VYSOKÉ UČENÍ TECHNICKÉ V BRNĚ
BRNO UNIVERSITY OF TECHNOLOGY

FAKULTA ELEKTROTECHNIKY A KOMUNIKAČNÍCH TECHNOLOGIÍ
ÚSTAV RADIOELEKTRONIKY

FACULTY OF ELECTRICAL ENGINEERING AND COMMUNICATION
DEPARTMENT OF RADIO ELECTRONICS

NÍZKOŠUMOVÝ ZESILOVAČ PRO PÁSMO UHF

LOW NOISE AMPLIFIER FOR UHF BAND

BAKALÁŘSKÁ PRÁCE
BACHELOR'S THESIS

AUTOR PRÁCE Ivan Alejandro Benites Ayala
AUTHOR

VEDOUCÍ PRÁCE prof. Ing. Miroslav Kasal, CSc.
SUPERVISOR

BRNO, 2016

VYSOKÉ UČENÍ TECHNICKÉ V BRNĚ

Fakulta elektrotechniky a komunikačních technologií

Ústav radioelektroniky

Bakalářská práce

bakalářský studijní obor

Elektronika a sdělovací technika

Student: Ivan Alejandro Benites Ayala ***ID:*** 159274

Ročník: 3 ***Akademický rok:*** 2015/16

NÁZEV TÉMATU:

Nízkošumový zesilovač pro pásmo UHF

POKYNY PRO VYPRACOVÁNÍ:

Prostudujte teoretický návrh nízkošumového zesilovače (LNA) pro pásmo UHF 430 - 440 MHz s moderními aktivními prvky. Zaměřte se na šumové přizpůsobení a vyšetřování stability zesilovače. Seznamte se s možností návrhu a simulace nízko úrovňových zesilovačů v návrhovém a simulačním prostředí ANSYS Designer. Zisk zesilovače nechť je 15 až 20 dB, vstup a výstup nesymetrický s impedancí 50 Ohmů.

Zvažte otázku selektivních obvodů na vstupu zesilovače. Navrhněte desku plošného spoje a konstrukční řešení. Zesilovač realizujte. Změřte jeho parametry a porovnejte je s teoretickými předpoklady.

DOPORUČENÁ LITERATURA:

[1] HANUS, S., SVAČINA, J. Vysokofrekvenční a mikrovinná technika. Skriptum. Brno: FEKT, VUT v Brně, 2002.

[2] ROHDE, U.L., NEWKIRK, D.P. RF/Microwave Circuit Design for Wireless Applications. Hoboken: John Wiley & Sons, 2000.

Termín zadání: 8. 2. 2016 ***Termín odevzdání:*** 26. 5. 2016

Vedoucí práce: prof. Ing. Miroslav Kasal, CSc.

Konzultanti bakalářské práce:

doc. In Tomáš Kratochvíl, Ph.D.

předseda oborové rady

UPOZORNĚNÍ:

ABSTRACT

This bachelor's thesis presents the design of a low noise amplifier for UHF band $430 - 440$ MHz. This thesis is mainly focused on stability study and matching networks. For this design is used Ansoft Designer program, through which it is possible to simulate the low noise amplifier. The ATF-54143 transistor is used as main device for this design. Morover, there is designed a band pass filter, which satisfies the requirements for this low noise amplifier.

KEYWORDS

Low noise amplifier, ATF-54143 transistor, operating point, stability, matching network, noise figure, band pass filter.

BENITES AYALA, Ivan Alejandro. *Low Noise Amplifier for UHF*. Brno: Brno University of Technology, Faculty of Electrical Engineering and Communication, Department of Radio Electronics, 2016. 53 s. Bachelor‘s thesis. Supervisor: prof. Ing. Miroslav Kasal, CSc.

DECLARATION

I declare that I have worked on this thesis independently, using only the primary and secondary sources listed in the bibliography at the end of the work.

I further declare that, as the author of this term project, I did not infringe the copyrights of third parties, in particular, I did not intervene in the personal and property copyright of others, and I am fully aware of the consequences of violation of § 11 et seq Law no. 121/2000 Coll., on copyright, rights related to copyright and amending some laws (Copyright Act), as amended, including possible criminal consequences arising from the provisions of Part II, Title VI. Part 4 of the Penal Code no. 40/2009.

Brno ....

(Author's signature)

ACKNOWLEDGEMENT

I would like to thank my supervisor prof. Ing. Miroslav Kasal for his help, guidance, and valuable advice.

Faculty of Electrical Engineering
and Communication

Brno University of Technology
Technicka 12, CZ-61600 Brno, Czech Republic

http://www.six.feec.vutbr.cz

Experimentální část této diplomové práce byla realizována na výzkumné infrastruktuře vybudované v rámci projektu CZ.1.05/2.1.00/03.0072
Centrum senzorických, informačních a komunikačních systémů (SIX)
operačního programu Výzkum a vývoj pro inovace.

TABLE OF CONTENTS

INDEX OF ILLUSTRATIONS

INDEX OF TABLES

INTRODUCTION

A Low Noise Amplifier (LNA) is one of the basic functional blocks in radio frequency communication systems. The main function of the LNA in terms of the input of the analog processing chain is to amplify the signal without adding significant noise.

