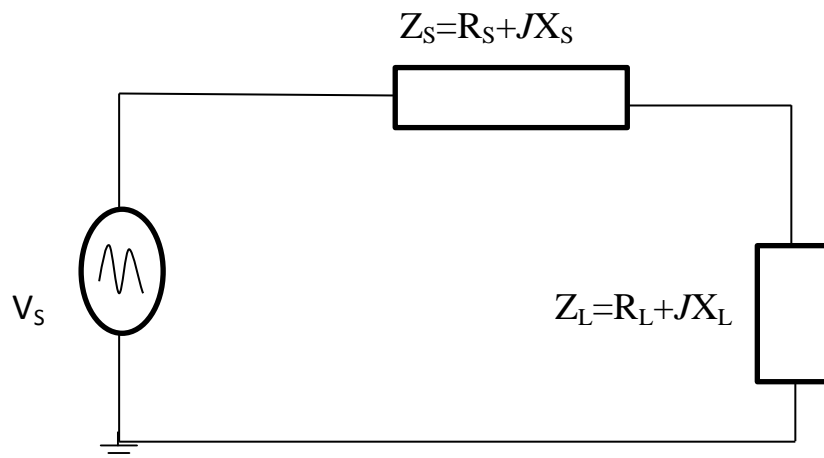


Figure 3.3: Source having complex impedance connected with complex load

From the above Figure (3.3), power delivered to the load  $R_L$  is given as-

$$P_L = \frac{|V_{RL}|^2}{R_L} = \frac{R_L |V_S|^2}{|Z_L + Z_S|^2} \quad (3.10)$$

Now separating the real and imaginary part of Equation (3.10) we have

$$P_L = \frac{|V_{RL}|^2}{R_L} = \frac{R_L |V_S|^2}{(R_L + R_S)^2 + (X_L + X_S)^2} \quad (3.11)$$

Now by differentiating the above equation with respect to  $X_S$  and  $R_S$  and equating to zero, the condition for maximum power transfer can be achieved. The condition for maximum power transfer is  $R_L$  should be equal to  $R_S$  and  $X_L + X_S$  should be equal to zero. In other words it can be said that for maximum power transfer the real impedance of load should be equal to the real impedance of the source and summation of imaginary part load impedance and source impedance must be zero. So for maximum power transfer the load impedance  $Z_L$  should be the complex conjugate of source impedance  $Z_S$ . Also it is to be noted that even in case of maximum power source will deliver only half of the available power to the load.

### 3.2.2 DIFFERENT GAIN EXPRESSION

The gain of the device is its ability to amplify the amplitude or the power of the input signal. It is defined as the ratio of the output to the input signal and is often referred to in terms of decibels.

$$\text{Voltage Gain} = 20 \log \frac{V_{out}}{V_{in}}$$

Power gain is generally defined as the ratio of the power actually delivered to the load to the power actually delivered by the source. Three power gains are commonly used in LNA design.

- (1)  $G_T$ , transducer power gain
- (2)  $G_P$ , operating power gain
- (3)  $G_A$ , available power gain

Besides these three gain definitions, there are three additional gain definitions we can use to evaluate the LNA design.

- $G_{umx}$ , maximum unilateral transducer power gain
- $G_{max}$ , maximum transducer power gain
- $G_{msg}$ , maximum stability gain

Besides these six gain definitions, there are two gain circles that are helpful to the design of input and output matching networks.

- GPC, power gain circle
- GAC, available gain circle

### (1) Transducer Power Gain

Transducer power gain,  $G_T$  is defined as the ratio between the power delivered to the load and the power available from the source [27].

$$G_T = \frac{1 - |\Gamma_s|^2}{|1 - S_{11}\Gamma_s|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - \Gamma_{out}\Gamma_L|^2} \quad (3.12)$$

The transducer gain expressions are too complex for manual design. To effect an approximate gain solution, let us ignore the feedback, that is assume that  $S_{12} = 0$ . If an amplifier has no feedback, signals pass one way through it. Accordingly this is called unilateral gain. If  $S_{12} = 0$ , then  $\Gamma_{in} = S_{11}$  and  $\Gamma_{out} = S_{22}$  and transducer gain becomes the unilateral gain.

### (2) Operating Power Gain

Operating power gain,  $G_P$  is defined as the ratio between the power delivered to the load and the power input to the network.

$$G_A = \frac{1 - |\Gamma_s|^2}{|1 - S_{11}\Gamma_s|^2} |S_{21}|^2 \frac{1}{1 - |\Gamma_{out}|^2} \quad (3.13)$$

The LNA scattering matrix is normalized in terms of the source and load resistance in Equation (3.14).

$$\Gamma_s = \Gamma_L = 0 \quad (3.14)$$

Thus, the input and output reflection coefficient are simply expressed in terms of Equation (3.15) and (3.16)

$$\Gamma_{in} = S_{11} \quad (3.15)$$

$$\Gamma_{out} = S_{22} \quad (3.16)$$

From these equations,

$$G_P = \frac{1}{1-|S_{11}|^2} |S_{21}|^2 \quad (3.17)$$

### (3) Available Power Gain

Available power gain,  $G_A$  is defined as the ratio between the power available from the network and the power available from the source [27].

$$G_A = \frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1}{1-|\Gamma_{out}|^2} \quad (3.18)$$

From Equations 3.15 and 3.16

$$G_A = |S_{21}|^2 \frac{1}{1-|S_{22}|^2} \quad (3.19)$$

The power available from the source is greater than the power input to the LNA network,  $G_P > G_T$ . The closer the two gains are the better the input matching. Similarly, because the power available from the LNA network is greater than the power delivered to the load,  $G_A > G_T$ .

### (4) Maximum Unilateral Transducer Power Gain

Maximum Unilateral transducer power gain  $G_{umx}$  is the transducer power gain when we assume that the reverse coupling of the LNA,  $S_{12}$  is zero, and the source and load impedances are conjugate matched to the LNA. That is  $\Gamma_S=S_{11}$  and  $\Gamma_L=S_{22}$ . If  $S_{12} = 0$  and the input and output reflection coefficient are  $\Gamma_{in}=S_{11}$  and  $\Gamma_{out}=S_{22}$ . Thus from Equation 3.13 we get Equation 3.20.

$$G_{umx} = \frac{1}{|1-|S_{11}|^2|} |S_{21}|^2 \frac{1}{|1-|S_{22}|^2|} \quad (3.20)$$

### (5) Maximum Transducer Power Gain

Maximum transducer power gain,  $G_{max}$  is the simultaneous conjugate matching power gain when both the input and output are conjugate matched. That is  $\Gamma_S=S_{11}$  and  $\Gamma_L=S_{22}$ . When the reverse coupling,  $S_{12}$  is small,  $G_{umx}$  is close to  $G_{max}$ .

$$G_{max} = \frac{|S_{21}|}{|S_{12}|} (k - \sqrt{k^2 - 1}) \quad (3.21)$$

Where, K is stability factor

### (6) Maximum Stability Gain

Maximum Stability gain,  $G_{msg}$  is the maximum of  $G_{max}$  when the stability condition  $K > 1$  is still satisfied.

$$G_{msg} = \frac{|S_{21}|}{|S_{12}|} \quad (3.22)$$

### (7) Power Gain Circle

Power gain circle in short form is *GPC*. From Equation 3.13,  $G_p$ , is solely a function of the load reflection  $\Gamma_L$ . Thus we can draw power gain contours on the smith chart of  $\Gamma_L$ . The location for the peak of the contour corresponds to  $\Gamma_L$  producing the maximum  $G_p$ . We can move the peak location by changing the design of the output matching network. The best location for the contour peak is at the center of the smith chart that is where  $\Gamma_L = 0$ .

### (8) Available Gain Circle

Available gain circle in short form is *GAC*. From Equation 3.18, we can see that  $G_A$  is solely a function of the source reflection  $\Gamma_S$ . Thus we can draw available gain contours on the smith chart of  $\Gamma_S$ . The location for the peak of the contour corresponds to  $\Gamma_S$  producing the maximum  $G_A$ . we can move the peak location by changing the design of the input matching network. The best location for the contour peak is at the center of the smith chart that is where  $\Gamma_S = 0$ .

## 3.3 STABILITY

In the presence of feedback paths from the output to input, the circuits might become unstable for certain combination of source and load impedances. An LNA design that is normally stable might oscillate at the extremes of the manufacturing or voltage variations, and perhaps at unexpectedly high or low frequencies. The stability factor is given as in Equation 3.23

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} \quad (3.23)$$

Where

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| \quad (3.24)$$

Where  $K > 1$  and  $\Delta < 1$ , the circuit is unconditionally stable. That is, the circuit does not oscillate with any combination of source and load impedance. We should perform the

stability evaluation for the S parameters over a wide frequency range to ensure that  $K$  remains greater than one at all frequencies. As the coupling ( $S_{12}$ ) decrease, which is as reverse isolation increase, stability improves. Techniques such as resistive loading and neutralization are used to improve stability for LNA. Equation 3.23 is valid for small signal stability. If the circuit is unconditionally stable under small signal conditions, the circuit is less likely to be unstable when the input signal is large. Aside from the two metrics  $K$  and  $\Delta$ , the source and load stability circles can be used to check LNA stability.

### **3.4 NOISE**

Noise is a random process means the value of the noise cannot be predicted at any time if the past value is known. Instantaneous value of noise cannot be predicted, a statistical model provides knowledge about some other important properties of the noise i.e. average power of noise that will be predictable.

#### **3.4.1 NOISE IN MOSFET**

Design of LNA using CMOS will be discussed, therefore we will discuss the main source of noise and their expression in order to be able to express them both qualitatively and quantitatively in the circuits. Several kinds of noise are discussed below

##### **(i) Thermal Noise**

It is the random motion of electrons in a conductor introduces fluctuations in the voltage measured across the conductor even if the average current is zero. The thermal noise is proportional to the absolute temperature.

##### **(ii) Flicker Noise**

The interface between the gate oxide and the silicon substrate in a MOSFET entails an interesting phenomenon. Since silicon crystal reaches an end at this interface, many “dangling” bonds appear, giving rise to extra energy states. As charge carriers move at the interface, some are randomly trapped and later released by such energy states, introducing “flicker noise” in the drain current. Since LNA design for radio frequency and flicker noise is inversely proportional to frequency so it will be ignored. Noise due to gate resistance- the charge present in the channel fluctuates and it is capacitively coupled to gate and hence noise is produced and given as [28].



$$\overline{\hat{i}_{ng}^2} / \Delta f = 4kTg_g \delta \quad (3.25)$$

Channel of MOSFET generates thermal noise and is given as-

$$\overline{\hat{i}_{ng}^2} / \Delta f = 4kTg_{do} \delta \quad (3.26)$$

Where  $g_g$  is the gate conductance,  $g_{do}$  is open channel conductance [26].

### 3.4.1.1 Input equivalent noise source

In order to able to explain the behaviour of the circuit a noisy circuit is expressed by a noiseless circuit and two sources at input.

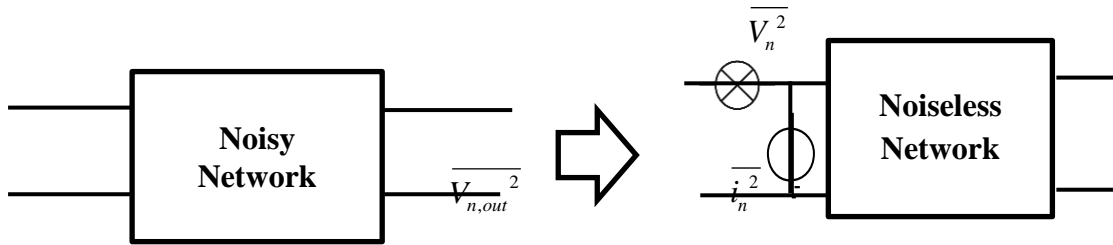


Figure 3.4: Input equivalent noise source

Two noise sources one current and one voltage source is necessary because if  $R_s$  is zero(short) the noise current will flow out and only noise voltage will remain and similarly if  $R_s$  is infinite(open) only noise current will affect the circuit.

### 3.4.1.2 Input equivalent noise source for MOS

As shown in figure 3.5 input equivalents for noise source-

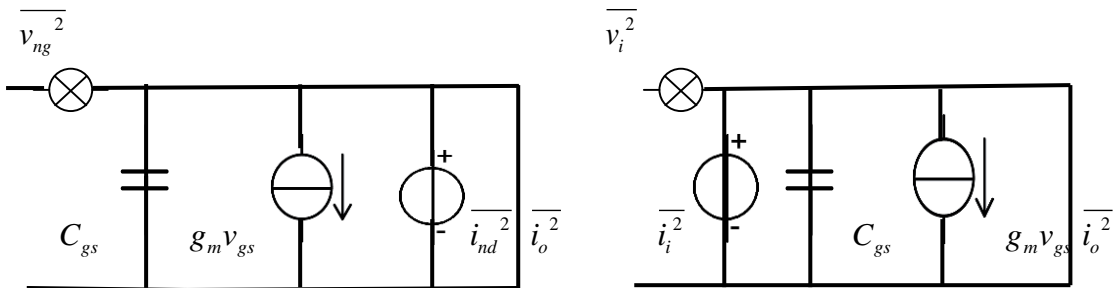


Figure 3.5: Input equivalent noise sources for MOS

By short circuiting the output of the actual circuit and equivalent circuit the relation between the equivalent current and voltage source with drain and gate noise current will be determined [27].

Case 1: when input is short circuited-

$$i_0 = g_m V_{ng} + i_{nd} \quad (3.27)$$

It should be equal to the value of  $i_o$  from equivalent circuit i.e.

$$i_0 = g_m V_i \quad (3.28)$$

By equating the above two equations, Equation (3.29) is obtained.

$$V_i g_m = V_{ng} g_m + i_{nd} \quad (3.29)$$

Therefore it can be written as-

$$\overline{V_i^2} = \overline{V_{ng}^2} + \overline{i_{nd}^2} \left(\frac{1}{g_m}\right)^2 \quad (3.30)$$

Case 2: when input is open circuited-

$$i_o = i_{nd} \quad (3.31)$$

And it should be equal to

$$i_o = \left(\frac{i_i}{w_c g_s}\right) g_m \quad (3.32)$$

From the above two equations, it can also be written as-

$$\hat{i}_i^2 = \left(\frac{w_c g_s}{g_m}\right)^2 \hat{i}_{nd}^2 \quad (3.33)$$

So from Equations (3.30) and (3.33) equivalent noise source are defined, and it is to be noted that both the noise sources are correlated.

### 3.4.1.3 Absolute value of equivalent noise source

The correlation coefficient of the two thermal noise source is given by Equation (3.34)

$$c = \frac{\overline{\hat{i}_{ng} \cdot \hat{i}_{nd}^*}}{\sqrt{\hat{i}_{ng}^2} \sqrt{\hat{i}_{nd}^2}} \quad (3.34)$$

And value of  $c$  for long channel is  $0.395j$  [5].

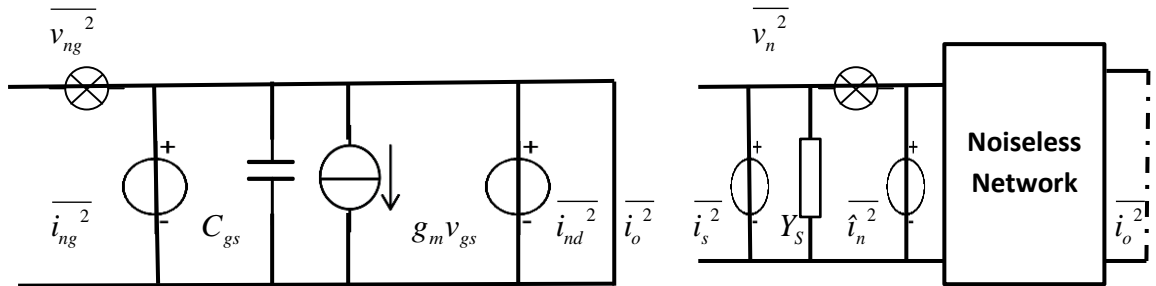


Figure 3.6: Noisy MOS replaced by noiseless source with input noise

Now again shorting the input and forcing equal current io in both the circuits, therefore

$$\overline{V_n^2} = \frac{\overline{\hat{i}_{nd}^2}}{g_m^2} = \frac{4kT\gamma g_{do}\Delta f}{g_m^2} \quad (3.35)$$

It is clear from the above equation that gate thermal noise does not contribute to this. The noise generated in Equation (3.35) can be expressed by the noise generated by a resistor of value  $R_n$  therefore

$$\overline{V_n^2} = 4KT\Delta f R_n \quad (3.36)$$

Where

$$R_n = \frac{\gamma g_{do}}{g_m^2} \quad (3.37)$$

$\overline{\hat{i}_{ng}^2}$  Will have contribution due to gate thermal noise  $\overline{\hat{i}_{ng}^2}$  and drain thermal noise  $\overline{\hat{i}_{nl}^2}$

$$\overline{\hat{i}_{nl}^2} = \overline{\hat{i}_{nd}^2} \frac{(\omega c_{gs})^2}{g_m^2} = \frac{4kT\gamma g_{do} (\omega c_{gs})^2 \Delta f}{g_m^2} \quad (3.38)$$

From Equation 3.35

$$\overline{\hat{i}_{nl}^2} = \overline{v_{nl}^2} (\omega c_{gs}) \quad (3.39)$$

$\overline{\hat{i}_{ng}^2}$  has two components one correlated with  $\overline{v_n^2}$  and other not correlated with it

$$\overline{\hat{i}_{ng}^2} = \overline{(i_{ngc} + i_{ngu})^2} = 4KT\delta g_g \Delta f |c|^2 + 4KT\delta g_g \Delta f (1 - |c|^2) \quad (3.40)$$

Therefore total equivalent noise current can be written into two parts-

(i)Uncorrelated term

$$\overline{\hat{i}_{nu}^2} = \overline{\hat{i}_{ngu}^2} \quad (3.41)$$

It can be represented by noise generated by Yu and value is given by-

$$Y_u = G_u = \frac{\overline{\hat{i}_{nu}^2}}{4KT\Delta f} = \delta g_g (1 - |c|^2) \quad (3.42)$$

So

$$Y_u = G_u = \frac{\delta \omega^2 c_{gs}^2 g_g (1 - |c|^2)}{5g_{do}} \quad (3.43)$$

(ii)Correlated term

$$\overline{\hat{i}_{nc}^2} = \overline{\hat{i}_{ngc}^2} \quad (3.44)$$

It can be represented by noise generated by  $Y_c$  and is given by the following equation.

$$Y_c = \frac{i_{nc}}{V_n} = \frac{g_m i_{nc}}{i_{nd}} + j\omega C_{gs} \quad (3.45)$$

Therefore, the model of MOS transistor as a noiseless system with two input noise sources is obtained and the noise sources can be characterized completely by  $Y_u$  and  $Y_c$ .

### 3.5 NOISE FIGURE

Noise factor is defined as the ratio of signal to noise ratio at input to signal to noise ratio at output and also defined as the ratio of total output noise power to the output noise power due to source alone[29]. Decibel representation of noise factor is known as noise figure. Generally signal to noise ratio is the parameter that is being used for analog circuits but in RF design since the ultimate aim is to maximize the SNR therefore noise figure is defined [29].

$$\text{Noise factor} = \text{SNR}_{\text{in}} / \text{SNR}_{\text{out}}$$

And

$$\text{Noise figure} = 10 \log_{10}(\text{Noise factor})$$

Now the expression for the noise factor of MOS transistor will be derived and also the condition for the noise factor to be minimum will be determined.