The main objective of this bachelor's thesis is to design a Low Noise Amplifier (LNA) for UHF band 430-440MHz, with a gain or amplification of 15-20dB, a minimum insertion noise, and input and output impedance of 50 ohms.

The first part discusses essential parameters and concepts for designing a LNA, such as noise figure, scattering parameters, stability, power gain, impedance matching networks, and bias circuits. In section 1.3 of the theoretical part, a band pass filter based on quarter-wave (λ/4-wave) coaxial resonator is described.

The second part is divided in four sub-sections where the proper design of the LNA and the Band-Pass Filter is described. This part also focuses on the complete simulation of both of them. Finally, the results regarding the realization and measurement of the LNA are presented. ATF-54143 transistor is used as a main device in the LNA designing. Simulation of the LNA has been performed using Ansoft Designer Program.

1 THEORETICAL PART

1.1 About Low-Noise Amplifier (LNA)

A Low-noise amplifier (LNA) is an electronic amplifier, which is designed to have some particular characteristics different from other amplifiers. The LNA generally is used in RF reception systems to amplify a very low-power signal and it's connected very close to the reception device (antenna) to reduce losses in the supply line.

In the design of LNA it will be essential to consider the noise, which comes from the low-power signals and LNA's components. Therefore, the LNA will reduce the noise while amplifying the useful signal, thereby the amplification of the desired signal is obtained for its post-processing.

1.2 Parameters of Low-Noise Amplifier (LNA)

1.2.1 Noise Factor, Noise Figure and Noise Temperature

The noise factor (F) is a figure that quantifies the noise power originating internally in a two-port network. The noise factor of a network can be defined as the radio of the available noise power from the two-port network with the internal noise sources to the available noise power without the internal noise sources [4] namely:

$$F = \frac{N_{out}}{kTBG_A}, \tag{1.1}$$

Where N_{out} is the available noise power available from the two-port's output, k is the Boltzmann constant, T_0 is the temperature (standard 290 K), B is the bandwidth used in the measurement and G_A is the two-port network available gain. Noise figure (NF) is the noise factor expressed in dB.

$$NF = 10 \log F \text{ [dB]}, \tag{1.2}$$

Yet, there is still another measure of the noise of a two-port network. Equivalent noise temperature T_e characterizes the noise properties of a two-port network in terms of kelvins. The relation between T_e and F is:

$$F = 1 + \frac{T_e}{T_0}, \tag{1.3}$$

Where T_0 is room temperature (290 K). Measure of the noise occurring in amplifiers is often performed using the Y-Factor method [5].

1.2.2 Scattering Parameters

In high frequency networks, direct measurements of voltages and currents become problematic. Therefore incident, reflected, and transmitted voltage waves facilitate the description of these networks [6]. The scattering parameters (S-parameters) relate these waves and provide representations of the high frequency networks in RF electronics. S-parameters are also widely used for modeling of passive and active components. Furthermore, the convertibility of S-parameters allows the derivation of ABCD-, Z- and Y-parameters [7].

If an amplifier is considered as a two-port network as shown in Figure 1.1 the scattering matrix representing such a network has the following form:

$$\begin{bmatrix} V_1^- \\ V_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} V_1^+ \\ V_2^+ \end{bmatrix}, \tag{1.4}$$

Where V_1^-, V_2^- are the reflected wave voltages, V_1^+, V_2^+ are the incident wave voltages on the ports, and S_{11}, S_{12}, S_{21}, S_{22} are the S-parameters that expressed using the complex numbers. The overall voltage at port 1 is $V_1 = V_1^+ + V_1^-$ and at port 2 is $V_2 = V_2^+ + V_2^-$. It should be mentioned that the DC power supply port is omitted in this kind of two-port representation. This means that S-parameters represent an amplifier for specific bias conditions. Morover, the S-parameters provide the network description for one frequency only.

The S-parameters expressed as a set of two equations take the following form:

$$V_1^- = S_{11}V_1^+ + S_{12}V_2^+, \tag{1.5}$$

$$V_2^- = S_{21}V_1^+ + S_{22}V_2^+. \tag{1.6}$$

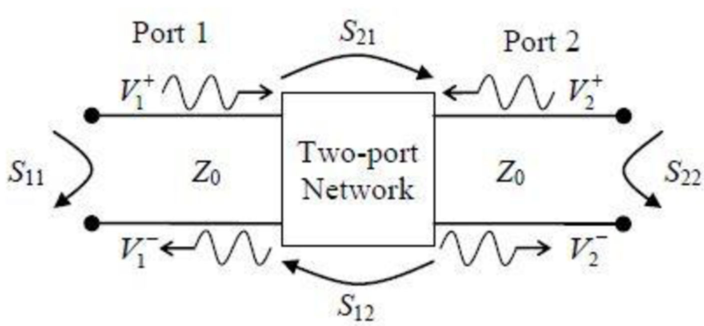


Figure 1.1: S-parameters in the two-port network.

The S_{11} parameter can be found from (1.5) as:

$$S_{11} = \left.\frac{V_1^-}{V_1^+}\right|_{V_2^+=0} \tag{1.7}$$

The condition $V_2^+ = 0$ in (1.7) means that there is not a wave coming from the port 2. In other words, we say that the port 2 is matched to a load. In majority RF systems, the termination impedance of the port two would have value 50Ω.

The remaining S-parameters of the two port network are found a similar way:

$$S_{21} = \left.\frac{V_2^-}{V_1^+}\right|_{V_2^+=0} \tag{1.8}$$

is transmission coefficient from port 1 to port 2,

$$S_{22} = \left.\frac{V_2^-}{V_2^+}\right|_{V_1^+=0} \tag{1.9}$$

is reflection coefficient at port 2,

$$S_{12} = \left.\frac{V_1^+}{V_2^+}\right|_{V_1^+=0} \tag{1.10}$$

is the transmission coefficient from port 2 to port 1.

Considering amplifiers, the parameters S_{21} is commonly expressed in the logatithmic scale and it relates to gain. Return loss (RL) which is commonly used as the measure of input and output matching of amplifiers can be derived, for example, for the input port from the parameter S_{21} as:

$$RL_{IN} = 20\log\left|\frac{1}{S_{11}}\right| \tag{1.11}$$

In other words, RL can be also understood as the decibel-difference between the incident power and reflected power.

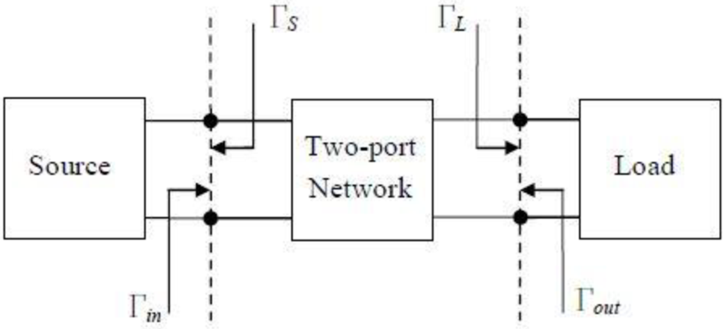


Figure 1.2: Terminated two-port network with source and load.

1.2.3 Stability

1.2.3.1 General Concept

We know consider a two-port network connected to a source and terminated with arbitrary impedances. The corresponding reference planes represented by reflection coefficients are shown in Figure 1.2. The two-port network is unconditionally stable if the magnitudes of the reflection coefficients Γ_{in}, and Γ_{out} are less than unity for passive loads [8], thus it should hold that

$$\text{for: } |\Gamma_S| < 1, |\Gamma_L| < 1, \tag{1.12}$$

$$|\Gamma_{in}| = \left|S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}\right| < 1, \tag{1.13}$$

$$|\Gamma_{out}| = \left|S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}\right| < 1, \tag{1.14}$$

From the equations 1.13 and 1.14 is apparent that Γ_S and Γ_L have an effect on the stability as the S-parameters remain constant for a particular frequency. Depending on the values of Γ_S and Γ_L the stability of the two-port network can be of two kinds, unconditionally stable or potentially unstable [7]. The unconditional stable two-port network will never oscillate for any passive loading connected to the network. The potentially unstable two-port network may oscillate if Γ_S and Γ_L reflection coefficients do not support the stability conditions (1.13) and (1.14).

1.2.3.2 Stability Factors

Source stability factors μ_S and load stability factor μ_L determinate the stability of a two-port network considering only the S-parameters.

$$\mu_S = \frac{1 - |S_{22}|^2}{|S_{22} - S_{11}^{*}\Delta| + |S_{21}S_{12}|}, \tag{1.15}$$

$$\mu_L = \frac{1 - |S_{11}|^2}{|S_{22} - S_{11}^{*}\Delta| + |S_{21}S_{12}|}, \tag{1.16}$$

$$\Delta = S_{11}S_{22} - S_{12}S_{21}. \tag{1.17}$$

Derivation of the stability factors is based on the conditions given by equations (1.13) and (1.14). It is enough to check only one of these two stability factors. Therefore the active two-port network is unconditionally stable if $\mu_S > 1$. In other words when $\mu_S > 1$ then also $\mu_L > 1$ and vice versa. In other case, when $\mu_S \leq 1$ the network is potentially unstable.

1.2.3.3 Stability Circles

The potentially unstable network will not oscillate if the Γ_S and Γ_L are selected from stable regions. The stable and unstable regions are separated by a stability circle. For example, the output stability circle represents all values of Γ_L when $|\Gamma_{IN}| = 1$. The center C_L and radius r_L of the circle (Figure 1.3) are given by

$$r_L = \left|\frac{(S_{12}.S_{21})^*}{|S_{22}|^2 - |S_{11}.S_{22} - S_{12}.S_{21} -|^2}\right|, \tag{1.18}$$

$$C_L = \frac{(S_{22} - (S_{11}.S_{22} - S_{12}.S_{21}).S_{11}{}^*)^*}{|S_{22}|^2 - |S_{11}.S_{22} - S_{12}.S_{21} -|^2}. \tag{1.19}$$

In addition μ_L defines the distance between the center of the Smith chart and the unstable region in the load plane (Figure 1.3).