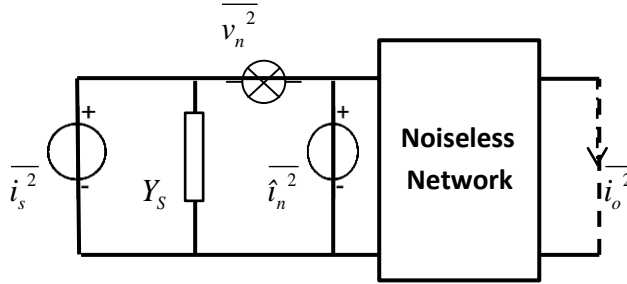


Figure 3.7: Noiseless network with equivalent noise current and voltage [30]

From the above circuit noise figure is-

$$F = \frac{\overline{\hat{i}_s^2} + \overline{(\hat{i}_n + Y_S V_n)(\hat{i}_n^* + Y_S^* V_n^*)}}{\overline{\hat{i}_n^2}} \quad (3.46)$$

It is known that the equivalent input current  $i_n$  is composed of two parts one is correlated with  $V_n$  and other is correlated with it. So considering correlated and uncorrelated terms-

$$i_n = i_{nu} + i_{nc} = i_{nu} + Y_c V_n \quad (3.47)$$

Therefore,

$$F = \frac{\overline{\hat{i}_s^2 + (i_{nu} + (Y_c + Y_s)V_n (\hat{i}_{nu}^* + (Y_c^* + Y_s^*)V_n^*))}}{\hat{i}_s^2} \quad (3.48)$$

$$F = \frac{\overline{\hat{i}_s^2 + (i_{nu} + (Y_c + Y_s)V_n (\hat{i}_{nu}^* + (Y_c^* + Y_s^*)V_n^*))}}{\hat{i}_s^2} \quad (3.49)$$

And using above equation 3.49,

$$F = \frac{G_u + [(G_c + G_s)^2 + (B_c + B_s)^2] R_n}{G_s} \quad (3.50)$$

Here  $G_c + jB_c$  is used for  $Y_c$  and  $G_s + jB_s$  for  $G_s$ .

For MOS  $G_c = 0$  and above equation can be written as-

$$G_u = \frac{\delta c_{gs}^2 \omega^2 (1 - |c|^2)}{5g_{do}}$$

$$B_c = \omega c_{gs} \left( 1 + \alpha |c| \sqrt{\frac{\delta}{5\gamma}} \right)$$

And

$$R_n = \frac{\gamma}{\alpha g_m}$$

These values are putting in Equation 3.50

$$F = 1 + \frac{\frac{\delta \alpha \omega^2 c_{gs} (1 - |c|^2)}{5g_m} + \left( \omega c_{gs} \left( 1 + \alpha |c| \sqrt{\frac{\delta}{5\gamma}} \right) + B_s \right)^2 \frac{\gamma}{\alpha g_m}}{G_s} \quad (3.51)$$

Now in order to optimize this noise figure, Equation 3.50 will be differentiated with respect to  $B_s$  and  $G_s$ , because these are the values which are coming from source and these values will have to be optimized in order to get minimum noise figure. So by taking differentiation with respect to  $B_s$  we have-

$$B_{s\_opt} = -B_c = -\omega c_{gs} \left( 1 + \alpha |c| \sqrt{\frac{\delta}{5\gamma}} \right) \quad (3.52)$$

Also by differentiating with respect to  $G_s$ -

$$G_{s\_opt} = \sqrt{\frac{G_u}{R_n}} = \alpha \omega c_{gs} \left( \sqrt{\frac{\delta (1 - |c|^2)}{5\gamma}} \right) \quad (3.53)$$

By using Equations 3.52 and 3.53

$$F_{\min} = 1 + \frac{2\omega}{\sqrt{5}w_T} \sqrt{\gamma\delta(1-|c|^2)} \quad (3.54)$$

This is the expression for minimum noise figure.

Also from the maximum power transfer theorem the condition for maximum power transfer and the condition for minimum noise figure cannot be satisfied simultaneously. Therefore one will have to always look for the trade off in between noise and power gain.

### 3.6 NOISE OPTIMIZATION METHODS

In low noise amplifier design, determination of the minimum noise figure is a common and Well-understood procedure. Typically, a small-signal model of the amplifier is assumed, an expression for F is formed, and differentiation leads to the unique conditions for optimized noise performance [30]. In the noise optimization techniques we seek the conditions that guarantee optimized noise performance for a specified fixed design parameter, such as gain or power consumption, under the condition of perfect input matching. Now, we fix the necessary design criteria and determine the appropriate small-signal model through the optimization procedure. Because the architecture permits selection of  $Q_L$  and  $L_S$  independently, so we can optimize the noise performance that coincides with the input match. There are two approaches for the optimization of noise figure. The first assumes a fixed transconductance,  $G_m$ , for the amplifier. The second assumes fixed power consumption. We know the equation of noise figure is [31].

$$F = 1 + \frac{R_L}{R_s} + \frac{R_g}{R_s} + \gamma\chi g_{do} R \left( \frac{\omega}{\omega_T} \right)_s^2 \quad (3.55)$$

To illustrate these approaches the expression for the F in Equation 3.55 can be recast to make its dependence on power dissipation. From equation 3.56 it is clear that condition for constant  $G_m$  is equivalent to the condition of constant  $\omega_T$ .

$$G_m = g_{m1} Q_{in} = \frac{g_{m1}}{\omega c_{gs} (R_s + w_T L_s)} = \frac{\omega_T}{\omega R_s \left( 1 + \frac{w_T L_s}{R_s} \right)} = \frac{\omega_T}{2wR_s} \quad (3.56)$$

To maintain a fixed  $\omega_T$ , we need only fix the value of  $\rho$ . Hence, we will reformulate F in terms of  $P_D$  and  $\rho$  to facilitate both optimizations. The equation of F in terms of  $P_D$  and  $\rho$  is given below [31].

$$F = 1 + \frac{\gamma\omega L}{3v_{sat}} P(\rho, P_D) \quad (3.57)$$

In this equation we have neglected the contribution of the gate resistance and inductor losses to the noise factor. In the Equation 3.57 ( $\rho$ ), is a ratio of two sixth order polynomials of  $\rho$  given by

$$P(\rho, P_D) = \frac{\frac{P_D}{P_0} P_1(\rho) + P_2(\rho) + \frac{P_0}{P_D} P_3(\rho)}{\rho^3 \left(1 + \frac{\rho}{2}\right)^2 (1 + \rho)} \quad (3.58)$$

Where,

$$P_1 = (1 + \rho)^2 + \frac{\delta}{5\gamma} (1 + \rho)^2 \left(1 + \frac{\rho}{2}\right)^2$$

$$P_2(\rho) = 2|c| \sqrt{\frac{\delta}{5\gamma}} (1 + \rho)^3 \left(1 + \frac{\rho}{2}\right)^2$$

$$P_3(\rho) = \frac{\delta}{5\gamma} \left(1 + \frac{\rho}{2}\right)^2 \rho^4$$

Because F is a function of two variables, one can define contours of constant noise figure in  $\rho$  and  $P_D$ . Equation (3.57) suggests that optimization of proceeds by minimizing with respect to one of its arguments, keeping the other one fixed.

### 3.6.1 FIXED $G_M$ OPTIMIZATION

To fix the value of the transconductance,  $G_m$ , we need only assign a constant value to  $\rho$ . We know,

$$\omega_T \approx \frac{g_m}{c_{gs}} = \frac{g_m}{\frac{2}{3} w L c_{ox}} = \frac{3}{2} \frac{\alpha \mu_{eff} V_{od}}{L^2} = \frac{3\alpha \rho v_{sat}}{L} \quad (3.59)$$

The appropriate value for  $G_m$  is easily determined by substituting Equation (3.59) into the expression for as found in Equation (3.56). The value of  $G_m$  which relates to  $\rho$ , is

$$G_m = \frac{2v_{sat}\rho\left(1 + \frac{\rho}{2}\right)}{2\omega R_s L(1 + \rho)^2} \quad (3.60)$$

Once  $\rho$  is determined, we can minimize the noise factor by taking-

$$\frac{\delta P(\rho, P_D)}{\delta P_D} = 0$$

Which, after some algebraic manipulations result in

$$P_{D,opt,G_m} = P_0 \sqrt{\frac{P_3(\rho)}{P_1(\rho)}} = P_0 \frac{\rho^2}{1 + \rho} \left[1 + \frac{5\gamma}{\delta\alpha^2}\right]^{\frac{1}{2}} \quad (3.61)$$

This expression gives the power dissipation which yields the best noise performance for a given  $G_m$  under the assumption of matched input impedance. We know  $Q_L$  is:

$$Q_L = \frac{P_0}{P_D} \frac{\rho^2}{1 + \rho} \quad (3.62)$$

By comparing Equations (3.61) and (3.62), we see immediately that this optimum occurs when [32]

$$Q_L = Q_{L,opt,G_m} = \sqrt{1 + \frac{5\gamma}{\delta\alpha^2}} \geq 1.18 \quad (3.63)$$

Hence, the best noise performance for a given transconductance is achieved at some specific input. Note that the value 1.87 is valid only for long-channel devices. For short-channel lengths, where  $\alpha < 1$ , we can expect the optimum  $Q_L$  to be somewhat larger. By substituting (3.63) into (3.55), we determine that the minimum noise factor (neglecting inductor and gate losses) is

$$F_{min,G_m} = 1 + \sqrt{\frac{4}{5}} \delta\gamma \left(\frac{\omega}{\omega_T}\right) \left\{ |c| + \sqrt{1 + \frac{5\gamma}{\delta\alpha^2}} \right\} \geq 1 + 1.33 \left(\frac{\omega}{\omega_T}\right) \quad (3.64)$$

The value 1.33 is only valid for long-channel devices; it may be three to four times larger in the presence of high electric fields.

### 3.6.2 FIXED $P_D$ OPTIMIZATION

An alternate method of optimization fixes the power dissipation and adjusts  $\rho$  to find the minimum noise factor. If we assume that  $\alpha \ll 1$ , then  $P(\rho, P_D)$  can be simplified to:



$$P(\rho, P_D) \approx \frac{\frac{P_D}{P_0} \left(1 + \frac{\delta}{5\gamma}\right) + 2|C| \sqrt{\frac{\delta}{5\gamma}} \rho^2 + \frac{P_0 \delta}{P_D 5\gamma} \rho^4}{\rho^3} \quad (3.65)$$

This expression is minimized for a fixed PD when  $\frac{\delta P(\rho, P_D)}{\delta \rho} = 0$

The solution of this equation, under the assumption that  $\alpha \ll 1$

$$\rho_{opt, P_D}^2 = \frac{P_D}{P_0} |C| \sqrt{\frac{5\gamma}{\delta}} \left[ 1 + \sqrt{1 + \frac{3}{|C|^2} \left(1 + \frac{\delta}{5\gamma}\right)} \right] \quad (3.66)$$

By comparing Equation (3.66) to Equation (3.65), it is clear that this value for  $\rho$  is equivalent to an optimum  $Q_L$  of

$$Q_{L, opt, P_D} \approx |C| \sqrt{\frac{5\gamma}{\delta}} \left[ 1 + \sqrt{1 + \frac{3}{|C|^2} \left(1 + \frac{\delta}{5\gamma}\right)} \right] \quad (3.67)$$

So, it is clear that the optimum  $Q_L$  for fixed power dissipation  $P_D$  is larger than the optimum  $Q_L$  for a fixed Gm. from these equations

$$F_{min, P_D} \approx 1 + 2.4 \frac{\gamma}{\alpha} \left( \frac{\omega}{\omega_T} \right) \geq 1 + 1.62 \frac{\omega}{\omega_T} \quad (3.68)$$

Where the value of 1.62 is valid only in the long-channel limit; the value will be somewhat larger for short-channel devices in velocity saturation.

### 3.7 LINEARITY

The linearity of the LNA is another concern that must be taken into account. Linear operation is crucial, particularly when the input signal is weak with a strong interfering signal in close proximity. This is because in such a scenario there is a possibility for undesired intermodulation distortion such as blocking and cross modulation. Third-order intercept (IP3) and 1-dB compression point ( $P_{1dB}$ ) are two measures of linearity. IP3 shows at what power level the third-order intermodulation product is equal to the power of the first-order output. IIP3 and OIP3 are the input power and output power respectively, that corresponds to IP3.  $P_{1dB}$  shows at what power level the output power drops 1 dB, as a consequence of non-linearity, relative the theoretical linear power gain [29].

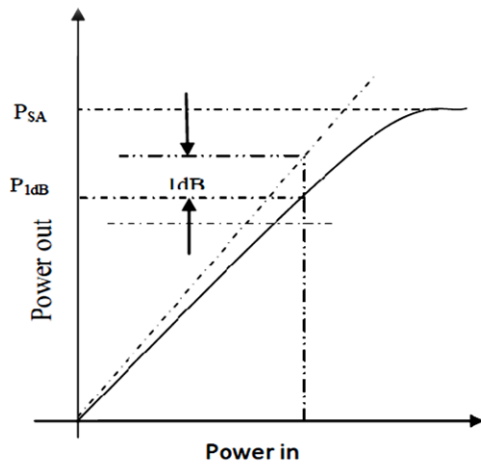


Figure 3.8: 1dB compression point

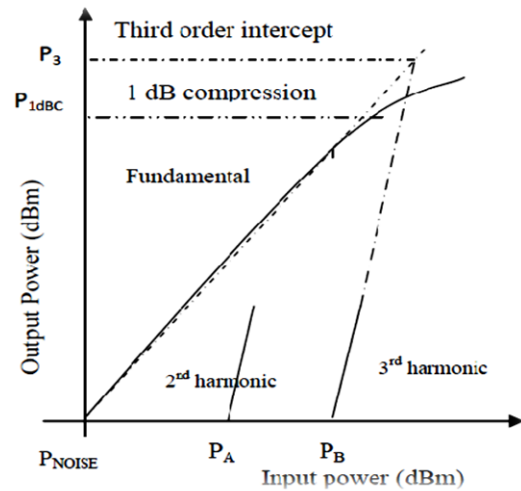


Figure 3.9: Third order intercept point

By knowing either  $IP_3$  or  $P_{1dB}$  the other can be estimated with the following rule-of-thumb formula:

$$IP_3 = P_{1dB} + 10dB \quad (3.69)$$

Third-order intercept point is shown in Figure 3.9. Both measurements indicate an upper distortion limit for the tolerable input power, whereas the noise figure sets a lower limit. The ratio of the two determines the dynamic range of the amplifier [10].

### 3.8 CENTRE FREQUENCY AND BANDWIDTH

As the LNA will operate with input signals of a particular frequency band, it is desired to design it with a centre frequency and bandwidth accordingly. Looking at the transfer function of the LNA, the differential of the two points around the centre frequency  $f_0$ , where the power gain is halved, is the bandwidth, denoted  $\Delta f$  in Figure 3.10. Although the target bandwidth should be specified numerically, by naming convention there are two options: narrowband and wideband.

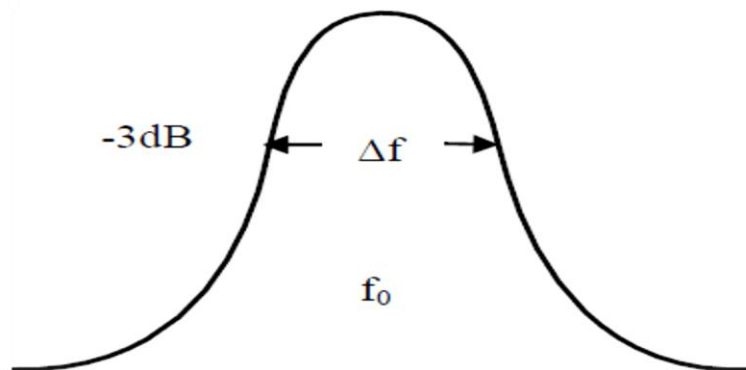


Figure 3.10: Illustration of centre frequency and bandwidth

And  $\Delta f$  is called the bandwidth and it is difference between higher frequency component and lower frequency components and  $f_0$  is the operating frequency and this work operating frequency is 6GHz and bandwidth of system has to calculated by the simulation of circuit. We can also calculate the relative bandwidth that tells us system is wideband and narrow band.

# CHAPTER 4

## DIFFERENT TOPOLOGIES FOR LNA DESIGN

Typical heterodyne receiver architecture as shown in Figure 4.1 and it is clear from the above figure that the signal received from the antenna has to pass from the band select band pass filter and then it goes to the LNA. The reflection from LNA is never desired because it will affect the filter characteristic of the band pass filter preceding the LNA and hence the overall performance of receiver will be altered. Further for the signal that is being received by LNA, one has to ensure that maximum part of incoming signal is received by the LNA.

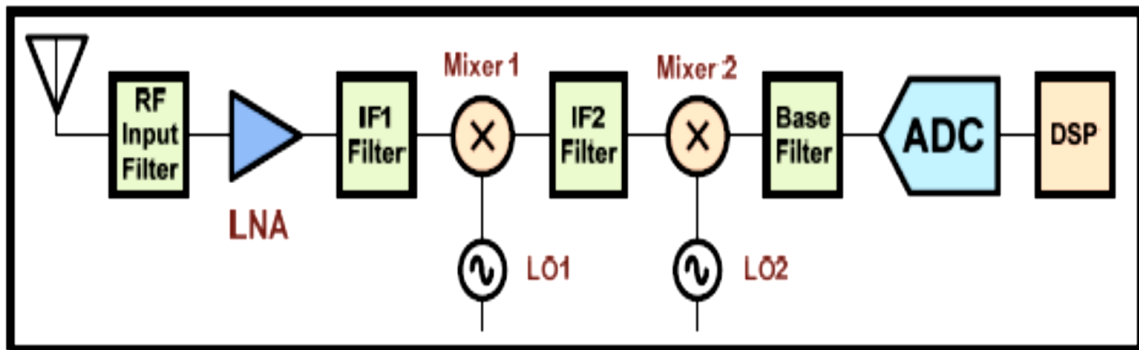


Figure 4.1: Basic heterodyne receiver architecture [36]

So the first stage of LNA will be the input matching stage which will insure that maximum signal is going into LNA for processing. And from above section it is clear that in order to get maximum power transfer theorem that is in for the signal received by the LNA to show minimum reflection the input impedance of the LNA must be complex conjugate of the source impedance. Here for simplicity it is assumed that source impedance is real and its value is  $50\Omega$ . Now it is also known that if a number of components are connected in cascade configuration than the equivalent noise figure of the chain will be-

$$F_{Tot} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots + \frac{F_n - 1}{G_1 G_2 \dots G_{n-1}} \quad (4.1)$$

From eqn.(4.1) it is clear that the noise figure of the system mainly depends upon the noise figure of the first stage because it directly gets added to the noise figure of the system while the noise figure of the system which comes after the first component gets divided by the gain of the preceding component and so on, so it is very important to ensure that the LNA is not adding much noise.

So from above discussion it is clear that mainly there are two conditions for LNA first is that the input impedance is matched with the source impedance and second condition is that noise figure of LNA should be minimum with high gain. Also from the above section it is clear that it is impossible to satisfy both the condition simultaneously and accurately, therefore one will have to go for trade off the respective stages of development of LNA will be discussed starting from the basic circuit of LNA possible. So that there are some topologies for LNA design are discuss below-

#### 4.1 RESISTIVE MATCHING

The following is the figure of the simple resistive type matching LNA circuit. As shown in the figure 4.2 the input impedance of MOSFET can be taken as infinity for particular frequency of interest so in order to get the matching  $R_{in}$  should be equal to  $R_s$ . so in the case  $V_{in}$  will be equal to half of the value of  $V_s$ . Further although it will be a broadband matching but it will show very poor performance because it can be seen that half of the signal power is already lost before going to the LNA and also this circuit is adding noise i.e. the thermal noise because of  $R_1$  is being added to the input stage of LNA.

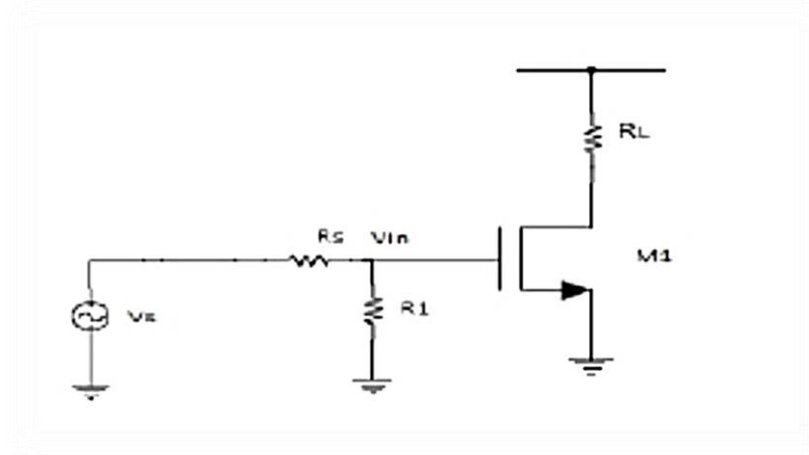


Figure 4.2: Resistive matching LNA [37]

Adding noise to the output stage of LNA will not be that dangerous because overall noise figure gets divided by the gain of preceding stages.