Stable and unstable regions must be observed very carefully on the Smith chart especially if the output stability circle encloses the origin $\Gamma_L = 0$ and if the S-parameter $|S_{22}| > 11$ [7].

The potentially unstable two-port network can be made unconditional stable by a resistive loading. Although employing the resistive loading results in overall performance degradation it is commonly applied stabilization method for potentially unstable two-port network [7].

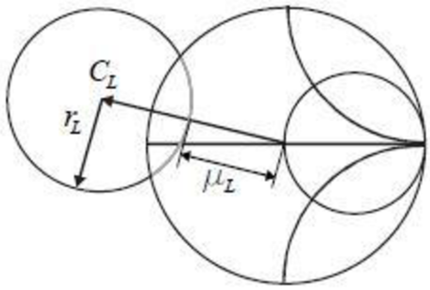


Figure 1.3: Load stability circle of a two-port plotted on the Smith chart Γ_L plane. C_L is a complex number representing the center and r_L is the radius of the circle. μ_L defines the distance between the center of the Smith chart and the unstable region.

1.2.4 Power Gain

If a two-port network is connected to a generator and to a load, four signal power levels can be identified as it is shown in Figure 1.4

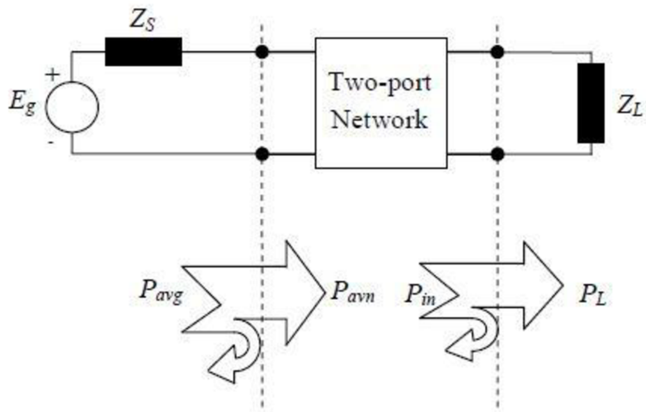


Figure 1.4: Power relations in a two-port network.

We denote these power levels at different positions of the circuit as available power from the source P_{avg}, available power from the two-port network P_{avn}, power input to the two-port network P_{in}, and power delivered to the load P_L. Knowing the power relations in every port, transducer power gain G_T, power gain G_P, and the available power gain G_A can be derived. The first mentioned gain G_T is dependent on both impedances Z_S and Z_L and has the following form

$$G_T = \frac{P_L}{P_{avg}} = \frac{(1-|\Gamma_L|^2)|S_{21}|^2(1-|\Gamma_S|^2)}{|(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_L)-S_{21}S_{12}\Gamma_L\Gamma_S|^2}, \qquad (1.20)$$

The simplified version of G_T can be obtained substituting either (1.13) or (1.14) which also helps to derive G_A and G_P. For example, if we put the Γ_L equal to the complex conjugate of the output reflection coefficient of the two-port network Γ_{out} and substitute it to the equation (1.20), we obtain the available power gain G_A. As a result, the G_A has the following form

$$G_A = \frac{P_{avn}}{P_{avg}} = \frac{|S_{21}|^2(1-|\Gamma_S|^2)}{(1-|\Gamma_{out}|^2)|1-S_{11}\Gamma_S|^2}. \qquad (1.21)$$

G_P is independent of Z_S and has the following form

$$G_P = \frac{P_L}{P_{in}} = \frac{(1-|\Gamma_L|^2)|S_{21}|^2}{(1-|\Gamma_{in}|^2)|1-S_{22}\Gamma_L|^2}. \qquad (1.22)$$

G_A and G_P can be used to derive the so-called gain circles which are drawn on the Smith chart. For example, in the low noise amplifier design, it is common to locate the avaible gain circles together with the noise figure circles.

In addition, gain relations as the Maximum Avaible Gain (MAG) and the Maximum Stable Gain (MSG) are often used in transistor data sheets for the description of maximum operating capabilities [8]. The first mentioned gain MAG is defined only when a transistor is unconditionally stable. The second mentioned gain MSG presents the highest theorically achievable gain if the device is terminated by passive terminations and stabilized at the verge of instability such that stability factor μ_S or $\mu_L = 1$, then

$$MSG = \frac{S_{21}}{S_{12}}. \qquad (1.23)$$

This gain, however, can not be achieved, because the required source and load terminations are situated on the circumference of the Smith chart [8].

1.2.5 Impedance Matching Networks

Impedance matching is a method which transforms given impedance into desired impedance in order to achieve a particular performance. The performance can be the maximum power transfer or minimum noise figure. The first mentioned performance aims to achieve maximum power transfer, thereby diminish losses caused by impedance mismatch. The second mentioned performance aims to minimize the noise figure by matching and optimum source impedance to the impedance to the generator impedance is called noise matching [2].

When designing a matching network, one should consider the following attributes [5]:

- **Complexity** - a simple topology composed of a minimum number of passive components is preferred, because it is cheaper and introduces less losses.
- **Bandwidth** - any type of matching network can be designed for a perfect match at single frequency only. Systems employing wide frequency operating range bring more complexity into the matching networks, because homogeneous conditions for every frequency component are often desired.
- **Implementation** - depending on a printed circuit board (PCB) area available and on feasibility, a matching network can be composed of stubs, lumped components, or their combinations. In addition, quarter wave transformers might be also used for impedance matching.
- **Adjustment** - adjustment components provide possibilities to optimize matching networks for different types of loading once a circuit is fabricated. For example, adjustable capacitors can be used in series or in parallel to a load.

1.2.6 Bias Circuits

Maintaining a stable operating point (Q-point) over temperature fluctuations is one of the elementary functions of the bias circuits. Biasing is a process of setting up the bias point. Usually this point is at the middle of the DC load line applying drain voltage and current [9]. In a field-effect transistor (FET), bias is the DC voltage supplied from a power supply which is applied at the drain. According to the selected Q-point, the biasing circuit is designed to operate the transistor at that Q-point. [9].

1.2.6.1 Passive DC Bias Circuit

It is a type of circuit based in passive components such as resistors, capacitors and inductors. It is not very used because the initial Q-point chosen usually changes due to the temperature sensitivity, which is typical in this kind of circuit.

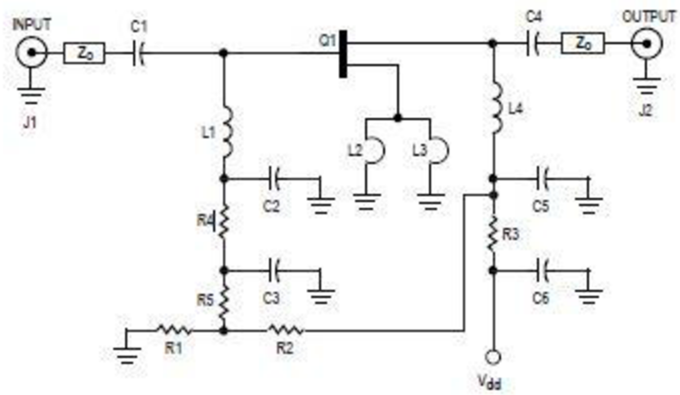


Figure 1.5: Typical ATF-54143 transistor with passive biasing. [9]

1.2.6.2 Active DC Bias Circuit

This type of circuit is implemented with active components such as diode or transistor as well as the passive components already mentioned before. A low frequency transistor is usually used as a stable current source, it means that the current will be constant and consequently the Q point as well. On the other hand, this type of circuit is generally used because is not affected by influence of the temperature or other factors. For designing of LNA is used the following kind of bias circuit

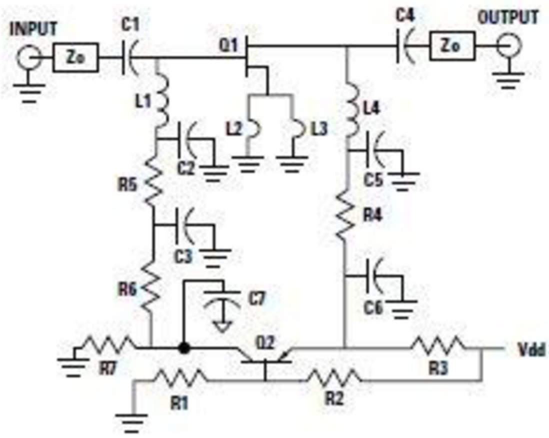


Figure 1.6: Typical ATF-54143 transistor with active biasing. [9]

1.3 Band Pass Filter

Band Pass Filter (BPF) and Low Noise Amplifier have to be designed individually. Once these two blocks are designed, integration of these two blocks make a single module named BPF-LNA.

The figure 1.7 shows the complete block of LNA with BPF, where these are connected through a matching network.

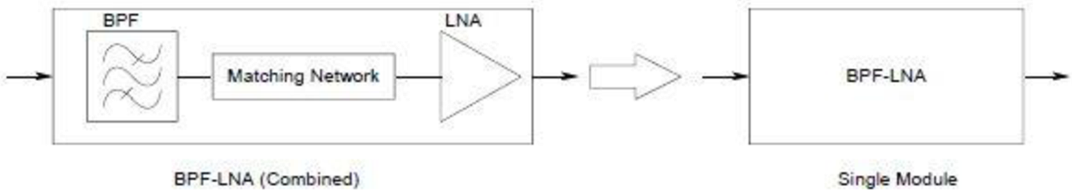


Figure 1.7: Complete LNA block diagram with BPF.

In RF transmitter and receiver, filters are key components which are used to pass or to reject signals based on frequency. Generally, there are four types of filters such as low-pass, high-pass, band-pass and band-stop filter. Combination of high-pass filter and low-pass filter make a band-pass filter (BPF) which is used to reject unwanted frequency bands and to pass a narrow pass-band [2].

There are several classes of band-pass filters such as Butterworth, Chebyshev and elliptical BPF. BPF can be designed in some ways like using lumped components or distributed components. In this thesis, a quarter-wave (λ/4-wave) coaxial resonator will be used as BPF, which will be discussed in detail in the subsequent section.