Noise figure of a circuit is given by following relation:

$$\text{Noise Figure} = \frac{\text{total output noise}}{\text{total noise due to source}}$$

Calculating noise figure from figure 4.2

Identifying the noise sources in the circuit

- (i) Due to source resistance  $R_S$

$$\overline{V_{nR_s}^2} = 4KTBR \quad (4.2)$$

- (ii) Due to termination resistance  $R_1$

$$\overline{i_L^2} = 4KTG_L B \quad (4.3)$$

- (iii) Due to induced drain current

$$\overline{i_d^2} = 4KT(\gamma g_{do}) B \quad (4.4)$$

In above equation  $\gamma$  is a constant that has value unity for linear region and reduced to 2/3 in saturation region and  $g_{do}$  is zero bias drain conductance.

- (iv) Due to load resistance  $R_L$

$$\overline{i_L^2} = 4KTG_L B \quad (4.5)$$

$G_L$  is the load transconductance.

So from equation 4.2 to 4.5 total output noise current is given by

$$\overline{i_0^2} = \overline{i_d^2} + \overline{i_L^2} + \left( \overline{V_n^2 R_S} + \overline{V_n^2 R_1} \right) g_m^2 \quad (4.6)$$

So noise figure using equation 4.2 to 4.6 is given by-

$$F = 1 + \frac{R_1}{R_S} + \frac{\gamma}{g_m R_S} + \frac{G_L G_S}{g_m^2} \quad (4.7)$$

From equation (4.7), it is observed that even if  $g_m$  becomes very large noise figure still will be dominated by termination resistance  $R_1$ . Hence, termination by resistor is not a good option as noise figure is very high. Clearly the value of noise figure that will be obtained from this configuration will be more than 3dB so this configuration is not practically applicable.

## 4.2 SHUNT-SERIES FEEDBACK AMPLIFIER

Another possible configuration can be shunt feedback amplifier which is as shown below-

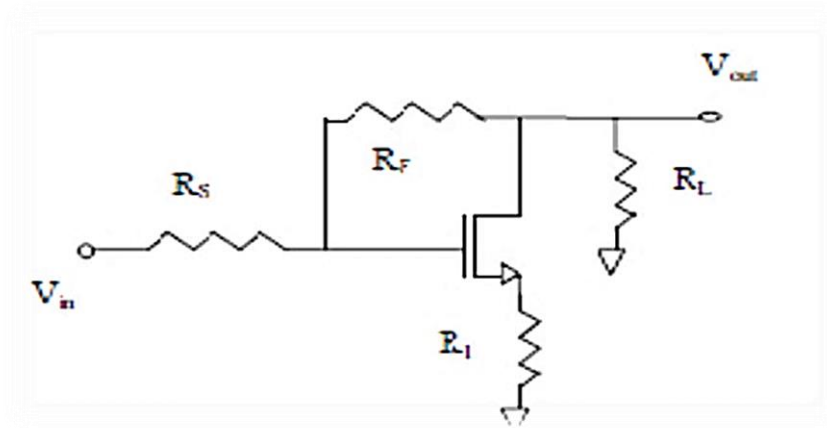


Figure 4.3: Shunt feedback amplifier [37]

It can be seen from this configuration that the main problem of previous configuration is not there i.e. half power was being attenuated before coming to the actual circuit which was responsible for the poor performance of resistive type matching circuit. It can be shown that the value of noise figure for this configuration can be represented as Equation (4.8).

$$F = 1 + \frac{R_S}{R_F} \left( 1 + \frac{1}{g_m R_S} \right)^2 + \frac{\gamma}{g_m R_S} \quad (4.8)$$

It is clear from the above equation that in order to get minimum noise figure  $g_m R_S$  should be more and  $R_F$  should be more, but the main problem is  $R_F$  induces noise so the both conditions contradicts each other. However the performance of this topology is better than the common source topology.

### 4.3 COMMON GATE AMPLIFIER

The common gate topology for low noise amplifier as shown in Figure 4.4, the noise figure of this configuration can be calculated by considering the effect of drain current. Since the drain current is injected into the input, it adds noise in the shunt with the input noise current.

By definition,

$$\text{Noise Figure} = \frac{\text{total output noise}}{\text{total noise due to source}} = \frac{\overline{\hat{i}_s^2} + \overline{\hat{i}_d^2}}{\overline{\hat{i}_s^2}}$$

$$F = 1 + \frac{4KT\gamma Bg_{do}}{4KTBG_s} \quad (4.9)$$

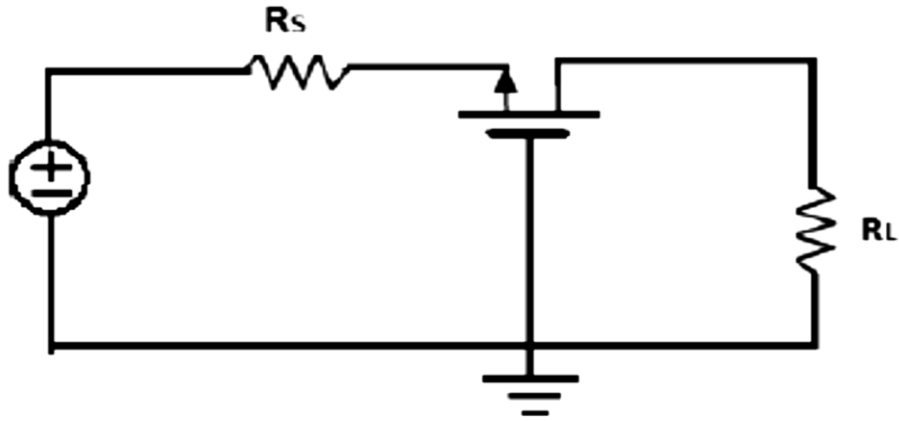


Figure 4.4: Common gate topology [37]

On solving Equation (4.8),

$$F = 1 + \frac{\gamma}{\alpha} \quad (4.10)$$

Since  $\gamma$  is  $2/3$  for saturation region and  $\alpha$  is unity. Putting these values in equation (4.10), noise figure is typically equal to 2.2dB and for short channel devices this value may increase to 3dB [37]. So  $(1/g_m)$  termination is not a good way for matching purposes.



## 4.4 INDUCTIVE SOURCE DEGENERATION

Ideal inductor contributes no internal noise. However, real inductors suffer from both internal and external noise [37]. Inductive source degeneration offers the best noise match among the all techniques discussed earlier.

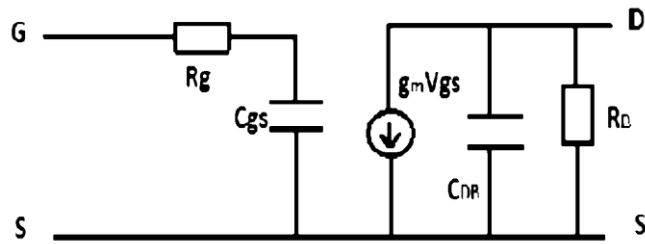


Figure 4.5: Equivalent model of MOSFET

Figure 4.5 shows the equivalent model of MOSFET consisting of voltage source in parallel with the load resistance  $R_D$ . As goal is to provide  $50\Omega$  proper termination but in reality the input of the device is reactive, with real and capacitive impedance. Capacitive reactance can be removed by adding inductive feedback to the source.

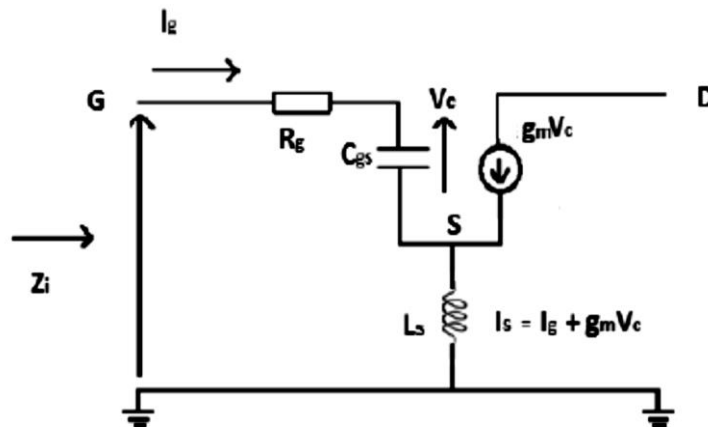


Figure 4.6: Equivalent MOSFET model with inductive source degeneration

Calculating input impedance from figure 4.6 by applying KVL at input, it comes out to be

$$Z_i = R_s + R_g + j(XL_s - XC_{gs}) \quad (4.11)$$

In Equation (4.11),

$$R_s = \frac{L_s g_m}{C_{gs}}$$

From equation (4.11), it can be observed that by adding series feedback,  $R_s + j\omega L_s$  term is added to original input impedance. Additionally, another inductor is added in series with the gate  $L_g$  that is selected to resonate with the  $C_{gs}$ .  $L_g$  is designed so that at resonant frequency it cancel out  $C_{gs}$  that is-

$$j(\omega L_s - \frac{1}{\omega C_{gs}}) = 0, \text{ such that } Z_i = \frac{L_s g_m}{C_{gs}} = 50\Omega \quad (4.12)$$

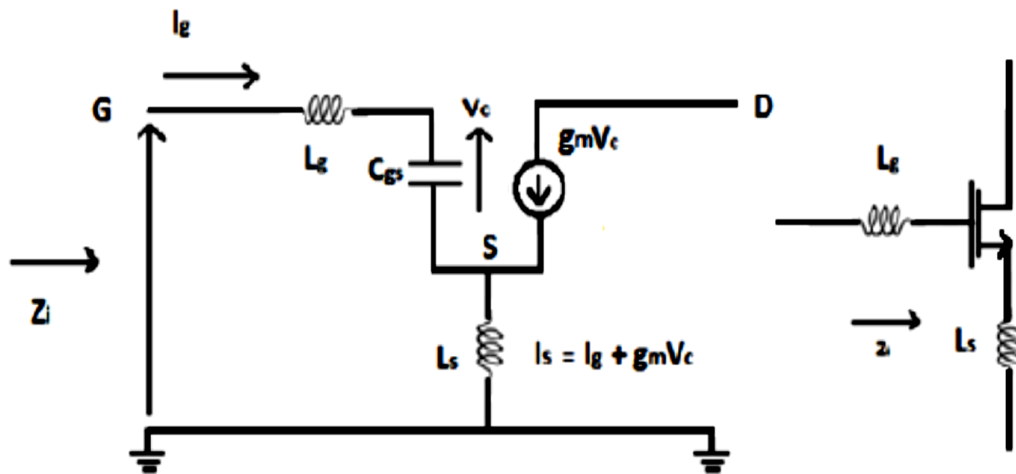


Figure 4.7: MOSFET model with inductive Gate and inductive source Degeneration

## 4.5 CASCODE INDUCTIVE SOURCE DEGENERATION

### Miller Effect

An important phenomenon that occurs in many analog circuits is related to “Miller Effect”. It states that figure 4.8 for any impedance can be converted to figure 4.9 provided that the impedance  $Z$  appears in parallel with the main signal.

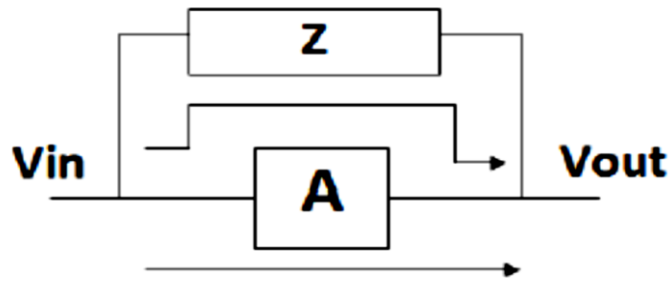


Figure 4.8: Typical Case for Miller's Theorem [38-39]

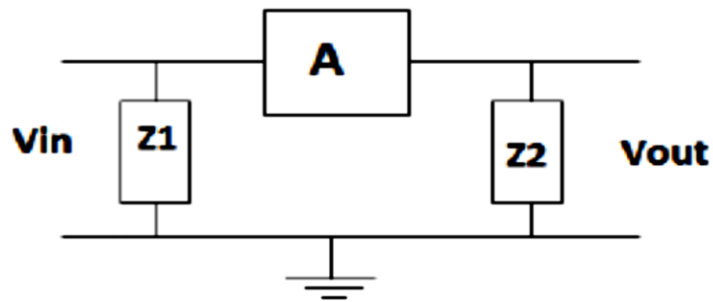


Figure 4.9: Application of Miller Theorem [38-39]

In figure 4.9

$$Z_1 = \frac{Z}{1-A} \quad (4.13)$$

And

$$Z_2 = \frac{Z}{1-A^{-1}} \quad (4.14)$$

Above Equation for  $Z_1$  and  $Z_2$  can be resistive, capacitive, or inductive and produces undesired results. For example if  $Z$  is capacitive in nature then at high frequencies it will limit the frequency response both at the input and output. So, some means should be provided in order to suppress the miller effect.

### Cascode Configuration

Cascode refers to the combination of common source and common gate. Main purpose of using the cascode configuration is to suppress the miller effect at the output and to increase the gain by increasing the output impedance of the circuit.

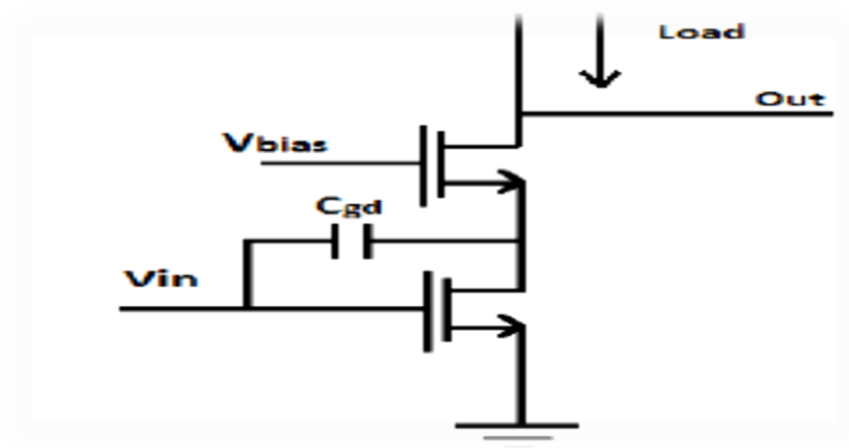


Figure 4.10: Cascode Configuration

From the figure 4.10 it is clear that the effect of miller capacitance  $C_{gd}$  has been suppressed at the output.

### Inductive Load

Instead of using resistive load inductive load is used for design of the low noise amplifier because of following advantages:

- The inductor between the Cascode source and supply blocks any RF leaking to the supply rail and may be varied in value to optimize the gain response of the LNA.
- To tune out the capacitances producing at the output, thereby providing band pass filtering.
- Compare to resistive load inductive load produces less noise

The advantages and disadvantages of different kinds of LNA topologies are shown in Table 4.1.

**Table 4.1: Advantages and Disadvantages of LNA topologies**

Type of Topology	Advantages	Disadvantages
Resistive termination common source	Broad band amplifier	Adding the noise from the resistor.

Common Gate	The input impedance is equal to $1/g_m$ . It is practical to get $50 \Omega$	The impedance varies with the bias current.
Shunt series feedback common source	Broad band amplifier	Adding noise from resistor
Inductive degeneration common source	The source and gate inductors make the input impedance $50 \Omega$ . Not adding the noise from the input.	The inductor is off chip at low frequency and low isolation.
Cascode inductor source degeneration	Isolation of input and output is good, higher gain, lower noise figure.	The inductor is off chip at low frequency.

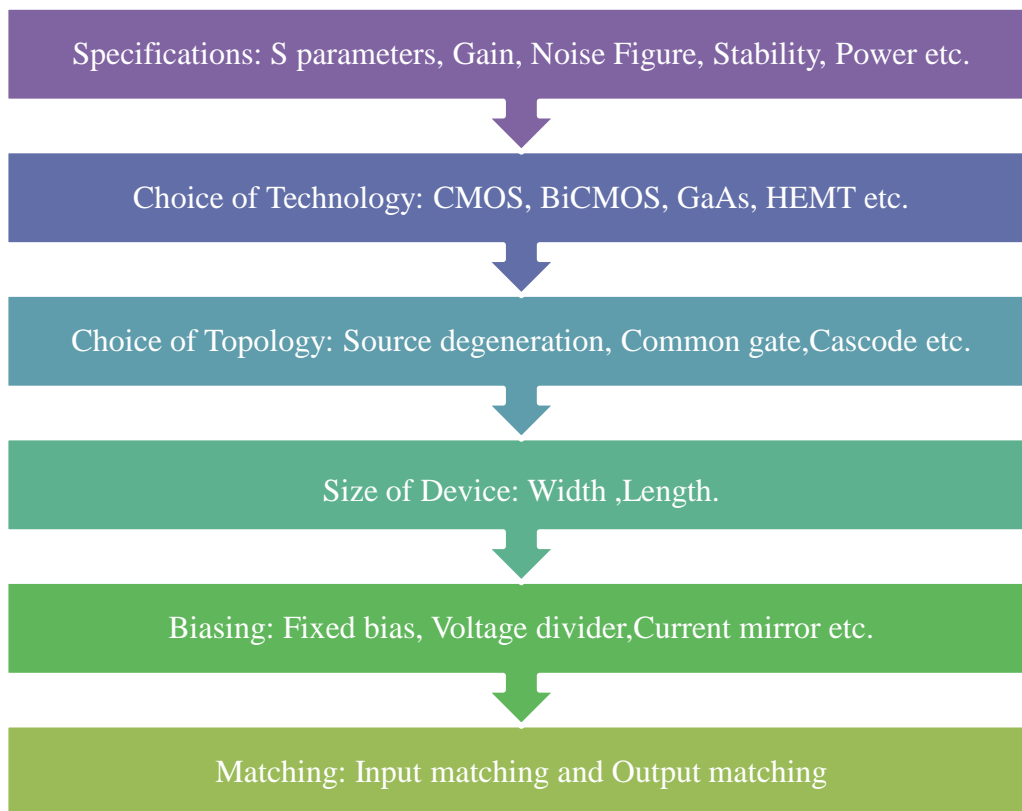
As all the above topologies, inductive source degeneration configuration is best from other topologies. So we have chosen inductive source degeneration configuration, with inductive load it provides sufficient gain without adding significant noise and also a good property it provides better matching in comparison to other topologies.

# CHAPTER 5

## DESIGN, SIMULATION AND RESULT

In the designing of low noise amplifier, some basic aspect of low noise amplifier should be kept in mind. Since in low noise amplifier there is trade-off between many factors i.e. gains, noise figure, stability, power consumption, and nonlinearity etc. And depending on these factors we have previously discussed different topology of low noise amplifier design. In this work we have chosen inductive source degeneration topology because this provides some beneficial advantages over other topologies as discussed in chapter 4. Inductive source degeneration topology is one of the most useful configurations for low noise amplifier design; because common source configuration provides low noise and solved the stability related problem.

### 5.1 FLOW CHART OF LNA DESIGN



Flow chart of LNA design is shown above that describes the design flow of low noise amplifier. The design starts from some basic specification means some standard output

values that is desired for any circuits design these specification are scattering parameters, Gain, Noise Figure, Stability, Power, linearity etc. and our aim, tends to achieve these specified values. Second step in the design is choice of technology so here; HEMT technology has chosen for this design. Third step is choice the topology, there are different types of topologies and among them we can use topology depending on our design. Fourth step is some basic calculation like device width, gate to source capacitance and other things after that we can do the biasing to set the operating point and last if there is some impedance mismatch so we have to do impedance matching by inserting a filter network

## 5.2 DESIGN SPECIFICATION

Low Noise Amplifier Design for 2.4 GHz operating frequency range by using 0.18um HEMT Technology, 0.18um is represents the length of device. That is the minimum length of device provided by the vender. HEMT Technology is used because of its several advantages such as low cost, low static power dissipation, low area, low noise and high dynamic range. If we goes to design any analog circuit it is mandatory to known the some design specification and that specification are standard and our goal is to achieve the desirable result. Table 5.1 shows the target specification for LNA design.

**Table 5.1: Target Specification**

Parameters	Target value
Centre frequency	2.4GHz
Voltage Gain(dB)	> 12dB
NF(dB)	< 2.0dB
Matching parameters	<-10dB
Stability Factor	> 1
Supply Voltage	3.0 V

## **5.3 SELECTION OF APPROPRIATE COMPONENTS**

Because transistors, spiral inductors and capacitors are often used in LNA circuit, the accurate RF models are very important to predict the silicon performance of gigahertz circuits. The characteristic of transistor in low frequency is different from the one in high frequency. The parasitic effects of transistor should be considered in circuit design, which are not included in the low frequency circuit design. So the transistor model for low frequency design is quite different with the model for high frequency design. Moreover in high frequency, the inductance and Q value varies with the operating frequency and capacitor also has parasitic effects. Advanced Design System (ADS) of Agilent is used for the design of various topologies of LNA. So transistor, inductor and capacitor models should also be chosen carefully for correct design from the design library Advanced Design System (ADS) simulator. The ADS library is used for selecting other active and passive components.

### **5.3.1 Transistor RF Models**

MOSFET models, especially the RF MOSFET models are required to predict the silicon performance accurately, such as sub-circuit short channel MOSFET models for RFIC designs. In the sub-circuit models, MOSFET is divided into two parts, an intrinsic part, and an extrinsic part. The intrinsic part represents the main active part of the device, which can be any compact model, such as Berkeley Short-Channel IGFET Model (BSIM). However, the extrinsic part consists of most of the parasitic elements, including all the terminal access series resistance, gate resistance, overlap and junction capacitance, and substrate network. The library in ADS contains various RF and microwave transistors models.

From these HEMT, ATF-54143 device is used for the designing of LNAs. The length of this transistor is constant 180.0nm and width is variable. The maximum allowable width of ATF-54143 in ADS is 105 $\mu$ m. Two or more transistors can be used in parallel for the bigger transistors. Various details of ATF-54143 have given in appendix A.

### **5.3.2 Inductor RF Models**

Inductors are employed for tasks such as bias feeding and impedance matching of transistor spiral inductor are similar to inductive transmission line. So many parameters



and careful trade off must be considered for the inductor design. Spiral inductors with reasonable Q and self-resonant frequency are widely used in the RFIC designs, such as fully integrated LNA, oscillator, and impedance matching network. They are proved to be most difficult passive components to be implemented on chip. Circular square spiral inductor is defined by side length, wire width, wire space, and number of turns. The component library in ADS contains the inductor model with its variable values having maximum and minimum limit set by the vendor. The inductance value is depending on these parameters such as turns number, diameter, and width. As the turn number decreases the inductance also decreases.

### **5.3.3 Capacitor RF Models**

Capacitors are another important passive components widely used in RF circuit design, such as impedance matching and DC block. The usage of a capacitor is primarily dependent upon the characteristics of its dielectric. The dielectric's characteristics also determine the voltage levels and the temperature extremes at which the device may be used. The component library in ADS contains the capacitor model with its variable values having maximum and minimum limit set by the vendor.

### **5.3.4. Resistor RF Models**

Resistance is the property of a material that determines the rate at which electrical energy is converted into heat energy for a given electric current. At very high frequencies and with low-value resistors (under 50 ohms), lead inductance, and skin effect may become noticeable. The component library in ADS contains various resistor models with its variable values having maximum and minimum limit set by the vendor.

## **5.4 LNA DESIGN WITH LUMPED COMPONENT**

Transistor must be biased at appropriate operating point before used. So that, transistor can works under values required and achieve less power consumption. In this project, passive biasing method is adopted [40-42]. The component readings are determined with reference from datasheet. By referring datasheet, data of  $V_{ds} = 3V$  and  $I_{ds} = 60mA$  had be chosen because it is believed that it can give optimum values in gain and noise figure With  $V_{gs} = 0.52$ ,  $I_{BB}=2mA$ ,  $V_{ds} = 3V$  and  $I_{ds} = 60 mA$ .

$$R_3 = \frac{V_{gs}}{I_{BB}} = 260\Omega, R_4 = \frac{V_{dd} - V_{ds}}{I_{ds} + I_{BB}} = 32.26\Omega, R_1 = \frac{(V_{ds} - V_{gs})R_3}{V_{gs}} = 1.24k\Omega$$

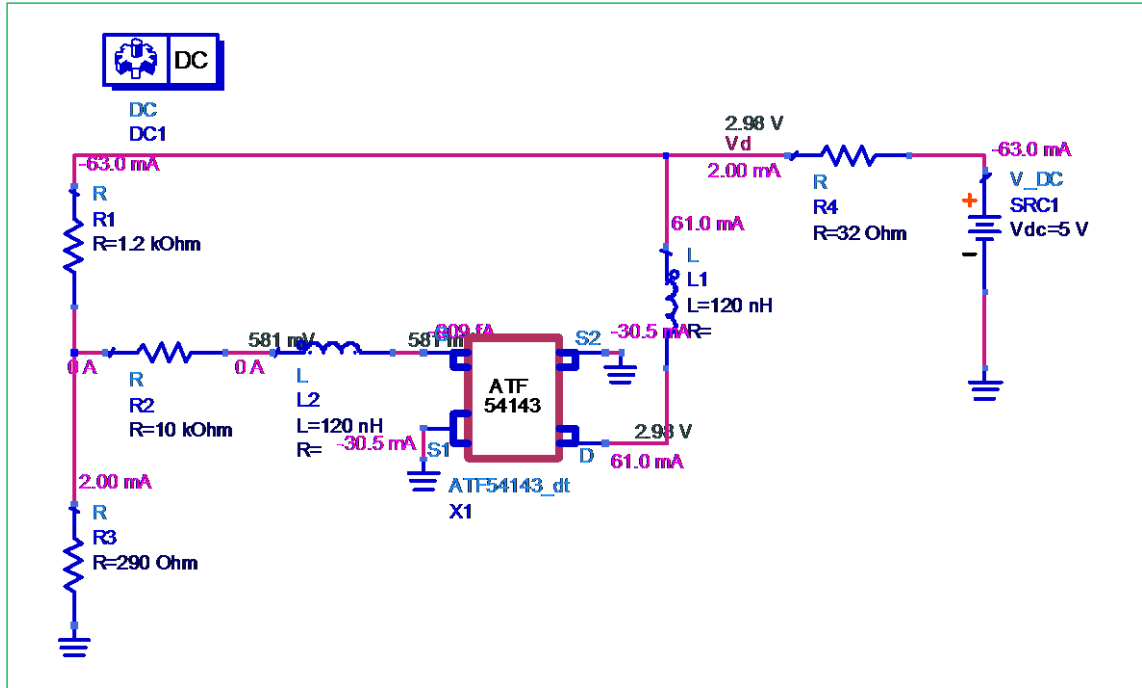


Figure 5.1: DC biasing Circuit

The  $R_1$ ,  $R_4$  and  $R_3$  in circuit are slightly adjusted from calculated readings in order to obtain better  $V_{ds}$  and  $I_{ds}$  reading in the simulation. Given  $Z_L=Z_S=Z_0=50\Omega$

From the data sheet which is given in appendix A, S parameter at 2.5 GHz is:

$$S_{11}=0.60\angle 176.2$$

$$S_{21}=6.01\angle 61.80$$

$$S_{12}=0.07\angle 30.10$$

$$S_{22}=0.13\angle -129.7$$

The reflection coefficient to source,  $\Gamma_S$  and reflection coefficient to load,  $\Gamma_L$ , and the equation is shown in (5.1) and (5.2).

$$\Gamma_s = \frac{Z_s - Z_0}{Z_s + Z_0} \quad (5.1)$$

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (5.2)$$

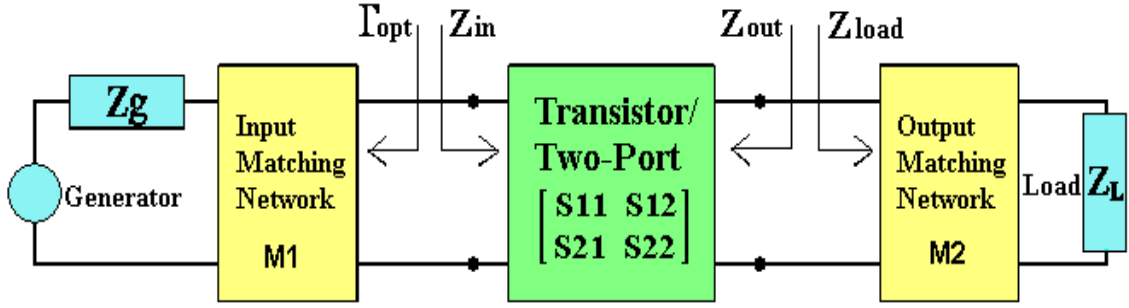


Figure 5.2: Block diagram of a single stage two port LNA model

Since given  $Z_L = Z_S = Z_0 = 50\Omega$ , the  $\Gamma_L$  and  $\Gamma_S$  calculated as zero. Reflection coefficient at the input and output:

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = S_{11} = 0.60 \angle 176.2 \quad (5.3)$$

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} = S_{22} = 0.13 \angle -129.7 \quad (5.4)$$

Power gain,  $G$

$$G = \frac{|S_{21}|^2 (1 - |\Gamma_L|^2)}{(1 - |\Gamma_{in}|^2) |1 - S_{22}\Gamma_L|} = 56.43 = 17.52dB \quad (5.5)$$

Available power gain,  $G_A$

$$G_A = \frac{|S_{21}|^2 (1 - |\Gamma_S|^2)}{|1 - S_{11}\Gamma_S|^2 (1 - |\Gamma_{out}|^2)} = 36.74 = 15.65dB \quad (5.6)$$

Transducer gain,  $G_T$

$$G_T = \frac{|S_{21}|^2 (1 - |\Gamma_S|^2) (1 - |\Gamma_L|^2)}{|1 - \Gamma_S\Gamma_{in}|^2 |1 - S_{22}\Gamma_L|^2} = 36.12 = 15.58dB \quad (5.7)$$

### 5.4.1 Stability Consideration

The stability of an amplifier or its resistance to oscillate is an important consideration in a design. It can be determined from the S parameters. Oscillation occurs when  $K < 1$ .

This is due to the dependence of and on the source and load matching networks. An amplifier is said to be unconditionally stable if the auxiliary condition along with *Rollet's condition*, defined as in equations below, are simultaneously satisfied [40],[42].

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

$$|\Delta| = |S_{11}S_{22} - S_{21}S_{12}| < 1$$

For this design it is found that  $K = 1.009 > 1$  and  $\Delta = 0.3427 < 1$

### 5.4.2 Noise Figure

Besides the stability and gain requirement, noise figure is another important consideration for a microwave amplifier. In receiver applications, it often required preamplifier with as low noise figure as possible as it has dominant effect on the noise performance of the system. From the ATF-54143 datasheet,  $F_{min} = 0.52$ ,  $\Gamma_{opt} = 0.26$  and  $R_N/Z_0 = 0.04$  at 2.4 GHz.

$$F = F_{Min} + \frac{4r_n |\Gamma_s - \Gamma_{opt}|^2}{(1 - |\Gamma_s|^2) |1 + \Gamma_{opt}|^2} = NF = 2.769dB$$

### LNA circuit before matching

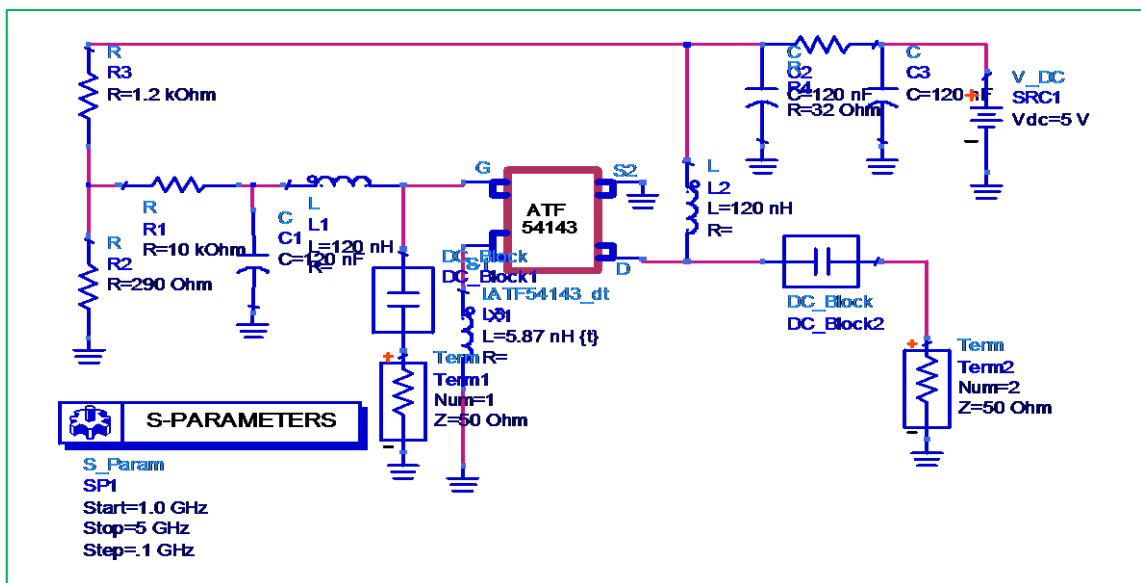


Figure 5.3: LNA circuit before matching network

### 5.4.3 Matching Network

The impedance matching basic idea is presented in Fig. 5.4, which ensembles that an impedance matching network placed between the load impedance and transmission line. Matching network is made to ideally avoid the unnecessary power loss. There are variety of factors that needed to be considered in the matching network selection e.g. complexity, implementation and adjustability. In this work, the LNA is designed by using lumped elements using L-C matching techniques.

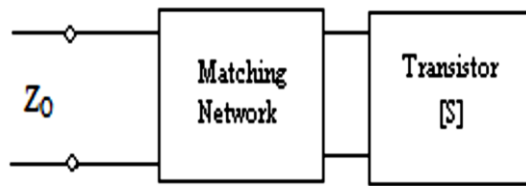


Figure 5.4: Impedance matching network

The parameters noise figure, stability factor, return loss and gain values are calculated by using above equations. While simulation results before the matching are listed as in Table 5.2.

**Table 5.2: Results differences in calculated and simulation before applying matching network**

Parameters	Calculated	Simulated
$ \Gamma_s $	0	$0.51\angle-164.1$
$ \Gamma_L $	0	$0.34\angle 21.3$
$ \Gamma_{in} $	$0.60\angle 176.2$	$0.702\angle 169.92$
$ \Gamma_{out} $	$0.13\angle -129.7$	$0.037\angle -70.502$
K	1.1478	0.971
G	17.52 dB	17.25 dB
$G_A$	15.65 dB	15.31 dB
$G_T$	15.58 dB	15.76 dB
$G_{Tmax}$	15.58 dB	15.63 dB
NF	2.769 dB	0.375 dB

## 5.5 PROPOSED SCHEMATIC DESIGN OF LNA

To match both impedance of input and output Lumped elements matching is added into LNA circuit with assistance of smith chart utility in ADS software. After matching from smith chart, the Lumped elements circuit will automatically generated. The components reading designed by ADS software do not exist in market. Thus, to fabricate, it must change to actual values. Matching has done using with the help of L-C matching technique. Other matching techniques such as Pi match, T-match are also used to match input and output port of amplifier. But here L-C section is used for simplicity and to reduce the complexity of circuit.

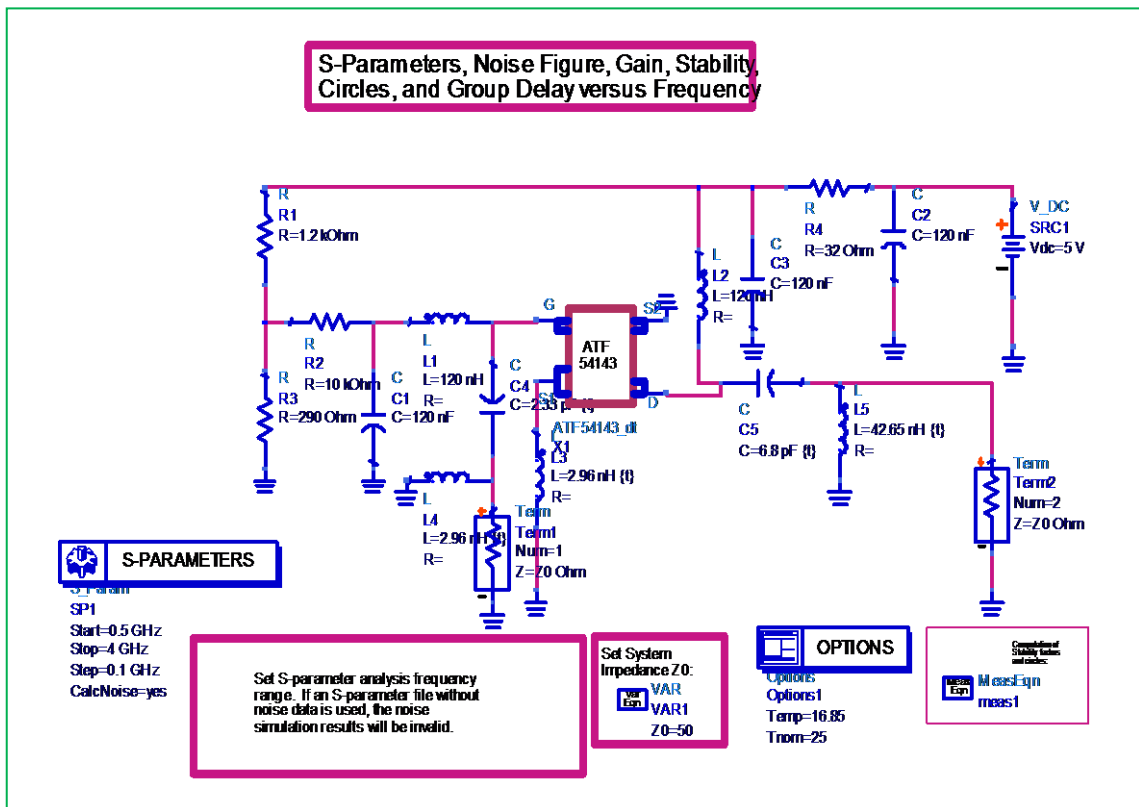


Figure 5.5: Complete LNA circuit with input and output matching circuit

The stability of this transistor is unconditionally stable at 2.4 GHz since its K constant is more than 1, which is 1.009 from the simulation. Therefore, no stability circle has to be drawn. Stability and maximum available gain are two of the more important considerations in choosing a two-port network LNA for use in amplifier design. As used here, stability measures the tendency of an LNA to oscillate.

## 5.6 SIMULATION AND RESULT

**1. S Parameter:** S parameters are scattering parameters as discussed in chapter 3, these are four  $S_{11}$ ,  $S_{12}$ ,  $S_{21}$ ,  $S_{22}$ .  $S_{11}$  is input reflection coefficient, represents measure of how well the input impedance is match to the reference impedance. It is also known as input return loss and the input reflection coefficient  $S_{11}$  for 2.4 GHz frequency is -15 dB as shown in figure 5.6.

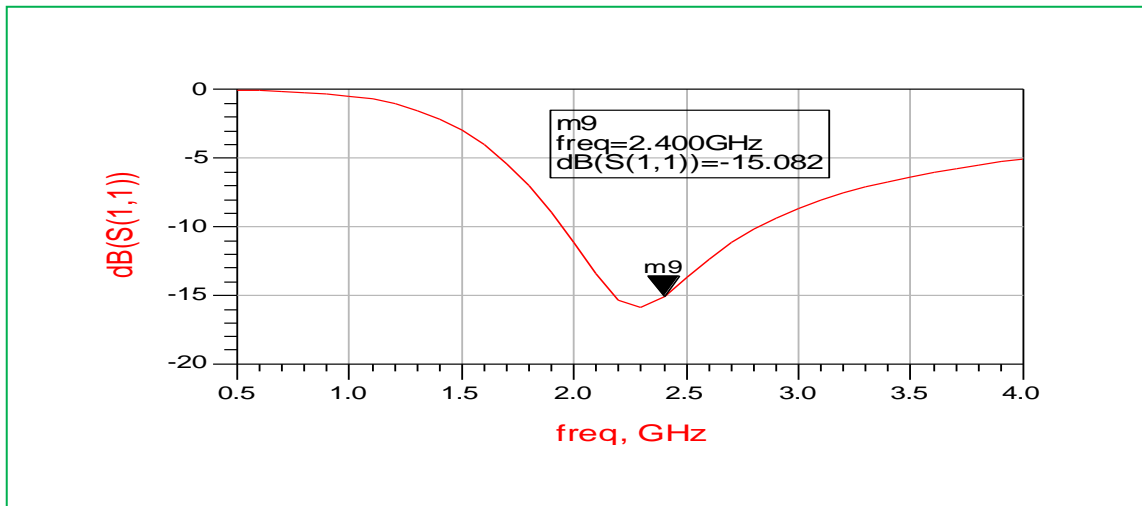


Figure 5.6: Scattering parameter  $S_{11}$

- $S_{22}$  is output reflection coefficient and it represents measure of how well the output impedance is matched to load impedance. It is also known as output return loss, and the value of  $S_{22}$  at the 2.4 GHz frequency is -10.20 dB as shown in figure 5.7.