1.3.1 Quarter-wave (λ/4-wave) Coaxial Resonator

Quarter-wave (λ/4-wave) coaxial resonator is constructed by shorting the center conductor of a coaxial cable to the shield at the far end of the circuit. The length of the cable is exactly λ/4 at the desired resonant frequency [2].

A short circuit is transformed to an open circuit a quarter wavelength away, so when the λ/4-wave coaxial resonator is part of an oscillator circuit, it is not electrically even present (high impedance); however, whenever the frequency of the oscillator attempts to go above or below the resonator's center frequency (due to load changes, temperature changes, etc.), the λ/4-wave section looks like a low impedance that works to attenuate other frequency components. The feature of a λ/4-wave coaxial resonator is its high quality factor, "Q."

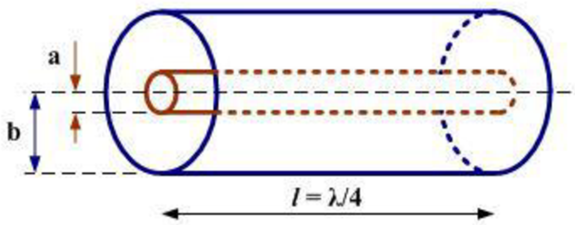


Figure 1.8: Dimension reference for λ/4-wave coaxial resonator.

a: Dimension of the inner conductor
b: Dimension of the outer conductor
l: Length of the quarter-wave (λ/4-wave) coaxial resonator

Coaxial resonator is often made from sections of normal coaxial cable (both flexible and semi-rigid). Quality factor of a λ/4-wave coaxial resonator is determined by the ratio of the center frequency (f_0) to the 3 dB power bandwidth, as shown in the figure below.

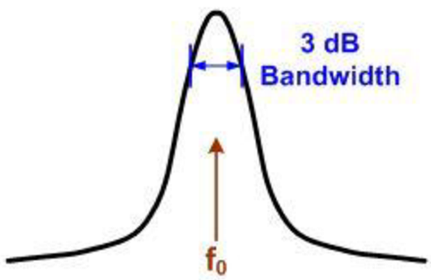


Figure 1.9: Q-factor quality of a λ/4-wave coaxial resonator.

2 PRACTICAL PART

2.1 Design and Simulation of LNA

2.1.1 General Considerations for LNA Design

Before starting to realize the design of a LNA, it is necessary to define objectives as gain, compression point, noise and others. It is important to define the purpose of the LNA, which will be reducing noise in the reception system. Moreover, it is necessary to determine the class of the amplifier such as A, B, AB, C, D, E or F.

Once these factors are determined, the next step is to choose an appropriate transistor, to propose a DC operating point, to study the stability of transistor and to calculate the values of biasing circuit. At this point it's convenient to start simulating the circuit, in order to prove the behavior of the amplifier. In the case of LNA, the simulation will be made by Ansoft Designer program.

Afterwards, the following step is to perform matching networks for input and output of the amplifier. These networks could be improved through the Ansoft Designer program in order to improve the objectives of the amplifier, as for example the gain, noise or pass-band.

2.1.2 Class A Amplifier

The operating point of the class A amplifier is determined between saturation region and cut-off region. It means that class A amplifier will work in the active region. The operating point is usually set in the middle of the active region (linear region). It is important to mention that class A amplifier is used in low-noise amplifiers because it has a linear behavior and low signal distortion. [10]

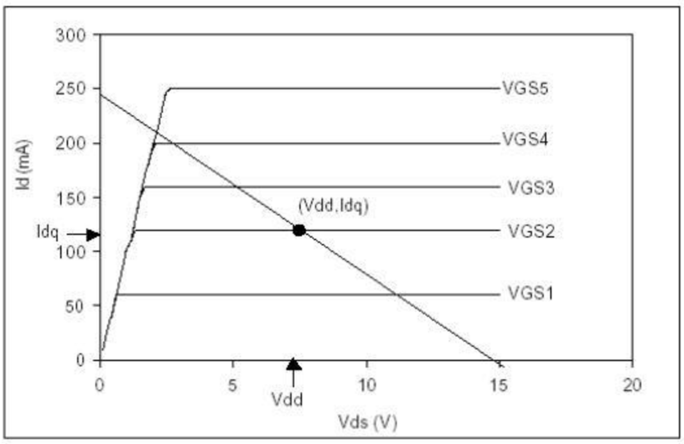


Figure 2.1: JFET characteristic curve - Load line.

Other features of A class amplifier are the signals of voltage and current in the output of the amplifier, which have the same waveforms as in the input. It means that the amplifier conducts over the entire range of the input cycle.

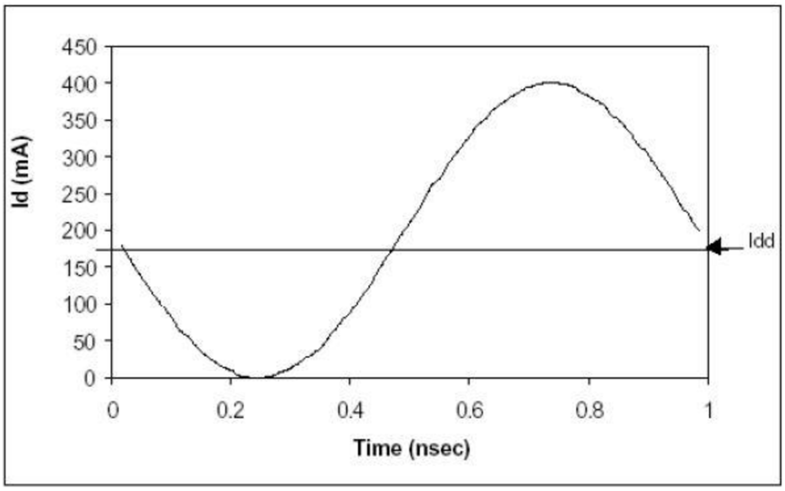


Figure 2.2: Wavefom of a class A amplifier.

2.1.3 Choice of the ATF-54143 Transistor

Once the main features and objectives of the LNA are known, it is possible to choose the appropriate transistor.

While choosing the appropriate transistor, there are some factors to take into account. First, it must operate in the frequency range of the LNA design, which is between 430-440 MHz of radio communication frequencies (RF). Second, the transistor technology should be appropriate for the LNA in order to obtain a low noise factor.

One of the most used transistor technologies for LNA designing is currently the pHEMT (pseudomorphic High Electron Mobility Transistor). Due to its construction, this technology has good performance in the electromagnetic spectrum of the RF; such as high linearity, low noise figure, gain and power.

According to this, the ATF-54143 transistor, made by Avago Technologies, was chosen for designing the LNA. In addition, this transistor has some favorable characteristics such as high compression point, adequate stability, high linear dynamic range and low cost factor important in modern times.

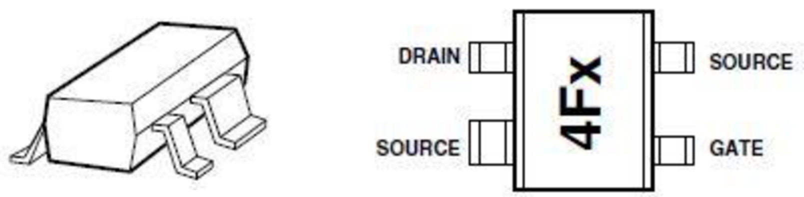


Figure 2.3: Package SOT-343 and pin connections of the ATF-54143 transistor. [9]

Table 2.1: Scattering parameters of the ATF-54143 transistor for $V_{DS} = 3\,V$ and $I_{DS} = 60$ mA at the frequency of 0.5 GHz. [9]

Freq.	Fmin.	Γopt	Γopt	Rn/50	Ga
GHz	dB	Mag.	Ang.	-	dB
0.5	0.15	0.34	42.3	0.04	28.50

Table 2.2: Typical noise parameters of the ATF-54143 transistor for $V_{DS} = 3\,V$ and $I_{DS} = 60$mA at the frequency of 0.5 GHz. [9]

Freq.	S11			S21		S12		S22		MSG/MAG
GHz	Mag.	Ang.	dB	Mag.	Ang.	Mag.	Ang.	Mag.	Ang.	dB
0.5	0.81	-80.8	26.04	20.05	128	0.03	52.4	0.4	-58.8	28.25

2.1.4 Choice of an Operating Point

The next step is to define the voltage V_{DS}, V_{GS} and the current I_{DS} of the ATF-54143 transistor in order to obtain the desired gain between 15-20 dB for the LNA designing.

It is possible to obtain an approximate gain of the transistor using the Scattering parameters, which are provided by Avago Technologies.

In the table 2.3 are shown mentioned parameters for different values of the V_{DS} and I_{DS} at the frequency of 0.5 GHz. These parameters will be replaced after in the equation (1.20) of the transducer power gain G_T.

Reflection coefficient to source Γ_S and reflection coefficient to load Γ_L are shown in the equation (2.1) and (2.2).

$$\Gamma_S = \frac{Z_S - Z_0}{Z_S + Z_0}, \tag{2.1}$$

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}. \tag{2.2}$$

Since given $Z_L = Z_S = Z_0 = 50\ \Omega$, the Γ_L and Γ_S are calculated as zero. Reflection coefficient at the input and output is defined by

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = S_{11}. \tag{2.3}$$

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S} = S_{22}. \tag{2.4}$$

For example for $V_{DS} = 3\ \mathrm{V}$ and $I_{DS} = 60\ \mathrm{mA}$ shown in table (2.3) the calculation of its transducer power gain G_T will be

$$G_T = \frac{P_L}{P_{avg}} = \frac{(1-|\Gamma_L|^2)|S_{21}|^2(1-|\Gamma_S|^2)}{|(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_L)-S_{21}S_{12}\Gamma_L\Gamma_S|^2} = |S_{21}|^2 = 402.$$

The gain in decibel units will be

$$G_{T(dB)} = 10 \cdot \log(G_T) = 10 \cdot \log(402) = 26.04\ \mathrm{dB}. \qquad (2.5)$$

Table 2.3: Scattering parameters of the ATF-54143 transistor for different values of the V_{DS} and I_{DS} at the frequency of 0.5 GHz. [9]