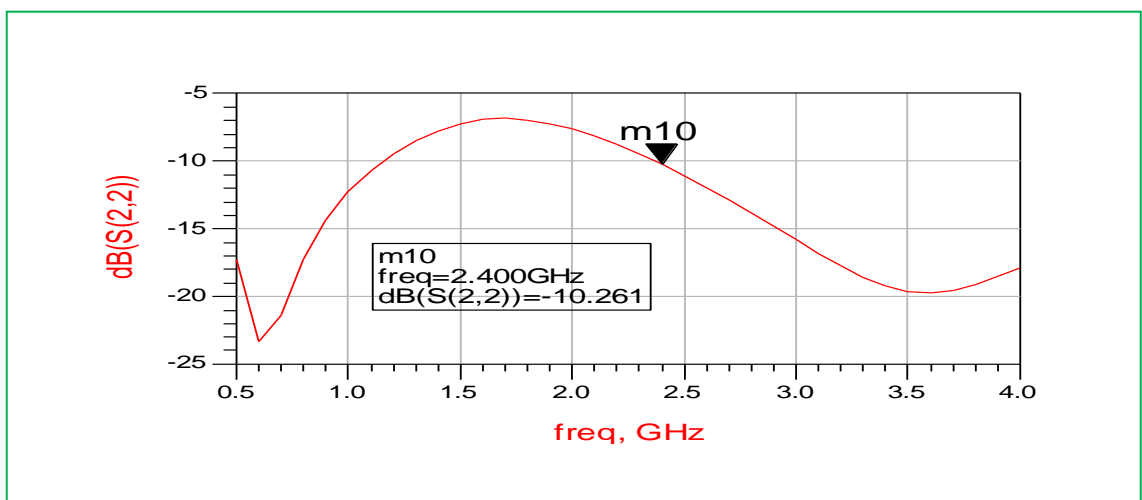


Figure 5.7: Scattering parameter  $S_{22}$

- $S_{21}$  is forward transmission coefficient and it represents, measure how well the signal goes from input to output. It is also known as gain and the value of  $S_{21}$  at 2.4 GHz frequency is 15.227 dB as shown in figure 5.8.

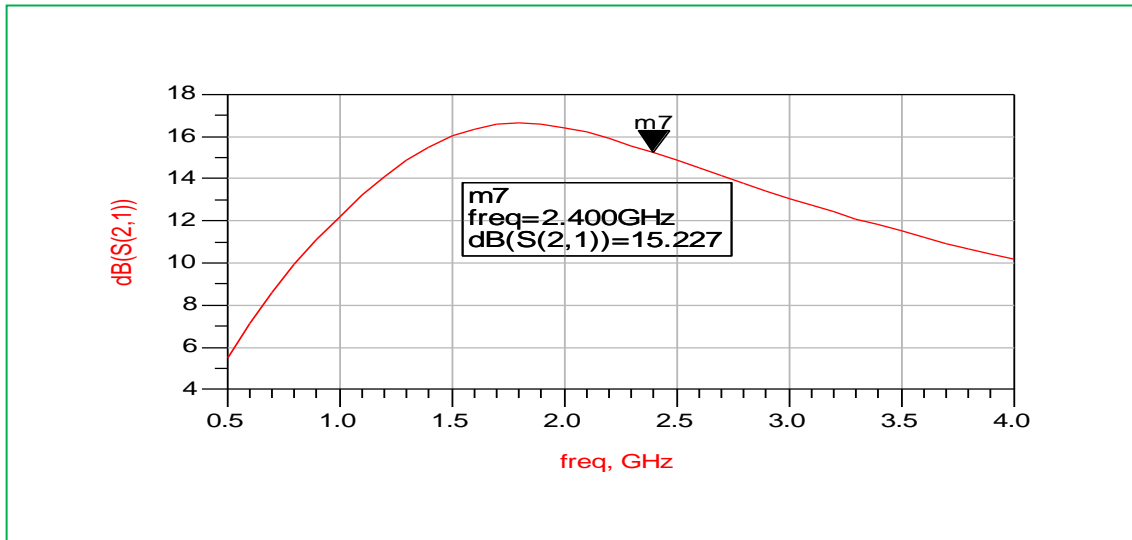


Figure 5.8: Forward Gain ( $S_{21}$ )

- $S_{12}$  is reverse transmission coefficient, and measure how much the input signal is reflected back and it is also known as reverse isolation and the value of  $S_{12}$  at 2.4 GHz frequency is -20 dB as shown in figure 5.9.

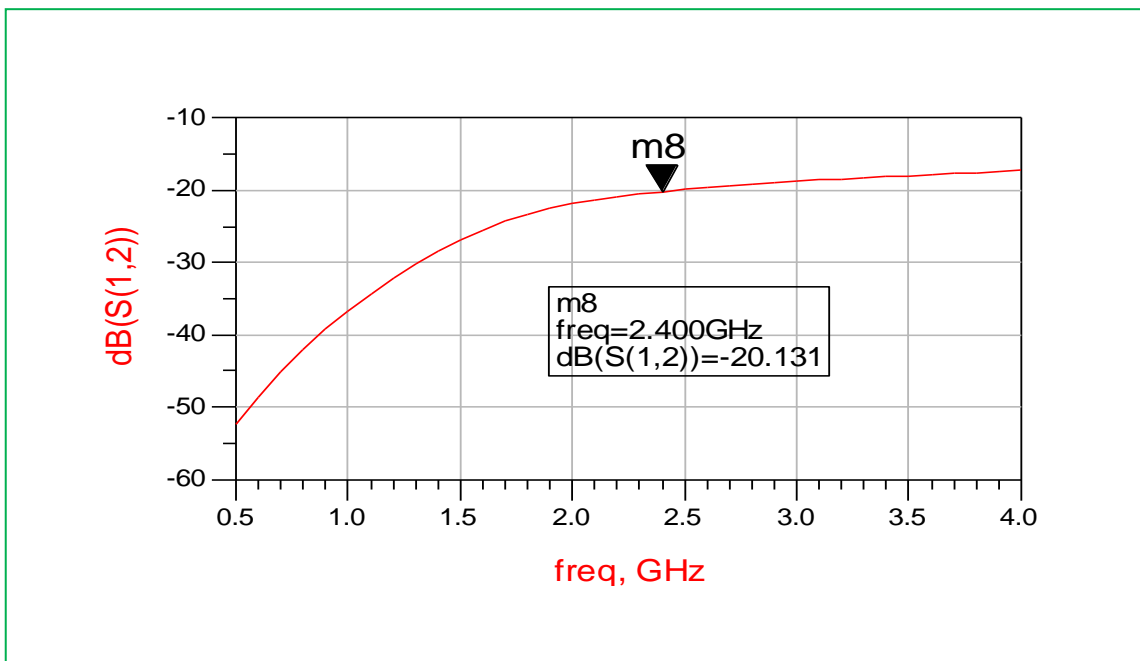


Figure 5.9: Scattering parameter  $S_{12}$



## 2. Noise Figure:

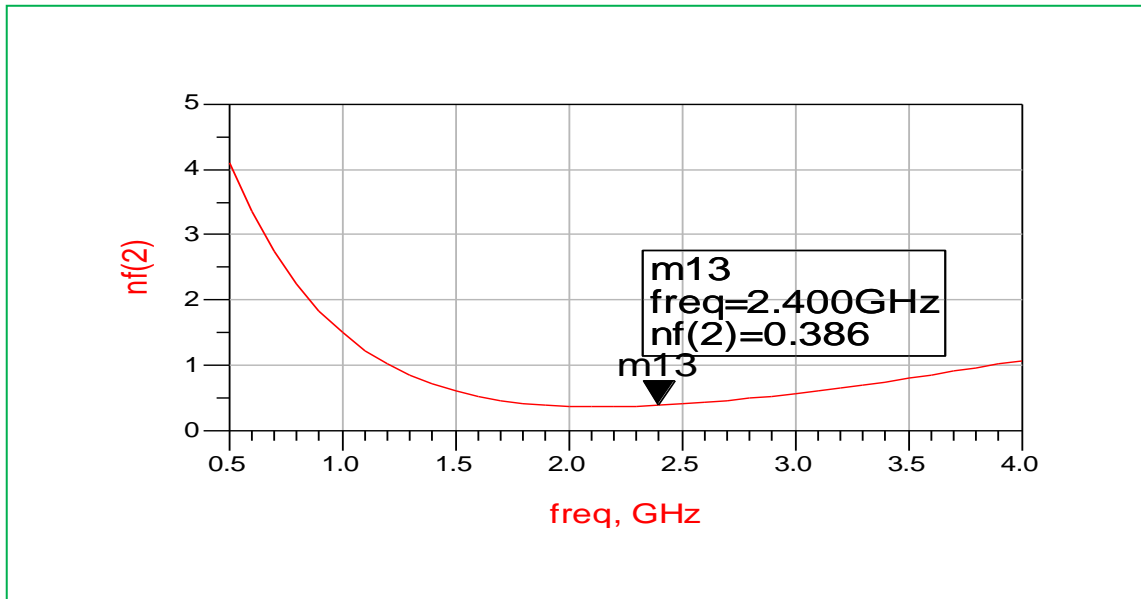


Figure 5.10: Noise Figure (NF)

Simulation result of Noise Figure as shown in figure 5.10. This Noise Figure indicates the noise performances of device at high frequency and its value is 0.386 dB at 2.4 GHz frequency and in the Figure 5.11 shows the minimum Noise Figure, with simulated value is 0.379 dB which is less than calculated value 2.769 dB

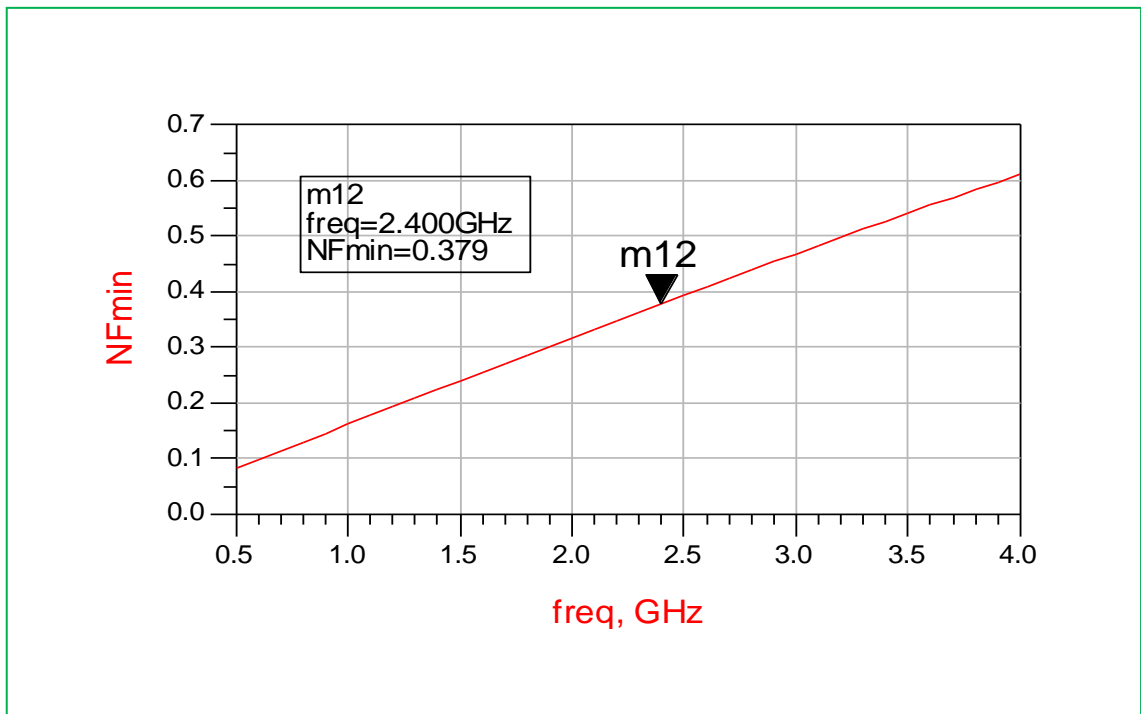


Figure 5.11: Minimum Noise Figure ( $NF_{\min}$ )

**3. Stability:** As we know that stability is defines by the factor K called as rollet stability factor, and this factor is describes in terms of s parameters, so that stability factor of circuit as shown in figure 5.12 and it must be greater than 1 so its value is 1.009 at 2.4 GHz frequency.

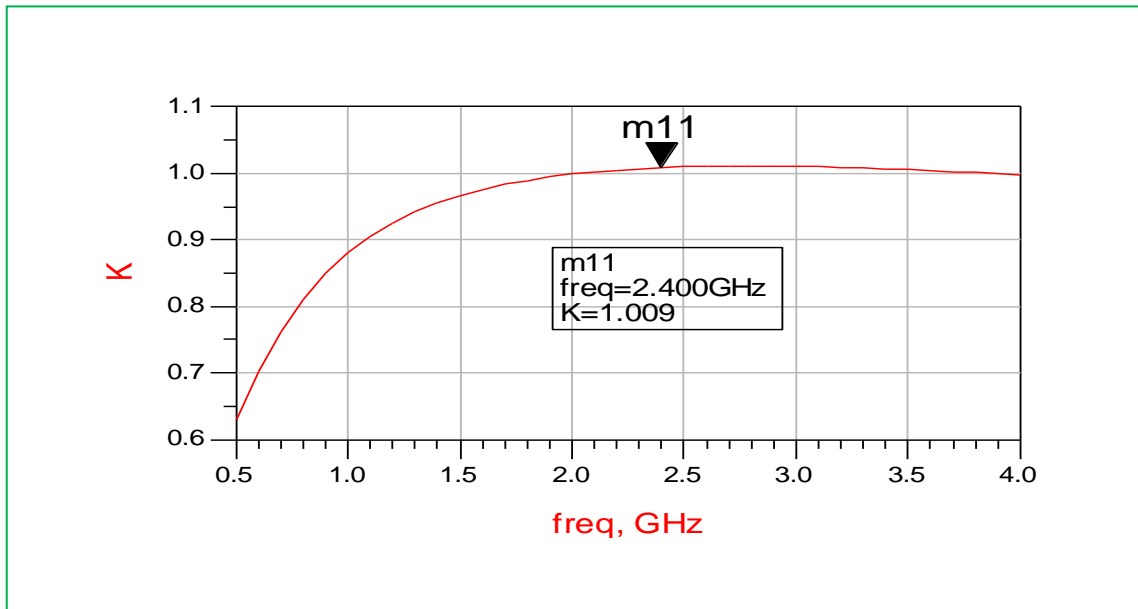


Figure 5.12: Stability factor

- Comparison Graph of different Gain is shown in figure 5.13.

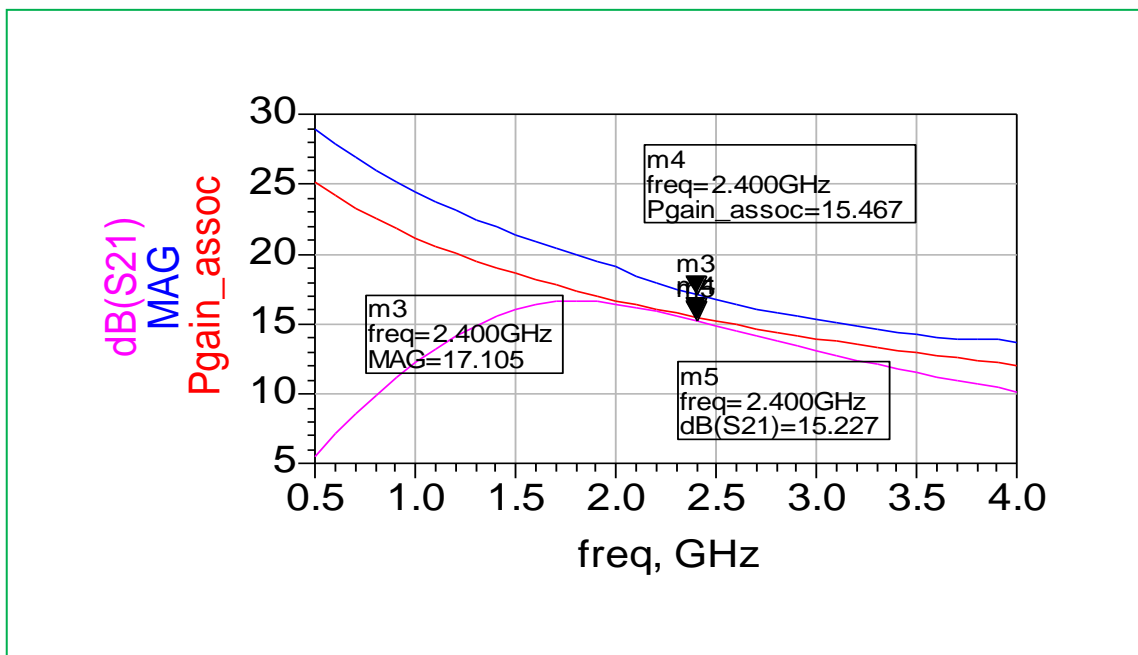


Figure 5.13: Gain Comparison

**4. Bandwidth Calculation:** As shown in the figure 5.8, the bandwidth is calculated by the help of gain graph, and the bandwidth of LNA is 500 MHz

## 5.7 LNA DESIGN WITH MICROSTRIP LINES COMPONENT

The main technical indexes of the LNA: working frequency from 2.40GHz to 2.50GHz, noise figure (NF) less than 2.0 dB, gain more than 12dB. Component choice is vital to design the LNA. According to the designing demand, Low noise PHEMT ATF-54143 manufactured by Avago has chosen and it is found that the minimum noise figure ( $F_{min}$ ) is close to 0.52dB and the gain is about 16.65dB when  $f = 2.40\text{GHz}$ ,  $V_{ds} = 3\text{V}$ ,  $I_{ds} = 20\text{ mA}$ . It can satisfy design demand, so this point is chosen as the quiescent point of ATF-54143. Bias circuit is shown in Figure 5.14.

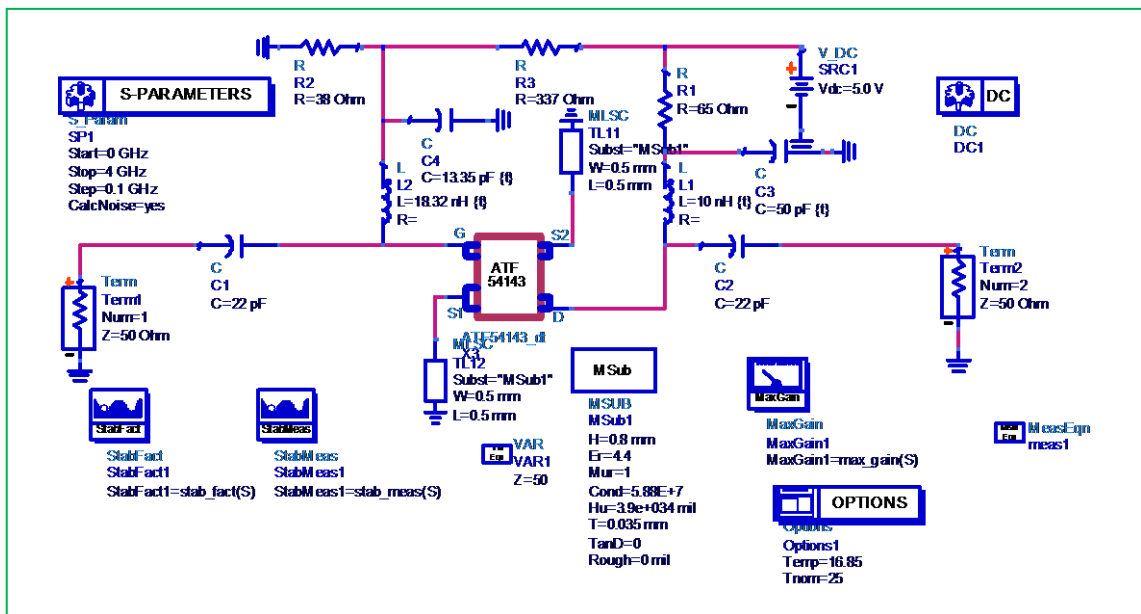


Figure 5.14: DC Bias circuit diagram

### 5.7.1 Stability of LNA

Because of the existence of reflected wave, RF amplifier will generate self-oscillation in some frequencies and cannot amplify signal. So stability is a very important factor we must take into consideration in the design of RF amplifier. We have to judge the amplifier is stable absolutely or not before designing matching circuit. Stability depends on S parameter of the transistor, matching network and biasing requirements. The sufficient and necessary conditions of absolute stability are

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

$$|\Delta| = |S_{11}S_{22} - S_{21}S_{12}| < 1$$

ATF-54143 isn't stable absolutely in some frequencies. In order to improve stability, negative feedback circuit is added in the source electrode. The circuit is shown in Figure 5.14. In the circuit, also short Microstrip lines of appropriate length between the source and ground has added. It acts as adding feedback inductance in the source. It is very complex that directly calculating stability figure by the datasheet, so one can use stability coefficient of the ADS software to analyse the stability of the amplifier. The simulation result is shown in Figure 5.15. From the figure it has found that K coefficient is greater than 1, so the transistor is stable absolutely in working frequencies.

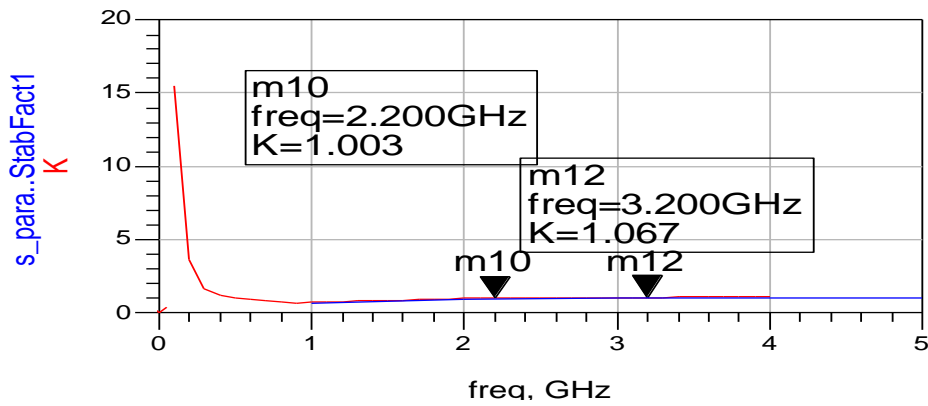


Figure 5.15: Stability factor

### 5.7.2 Design of Matching Network of Input Terminal

The design uses distributed parameter elements on account of high working frequency. It has to be balance gain and noise figure in designing because the input impedance when LNA has maximum gain is different that when LNA has minimal noise. The noise figure of receiving system is mainly determined by the first LNA, so the design uses minimal noise matching principle. In order to get minimal noise figure, reflection coefficient of input terminal must satisfy requirement of best source reflection coefficient ( $\Gamma_{opt}$ ). If termination impedance and characteristic impedance of Microstrip line is 50 ohm, matching network of input terminal has to convert  $\Gamma_{opt}^*$  into 50 ohm.

This design uses type L matching network. First, matching network of input terminal has designed with DA\_SmithChartMatch of ADS software, roughly determine its parameters approximately, and then optimize with tuning tool of ADS software. The matching network of input terminal is shown in Figure 5.16.

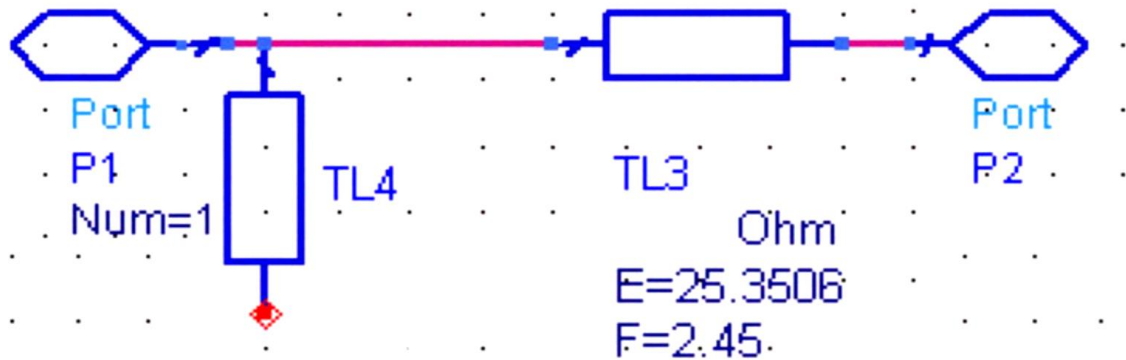


Figure 5.16: Matching network of input terminal

### 5.7.3 Design of Matching Network of Output Terminal

Matching network of output terminal is for increasing gain, improving gain flatness and input VSWR, so the design use maximum gain matching principle. Also matching network of output terminal has designed with DA\_SmithChartMatch of ADS software. The simulation results show that return loss of input terminal is good, but return loss of output terminal is bad. So, it has to optimize the whole circuit with Tuning tool of ADS software, find a balance in some performance parameters of the LNA such as  $S_{11}$ ,  $S_{22}$ ,  $S_{21}$ , noise figure. The matching network of output terminal is shown in Figure 5.17.

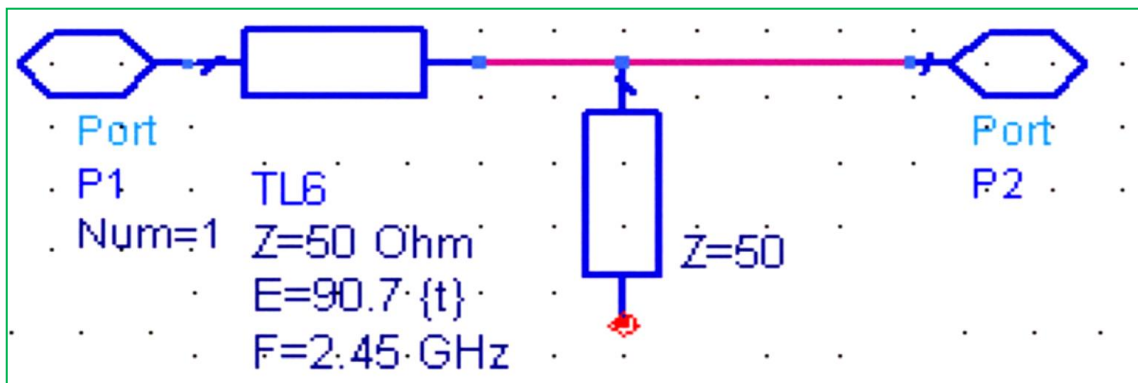


Figure 5.17: Matching network of output terminal

### 5.7.4 Determining the Size of Microstrip Line

FR<sub>4</sub> substrate has chosen for Microstrip lines. Its relative dielectric constant  $\epsilon_r = 4.4$ , thickness  $h = 0.8$  mm, permeability  $\mu_r = 1$ , metal conductivity  $Cond = 5.88e7$ ,

conductor thickness  $T = 0.035$  mm. Width and length of matching Microstrip line can be easily computed with Microstrip line computational tool LinCalc of ADS software. The schematic is shown in Figure 5.18.

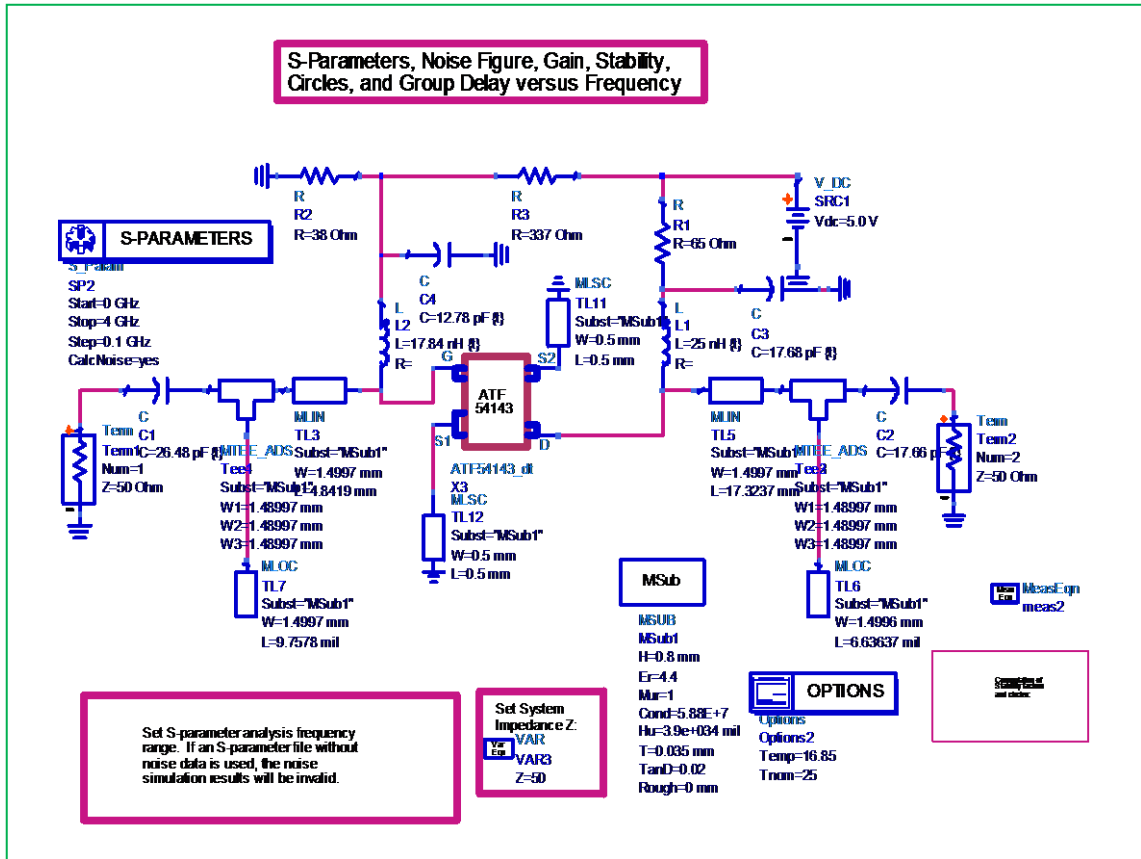


Figure 5.18: Schematic of Microstrip LNA

## 5.8 SIMULATION AND RESULT

The simulation results are shown in figure 5.19 to figure 5.24. From figure 5.19 and figure 5.20, it can be seen that in operating frequency range input reflection coefficient  $S_{11}$  and output reflection coefficient  $S_{22}$  are less than -13dB, and from figure 5.21 gain is more than 13dB, gain flatness is good and From Figure 5.23 noise figure is less than 0.7dB, stability factor is more than 1, the whole circuit is unconditionally stable.

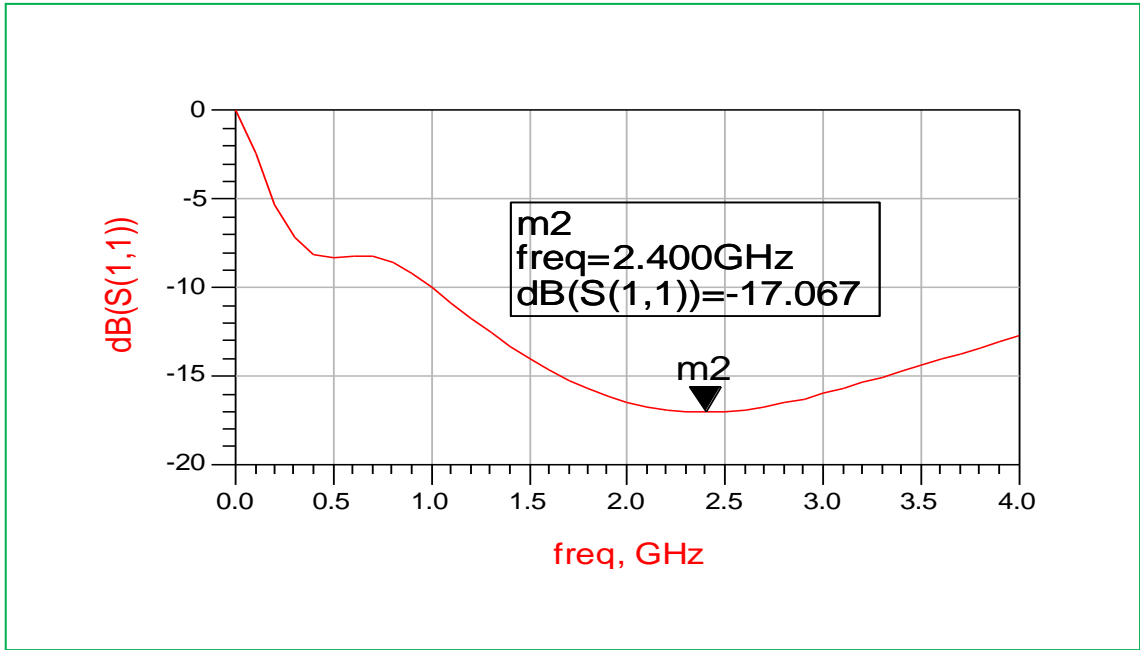


Figure: 5.19: Scattering parameter  $S_{11}$

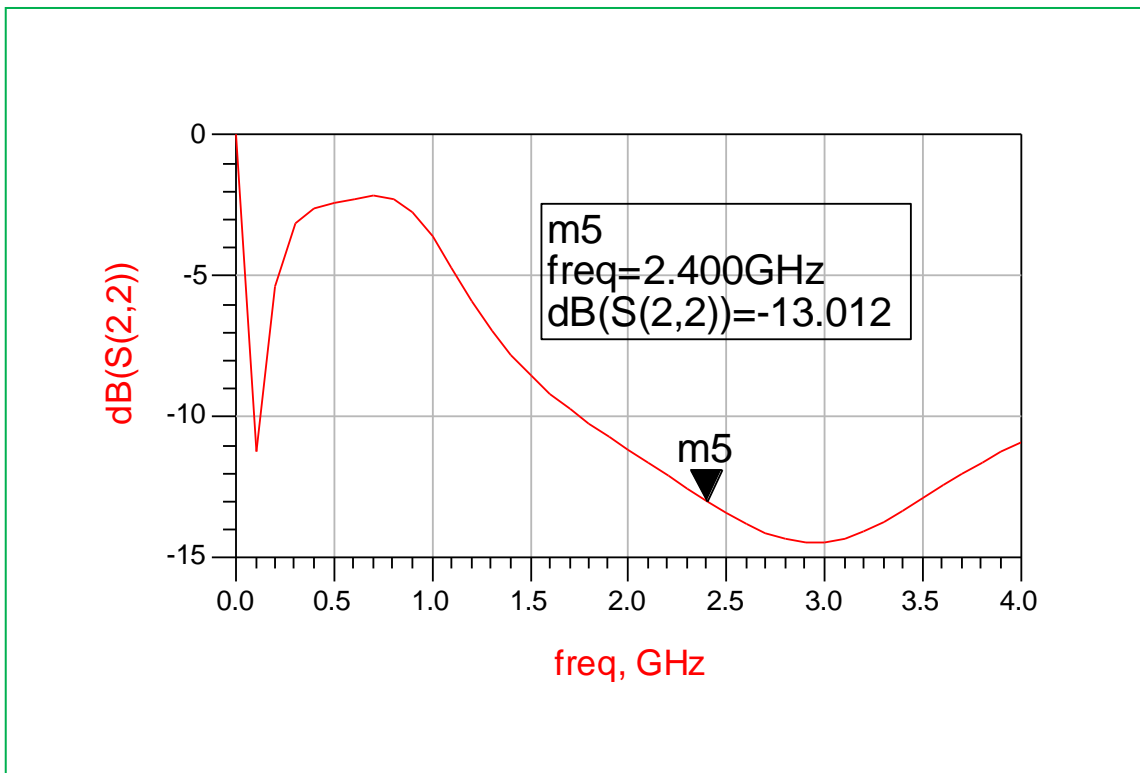


Figure 5.20: Scattering parameter  $S_{22}$

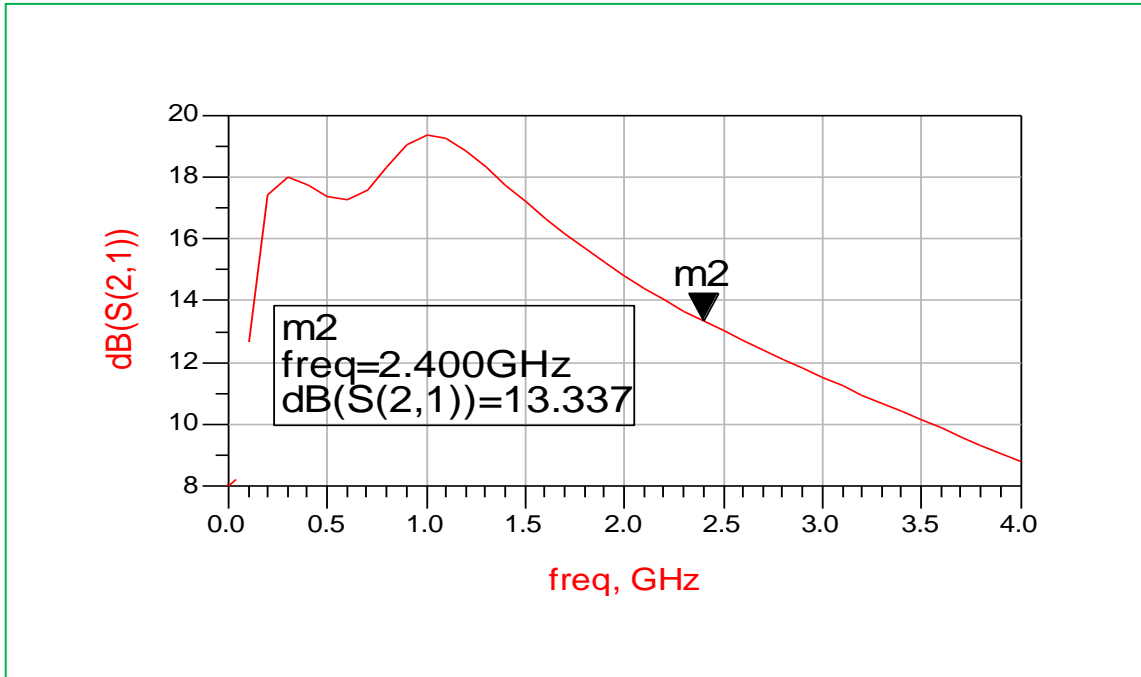


Figure 5.21: Forward Gain ( $S_{21}$ )

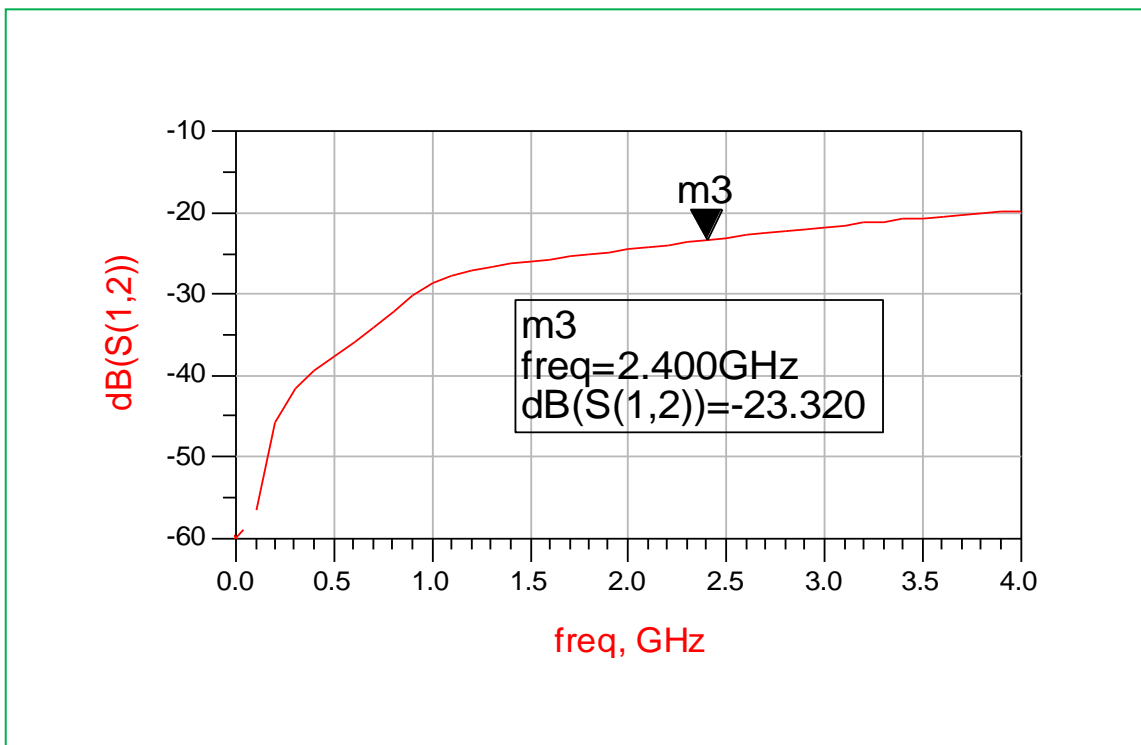


Figure 5.22: Scattering parameter  $S_{12}$



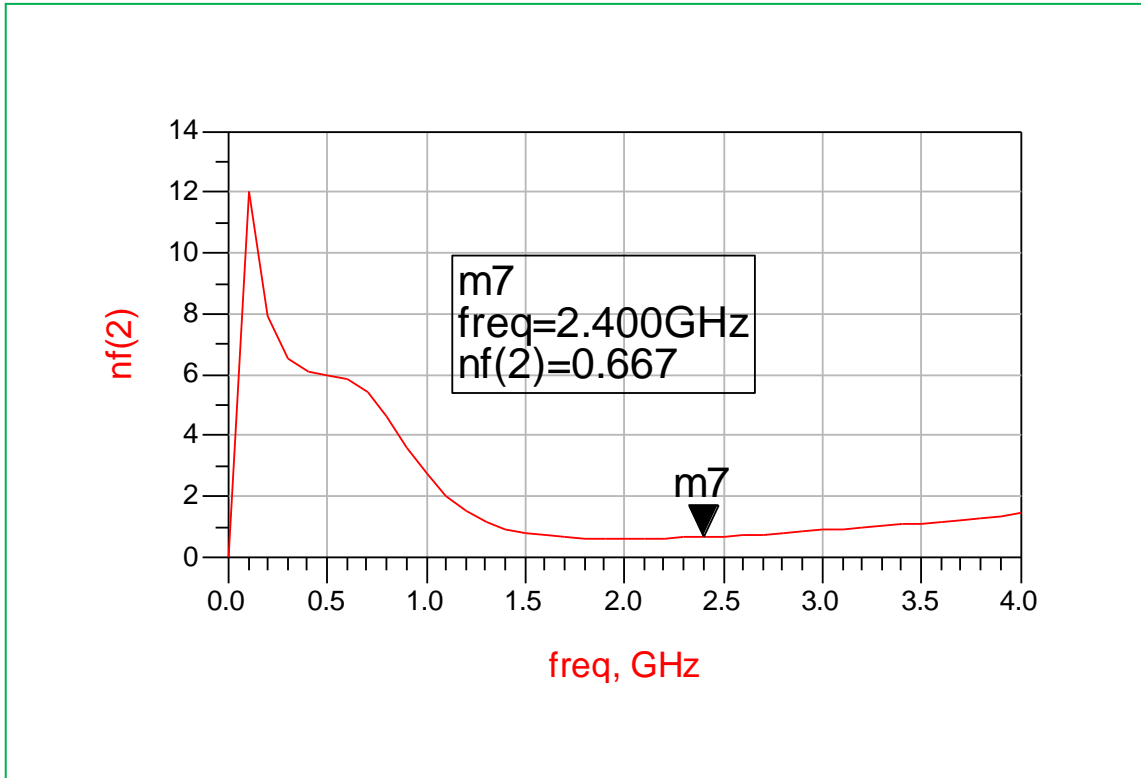


Figure 5.23: Noise Figure (NF)

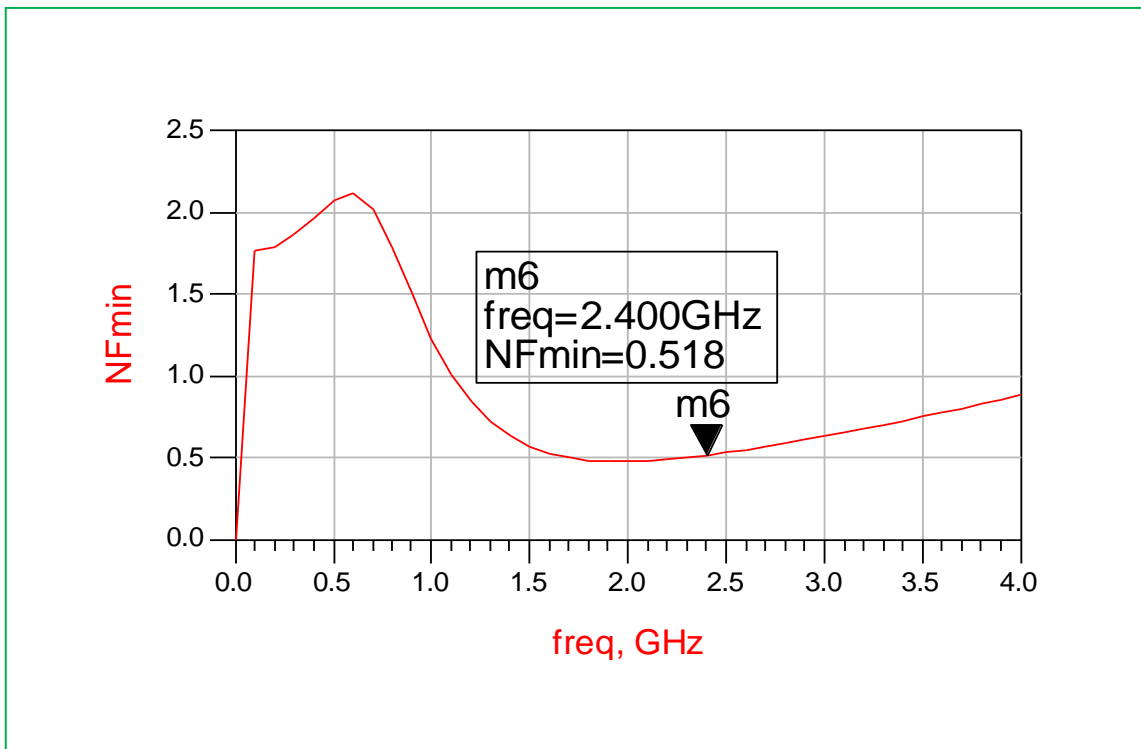


Figure 5.24: Minimum Noise Figure ( $NF_{min}$ )

## 5.9 RESULTS OBTAINED FROM SIMULATION

Based on the simulation result Table 5.3 shows the various performance parameters result of the design.

**Table 5.3: Performance Parameters**

Performance parameters	Using discrete component	Using Microstrip lines
$S_{11}$	-15.08 dB	-17.06 dB
$S_{12}$	-20.13 dB	-23.32 dB
$S_{21}$	15.27 dB	13.33 dB
$S_{22}$	-10.26 dB	-13.02 dB
Noise Figure(NF)	0.386 dB	0.667 dB
Minimum Noise Figure ( $NF_{min}$ )	0.379 dB	0.518 dB
Stability Factor(K)	1.009	1.067
Bandwidth	500 MHz	400 MHz

As shown in the above table the values of matching parameters  $S_{11}$ ,  $S_{12}$ , and  $S_{22}$  are -17 dB, -23dB and -13dB respectively, with targeted values as shown in table 5.1, of that is less than -10dB, so desired values have achieved. Targeted gain is greater than 12dB whereas gain of this design circuit is 13.33dB and noise figure must be less than 2dB as given in table 5.1, noise figure of this design is 0.667 dB and minimum noise figure is 0.518dB and all the other performance parameters values are better than target value.

# CHAPTER 6

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## CONCLUSION AND FUTURE WORK

### 6.1 CONCLUSION

In order to design any circuit, it is always better to abstract away its complexity. In this thesis design a low noise amplifier for 2.4 GHz applications covered all the important details required for the general low noise amplifier. And also a novel low noise amplifier is designed and simulated keeping view high gain and low noise figure for given frequency applications. This design used inductive source degeneration topology as it provided good matching and low noise figure as compared to other topology. Additionally, power constrained noise optimization technique was used for power and noise optimization. Since, HEMT Technology has become dominant for low noise amplifier design because of its several advantages such as low noise figure, high frequency of operation, low cost, low static power dissipation and low area, and this design of low noise amplifier used HEMT Technology. The LNA was designed to operate at 2.4 GHz frequency, simulated by using Advanced Design System designed by Agilent Technologies. The key issue in the design, Gain, noise figure, stability was considered. After the simulation it has achieved the Gain 15.23dB, Noise Figure 0.38dB, Stability Factor 1.009, input reflection coefficient -15dB, output reflection coefficient -10.26dB, reverse isolation coefficient -20.13dB, power consumption 180 mW with supply voltage 3.0 V using lumped components and Gain 13.33dB, Noise Figure 0.38dB, Stability Factor 1.009, input reflection coefficient -17dB, output reflection coefficient -13dB, reverse isolation coefficient -23.32dB, power consumption 60 mW with supply voltage 3.0 V using Microstrip lines. Their performance is compared in the simulation schematic. Optimization was performed according to get the desired responses in all the designs.

Furthermore, throughout the whole design, transistor was stable. The level of satisfaction of this thesis work is satisfactory. However, due to the parasitic effects and unavailability of required components, there are some deviations from expectations in

the simulated results. This thesis work gives a closer and wide view of all the relevant background theories and design technologies to the designer.

Table 6.1 shows the comparison of this work to some reported work according to operating frequency about 2-3 GHz.

**Table 6.1: Comparison Table**

<b>Parameters</b>	<b>[43]</b>	<b>[44]</b>	<b>[45]</b>	<b>This Work</b>
Frequency(GHz)	2.40	2.40	2.45	2.40
Supply Voltage (V)	1.2	1.5	0.9	3.0
Gain (dB)	13	21.4	15.3	15.2
Noise Figure (dB)	3.6	5.2	3.34	0.38
Input Reflection (dB)	-13.1	-19.0	-16.5	-15.1
Output Reflection (dB)	-11.5	--	--	-10.2
Isolation(dB)	--	--	--	-20.1

## 6.2 FUTURE WORKS

Due to less time period in this project work it could not possible to do more work but in future, for investigating design of Low Noise Amplifier for RF application following work may be extended:

- This work has been implemented as a schematic design, furthermore it can be continued to layout implementation and perform post layout simulation, and could be fabricated and used some wireless applications at 2.4 GHz.
- Power dissipation could have been minimized by body biasing of transistor used for designing of LNA.
- Spiral inductor occupy large area so either inductor less or active inductors may be used for design purpose.

- Band gap reference generators may be used instead of resistors for providing bias as they produce less noise and are temperature independent.

Though the responses are satisfactory, but still there are scopes to improve the performances. Some of the circuits can be improved in design and with more proper optimization to have better responses. In future, different classes of BPFs such as elliptical, Chebyshev with different orders can be designed with LNA, and LNA can be designed combined with antenna also. It will reduce the size of the device. This will provide more options to compare for the better one. Furthermore, exact values of components which were used in the designs can be purchased and made new prototypes, which may produce better responses of LNA (designed with lumped components). However, the acquired knowledge from this thesis work can help to design the whole RF receiver system in the ISM band.

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## Appendix A

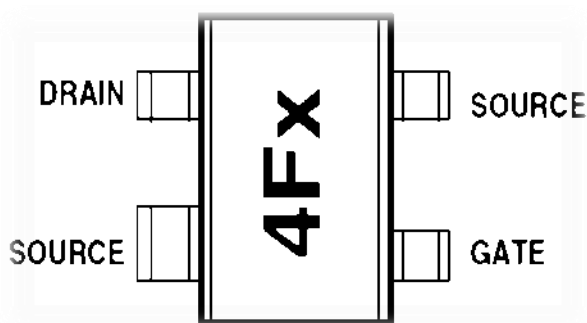
### ATF-54143

Low Noise Enhancement Mode Pseudomorphic HEMT in a Surface Mount Plastic Package

### Description

Avago Technologies' ATF-54143 is a high dynamic range, low noise, E-PHEMT housed in a 4-lead SC-70 (SOT-343) surface mount plastic package. The combination of high gain, high linearity and low noise makes the ATF-54143 ideal for cellular/PCS base stations, MMDS, and other systems in the 450 MHz to 6 GHz frequency range.

### Pin Connections and Package Marking



### Features

- High linearity performance
- Enhancement Mode Technology
- Low noise figure
- Excellent uniformity in product specifications
- 800 micron gate width
- Low cost surface mount small plastic package SOT-343 (4 lead SC-70)
- Tape-and-Reel packaging option available
- Lead-free option available

## Specifications

- 2 GHz; 3V, 60 mA (Typ.)
- 36.2 dBm output 3rd order intercept
- 20.4 dBm output power at 1 dB gain compression
- 0.5 dB noise figure
- 16.6 dB associated gain

## Applications

- Low noise amplifier for cellular/PCS base stations
- LNA for WLAN, WLL/RLL and MMDS applications
- General purpose discrete E-PHEMT for other ultra low noise applications

## ATF-54143 Electrical Specifications

$T_A = 25^\circ\text{C}$ , RF parameters measured in a test circuit for a typical device

Symbol	Parameter and Test Condition	Units	Min.	Typ.	Max.
$V_{gs}$	Operational Gate Voltage $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$	V	0.4	0.59	0.75
$V_{th}$	Threshold Voltage $V_{ds} = 3\text{V}$ , $I_{ds} = 4\text{ mA}$	V	0.18	0.38	0.52
$I_{dss}$	Saturated Drain Current $V_{ds} = 3\text{V}$ , $V_{gs} = 0\text{V}$	$\mu\text{A}$		1	5
$G_m$	Transconductance $V_{ds} = 3\text{V}$ , $g_m = \Delta I_{dss} / \Delta V_{gs}$ $\Delta V_{gs} = 0.75 - 0.7 = 0.05\text{V}$	mho	230	410	560
$I_{gss}$	Gate Leakage Current $V_{gd} = V_{gs} = -3\text{V}$	$\mu\text{A}$			200
NF	Noise Figure $f = 2\text{ GHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$ $f = 900\text{ MHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$	dB		0.5 0.3	0.9
$G_a$	$f = 2\text{ GHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$ $f = 900\text{ MHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$	dB	15	16.6 23.4	18.5
OIP3	$f = 2\text{ GHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$ $f = 900\text{ MHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$	dBm	33	36.2 35.5	
P1dB	$f = 2\text{ GHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$ $f = 900\text{ MHz}$ , $V_{ds} = 3\text{V}$ , $I_{ds} = 60\text{ mA}$	dBm		20.5 18.4	

**ATF-54143 Typical Scattering Parameters, VDS = 3V, IDS = 60 mA**

Freq. GHz	$S_{11}$		dB	$S_{21}$		$S_{12}$		$S_{22}$		MSG/MAG dB
	Mag.	Ang.		Mag.	Ang.	Mag.	Ang.	Mag.	Ang.	
0.1	0.99	-18.9	28.84	27.66	167.6	0.01	80.0	0.54	-14.0	34.42
0.5	0.81	-80.8	26.04	20.05	128.0	0.03	52.4	0.40	-58.8	28.25
0.9	0.71	-117.9	22.93	14.01	106.2	0.04	41.8	0.29	-83.8	25.44
1.0	0.69	-124.4	22.24	12.94	102.2	0.05	40.4	0.27	-88.5	24.13
1.5	0.64	-149.8	19.40	9.34	86.1	0.05	36.1	0.21	-105.2	22.71
1.9	0.62	-164.9	17.66	7.64	75.6	0.06	33.8	0.17	-114.7	21.05
2.0	0.62	-168.3	17.28	7.31	73.3	0.06	33.3	0.17	-117.0	20.86
2.5	0.60	176.2	15.58	6.01	61.8	0.07	30.1	0.13	-129.7	19.34
3.0	0.60	162.3	14.15	5.10	51.0	0.08	26.5	0.11	-146.5	18.04
4.0	0.62	137.1	11.81	3.90	30.8	0.09	17.1	0.10	165.2	14.87
5.0	0.66	115.5	9.87	3.11	11.7	0.11	6.8	0.14	131.5	13.27
6.0	0.69	97.2	8.22	2.58	-6.4	0.12	-3.9	0.18	112.4	11.72
7.0	0.70	80.2	6.85	2.20	-24.0	0.13	-15.8	0.20	94.3	10.22
8.0	0.72	62.2	5.58	1.90	-41.8	0.14	-28.0	0.23	70.1	9.02
9.0	0.76	45.0	4.40	1.66	-59.9	0.15	-39.6	0.29	50.6	8.38
10.0	0.83	28.4	3.06	1.42	-78.7	0.15	-55.1	0.38	36.8	8.71
11.0	0.85	13.9	1.60	1.20	-95.8	0.15	-68.6	0.46	24.4	7.55
12.0	0.88	-0.2	0.43	1.05	-111.1	0.15	-80.9	0.51	11.3	7.55
13.0	0.89	-14.6	-0.65	0.93	-128.0	0.15	-94.9	0.55	-5.2	6.70
14.0	0.88	-30.6	-1.98	0.80	-146.1	0.14	-109.3	0.61	-20.8	5.01
15.0	0.88	-45.0	-3.62	0.66	-162.7	0.13	-122.9	0.66	-35.0	3.73
16.0	0.88	-54.5	-5.37	0.54	-176.6	0.12	-133.7	0.70	-45.8	2.54
17.0	0.88	-62.5	-6.83	0.46	171.9	0.12	-143.2	0.73	-56.1	1.57
18.0	0.92	-73.4	-8.01	0.40	157.9	0.11	-156.3	0.76	-68.4	2.22

**Typical Noise Parameters, VDS = 3V, IDS = 60 mA**

<b>Freq GHz</b>	<b>F<sub>min</sub> dB</b>	<b>Γ<sub>opt</sub> Mag.</b>	<b>Γ<sub>opt</sub> Ang.</b>	<b>R<sub>n/50</sub></b>	<b>G<sub>a</sub> dB</b>
0.5	0.15	0.34	42.3	0.04	28.50
0.9	0.20	0.32	62.8	0.04	24.18
1.0	0.22	0.32	67.6	0.04	23.47
1.9	0.42	0.27	116.3	0.04	18.67
2.0	0.45	0.27	120.1	0.04	18.29
2.4	0.52	0.26	145.8	0.04	16.65
3.0	0.59	0.29	178.0	0.05	15.56
3.9	0.70	0.36	-145.4	0.05	13.53
5.0	0.93	0.47	-116.0	0.10	12.13
5.8	1.16	0.52	-98.9	0.18	11.10
6.0	1.19	0.55	-96.5	0.20	10.95
7.0	1.26	0.60	-77.1	0.37	9.73
8.0	1.63	0.62	-56.1	0.62	8.56
9.0	1.69	0.70	-38.5	0.95	7.97
10.0	1.73	0.79	-21.5	1.45	7.76

# An Unconditionally Stable Front End Low Noise Amplifier Design for 2.4 GHz ISM Band

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## ABSTRACT

This paper presents the design and simulation of a 2.4 GHz ISM band front end single stage Low Noise Amplifier (LNA) for wireless transceiver system. This amplifier uses AVAGO ATF-54143 transistor which is a low noise and high dynamic range Pseudomorphic high electron mobility transistor. The proposed method is addressed to optimize noise performance and power efficient while maintaining good input and output matching. The design simulation has been performed using Advance Design Simulation (ADS) software. The designed LNA offers minimum noise figure less than 0.38 dB, forward gain ( $S_{21}$ ) greater than 15 dB, input return loss less than -15 dB and output return loss ( $S_{22}$ ) less than -10 dB for the frequency range of 2.0 to 3.0 GHz. The Designed LNA can be used in various applications like Bluetooth, WI-FI, RFID, digital cordless telecommunication.

## Keywords

Advance Design System, Pseudomorphic high electron mobility transistor, Forward Gain, Low Noise Amplifier, Noise Figure, Stability, Wireless LAN.

## 1. INTRODUCTION

The fast progress of wireless communications emerges consumer equipment in low-cost and high-performance requirements, especially for portable devices such as handsets, Bluetooth devices. Light weight and long working time under limited battery capacity are major concerns, pushing the circuit designs to low power operations. In addition, some special demands from wireless sensor applications require the circuits working at very low supply voltage conditions.

Among the circuits of wireless communication equipment, the low noise amplifier plays an important role in the overall performance of a RF receiver. It is the first component in any RF receiver. The function of low noise amplifier (LNA) is to amplify low-level signals coming from antenna with as low as possible noise added by circuit, provides enough amplification and minimum degradation of signal-to-noise ratio. Additionally, for large signal levels, the low noise amplifier will amplified the received signal without introducing any noise, hence eliminating channel interference. LNA is located at the first stage of microwave receiver and it has dominant effect on the noise performance of the overall system. In LNA design, it is necessary to compromise its simultaneous requirements for high gain, low noise figure, stability, good input and output matching. The proposed LNA design is carried out with a systematic procedure and simulated by Advanced Design System (ADS2008) designed by Agilent. In

the LNA circuit designing, IEEE 802.11 stand-art is used as references. The low noise amplifier system should meet this standard requirement and should operate properly for wireless applications.

## 2. SELECTION OF TRANSISTOR

In this design Agilent's ATF-54143 PHEMT has been used [1]. It is a high dynamic range, low noise, Pseudomorphic HEMT (High Electron Mobility Transistor) housed in a surface mount plastic package. Based on its featured performance, ATF-54143 is suitable for low noise amplifier application in cellular and personal communication service (PCS) base stations, low earth orbit(LEO) systems, Bluetooth devices and other systems requiring super low noise figure with good intercept in the 450MHz to 10 GHz frequency range. Based on S-parameters of the transistor and certain performance requirements, a systematic procedure is developed for the design of LNA. In LNA design, the most important factors are low noise, moderate gain, matching and stability [2-3]. Besides these factors, power consumption and layout design size also need to be considered in designed works.

## 3. CIRCUIT ANALYSIS USING ADS

To design an amplifier, the input and output matching network are considered to achieve the required stability, noise-figure, and a bandwidth. Super high frequency amplifier is a typical active circuit used to amplify the amplitude of radio frequency signal. Using active or passive biasing, the basic amplifier design is shown in Fig 1.

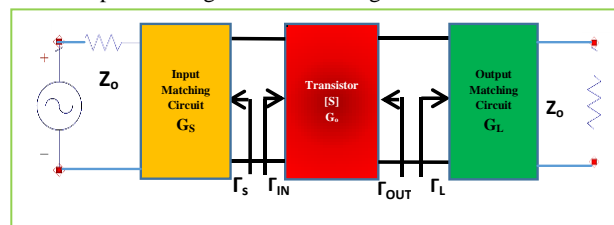


Fig.1. Block diagram of a single stage two port LNA model

Fig.1 shows a process for single-stage amplifier design including input/output matching networks. Analog integrated circuit designers accustomed to working with lower frequency circuits tend to have only a passing familiarity with two staples of traditional RF design as Smith charts and S parameters.

$$\Gamma_s = \frac{Z_s - Z_0}{Z_s + Z_0}, \Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (1)$$

The relationship between the normalized load impedance, source and load reflection coefficient are given by equation (1). The LNA design formula and equation were referred to [4]. Input/output matching circuit is essential to reduce the unwanted reflection of signal and to improve efficiency of the transmission from RF source to load.

### 3.1 Power Gain

The power gain is the ratio of power actually delivered to some load and the power actually delivered by source. Three widely used definitions for the power gain of the two-port network are the transducer power gain  $G_T$ , the available power gain  $G_A$ , and the power gain  $G_P$ , also called the operating gain. The two-port network is shown in Fig. 2

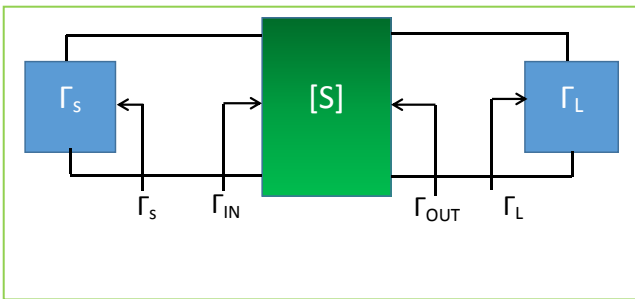


Fig.2. General two-port network

The matrix elements are  $S_{11}$ ,  $S_{12}$ ,  $S_{21}$  and  $S_{22}$  which are referred to as the scattering parameters.  $\Gamma_s$ ,  $\Gamma_L$  will lead to a stable input and output impedances.

### 3.2 Transducer Power Gain

Transducer Power Gain is the ratio of  $P_{Avg}$ , maximum power available from source to  $P_L$ , power delivered to the load [4-7].

$$G_T = \frac{\text{Output power of network}}{\text{max input power}} = \frac{P_L}{P_{avg}}$$

$$G_T = \frac{1 - |\Gamma_s|^2}{|1 - S_{11}\Gamma_s|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - \Gamma_{out}\Gamma_L|} \quad (2)$$

### 3.3 Available Power Gain

Available power gain is the ratio of  $P_{avg}$  power available from the source, to  $P_{avn}$ , power available from two-port network.

$$G_A = \frac{\text{Output power of network}}{\text{max.input power}} = \frac{P_{avn}}{P_{avg}}$$

$$G_A = \frac{1 - |\Gamma_s|^2}{|1 - S_{11}\Gamma_s|^2} |S_{21}|^2 \frac{1}{|1 - \Gamma_{out}|^2} \quad (3)$$

### 3.4 Operating Power Gain

Operating power gain is the ratio of power  $P_L$  delivered to the load to power  $P_{in}$  supplied to two-port network.

$$G_P = \frac{\text{Output power of network}}{\text{input power of network}} = \frac{P_L}{P_{in}}$$

$$G_P = \frac{1}{|1 - \Gamma_{in}|^2} |S_{21}|^2 \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|} \quad (4)$$

### 3.5 Noise Figure

Every device is a source of internally generated noise. The noise entering the device and the internal noise must be added to obtain the total input system noise. If the device is an amplifier, the total system noise power will be amplified the output by the gain of the device. If the output load is matched, this gain will be the available gain. The signal to noise ratio of input port to that of output port is referred as noise figure and is larger than 1 dB [2].

$$F = F_{Min} + \frac{4r_n |\Gamma_s - \Gamma_{opt}|^2}{(1 - |\Gamma_s|^2) |1 + \Gamma_{opt}|^2} \quad (5)$$

Transistor noise factor  $F$  is a function of  $\Gamma_s$ ,  $F_{min}$ ,  $R_n$ , and  $\Gamma_{Opt}$  where  $F_{min}$ ,  $R_n$ , and  $\Gamma_{Opt}$  are known as the transistor noise parameters and are usually known.

### 3.6 Simultaneous Conjugate Matching

The transducer, available, and operating power gains become equal to the maximum available gain ( $G_{max}$ ) when both the generator and the load are conjugate matched to the two-port, i.e.  $\Gamma_{in} = \Gamma_s^*$  and  $\Gamma_L = \Gamma_{out}^*$ .

$$\Gamma_s^* = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \quad (6)$$

$$\Gamma_L^* = S_{22} + \frac{S_{12}S_{21}\Gamma_s}{1 - S_{22}\Gamma_s} \quad (7)$$

## 4. CIRCUIT ANALYSIS USING ADS

Transistor must be biased at appropriate operating point before used. So that, transistor can works under values required and achieve less power consumption. In this project, passive biasing method is adopted. The component readings are determined with reference from datasheet. By referring datasheet, data of  $V_{DS} = 3V$  and  $I_{DS} = 60mA$  had be chosen because it is believed that it can give optimum values in gain and noise figure, with  $V_{GS} = 0.52V$ ,  $I_{BB} = 2mA$ ,  $V_{DS} = 3V$  and  $I_{DS} = 60mA$ .

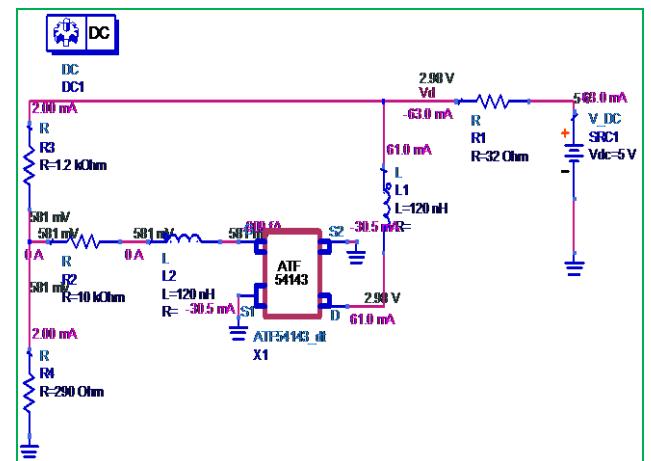


Fig.3. DC Biasing circuit

The Values of resistance  $R_1$ ,  $R_3$  and  $R_4$  can be calculated as:

$$R_3 = \frac{V_{GS}}{I_{BB}} = 260\Omega$$



$$R_4 = \frac{V_{DD} - V_{DS}}{I_{DS} + I_{BB}} = 32.26 \Omega$$

$$R_1 = \frac{(V_{DS} - V_{GS})R_3}{V_{GS}} = 1.24k\Omega$$

The  $R_1$ ,  $R_3$  and  $R_4$  in circuit are slightly adjusted from calculation readings in order to obtain better  $V_{DS}$  and  $I_{DS}$  reading in the simulation.

#### 4.1 Stability Consideration

The stability of an amplifier or its resistance to oscillate is an important consideration in a design. It can be determined from the S parameters. Oscillation occurs when  $|\Gamma_{in}| > 1$  or  $|\Gamma_{out}| > 1$ . This is due to the dependence of and on the source and load matching networks. By using S-parameters of ATF-54143 PHEMT for  $V_{DS} = 3V$  and  $I_{DS} = 60$  mA, Stability Factor K and  $\Delta$  can be calculated. For unconditionally stable LNA,

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1 \quad (8)$$

$$|\Delta| = |S_{11}S_{22} - S_{21}S_{12}| < 1 \quad (9)$$

For the frequency range 2 GHz to 3 GHz, it can be shown that

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

The designed LNA with matching network and DC-biasing network is shown in Fig.4. It has L-type matching networks at input and output which have been designed at 2.4 GHz using Quick Smith software and tuned for the frequency range 2 GHz to 3 GHz. The DC-biasing network has been designed using RF chokes and dc-blocking capacitors.

#### 5. SIMULATION RESULTS

For the proposed LNA shown in Fig.4, simulation results are shown in Fig. 5 to 10. The designed low noise amplifier offers forward gain  $S_{21} > 15$  dB, input return loss  $S_{11} < -15$  dB, output return loss  $S_{22} < -10$  dB and minimum noise figure  $NF_{min} < 0.38$  dB, for the frequency range of 2 GHz to 3 GHz which are shown in Fig.5, Fig.7, Fig.9 and Fig.10 respectively. The Stability Factor plot is shown in Fig.8. The Stability Factor  $K > 1$  for 2 GHz to 3 GHz which indicates that LNA is unconditionally stable in this frequency range.

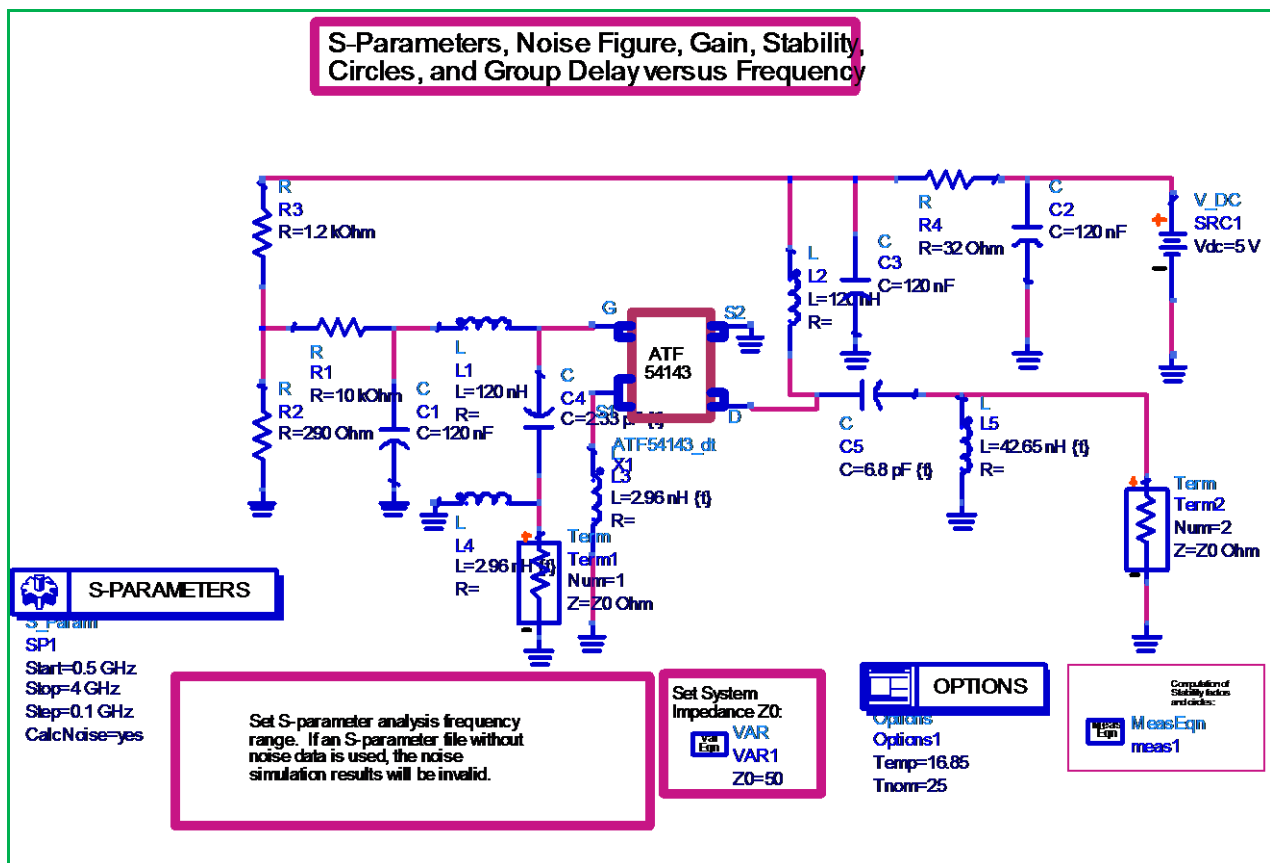


Fig 4: Complete LNA circuit after input and output matching

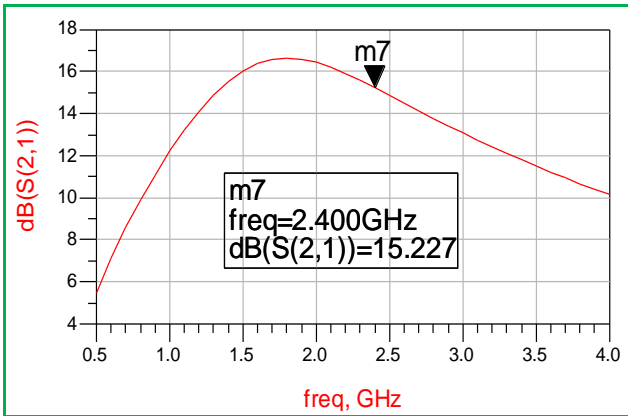


Fig.5. Forward Gain ( $S_{21}$ )

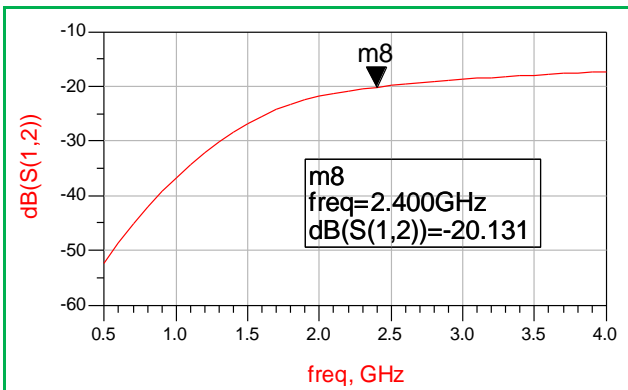


Fig.6. Isolation ( $S_{12}$ )

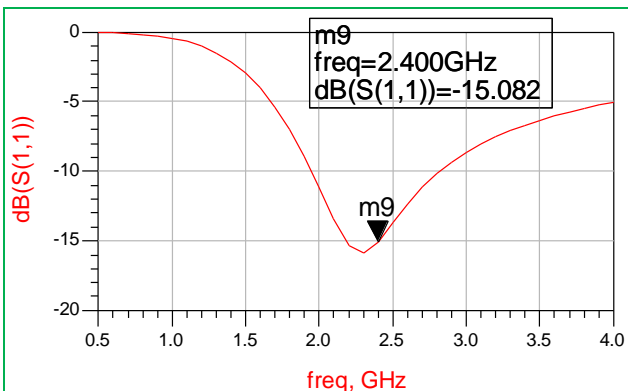


Fig.7. Input Reflection Coefficient ( $S_{11}$ )

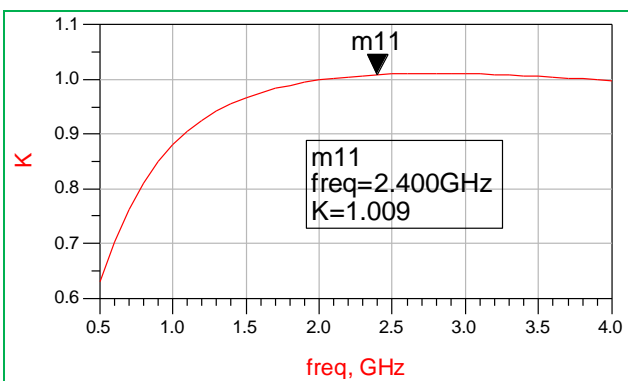


Fig.8. Stability Factor K

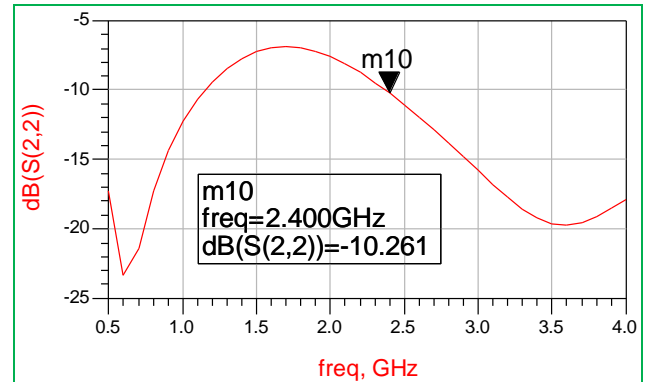


Fig.9. Output Reflection Coefficient ( $S_{22}$ )

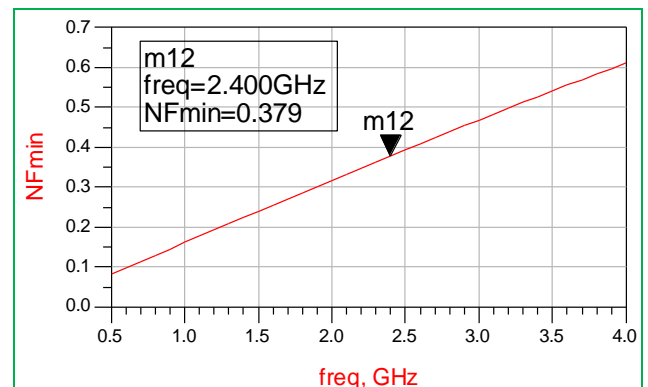


Fig.10. Minimum Noise Figure

## 6. CONCLUSIONS

A 2.4 GHz single stage low noise amplifier is design and simulated. The input insertion loss  $S_{11}$  is -15 dB and the output insertion loss  $S_{22}$  is -10 dB. The better performance in NF of the amplifier is achieved. The synthesized LNA guided by the proposed methodology achieves up to 15 dB power gain with a suppressed NF as low as 0.38 dB and provides good input and output matching over 2 GHz to 3 GHz. A 2.4 GHz single stage low noise amplifier is used in IEEE 802.11b and 802.11g standards for wireless local area network (WLAN).most popular application are Bluetooth. We can also use the designed LNA for the purpose of RF Transceiver systems [5-6]. This can also suitable for the switching circuits.

## 7. ACKNOWLEDGMENTS

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