	$V_{DS} = 3\ \mathrm{V}$	$V_{DS} = 3\ \mathrm{V}$	$V_{DS} = 3\ \mathrm{V}$	$V_{DS} = 4\ \mathrm{V}$
Freq. = 0.5 GHz	$I_{DS} = 40\ \mathrm{mA}$	$I_{DS} = 60\ \mathrm{mA}$	$I_{DS} = 80\ \mathrm{mA}$	$I_{DS} = 60\ \mathrm{mA}$
$\lvert S_{11}\rvert$	0.83	0.81	0.80	0.81
$\lvert S_{21}\rvert$	18.77	20.05	18.45	20.22
$\lvert S_{12}\rvert$	0.036	0.03	0.04	0.03
$\lvert S_{22}\rvert$	0.44	0.4	0.29	0.42
G_T[dB]	25.46	**26.04**	25.32	26.11

In addition, it is important to know that there is a particular case of unilateral transducer power gain G_{Tu} for the condition of the parameter $S_{12} = 0$, which is defined by

$$G_{Tu} = \frac{|S_{21}|^2(1-|\Gamma_S|^2)(1-|\Gamma_L|^2)}{|1-S_{11}\Gamma_S|^2|1-S_{22}\Gamma_L|^2}. \qquad (2.6)$$

And maximum unilateral transducer power gain $G_{Tu(max.)}$ for the $\Gamma_S = S_{11^*}$ and $\Gamma_L = S_{22^*}$

$$G_{Tu(max.)} = \frac{|S_{21}|^2}{(1-|S_{11}|^2)(1-|S_{22}|^2)}. \qquad (2.7)$$

After obtaining the approximate gain of the transistor by analysis of Scattering parameters, it was observed that the gain for the chosen example with values of $V_{DS} = 3\ \mathrm{V}$ and $I_{DS} = 60\ \mathrm{mA}$ satisfies the gain condition for the LNA designing.

Moreover, for designing the LNA it is advisable to set the operating Q-point in the middle of the active zone of the transistor in order to obtain an optimal range of amplification between zones of the cut-off and saturation. Figure 2.4 shows the load line of the ATF-54143 transistor, the operating Q-point in the middle of the active zone and the obtained values for $V_{DS} = 3$ V, $I_{DS} = 60$ mA and $V_{GS} = 0.59$ V.

Therefore, the values chosen for the operating Q-point satisfy for the appropiate designing of this LNA.

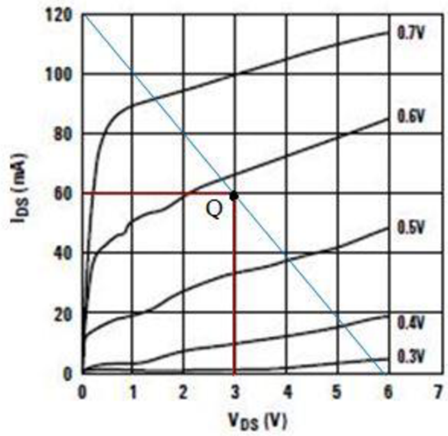


Figure 2.4: Load line of the ATF-54143 transistor for the $V_{DS} = 3$ V, $I_{DS} = 60$ mA and $V_{GS} = 0.59$ V . [9]

2.1.5 Transistor Stability

This section focuses on study of stability of the ATF-54143 transistor. For this purpose of improving the transistor stability is used the Ansoft Designer Software Version SV2, which simulates the properties of this transistor as a two-port network. Therefore, different graphs and schemes are observed such as the Scattering parameters, the Rollett's stability factor (K) and stability circles on the Smith chart for the respective analysis of this transistor.

Before starting the simulation, it is necessary to download the Scattering parameters of this transistor, which are provided by Avago Technologies for different operating points and frequencies. For this design were chosen $V_{DS} = 3$ V, $I_{DS} = 60$ mA at a frequency of 0.5 GHz.

Figure 2.5 shows the basic schema of the ATF-54143 transistor stability in Ansoft Designer, which is analyzed independently. The next figure shows the Scattering parameters such as S_{11}, S_{12}, S_{21} and S_{22} at frequency range from 0.1 GHz to 18 GHz in order to verify the behavior of the transistor at this frequency range.

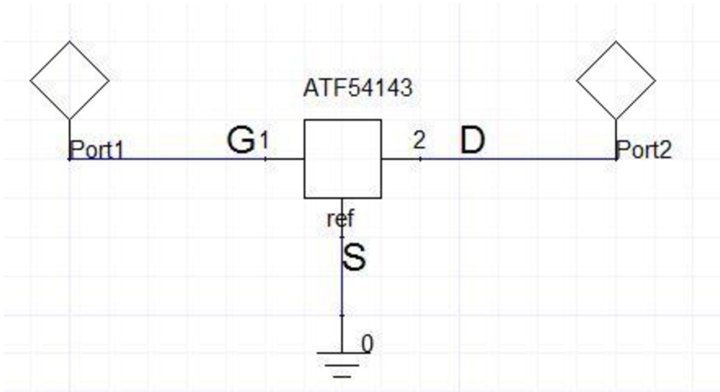


Figure 2.5: Basic scheme of the ATF-54143 transistor stability in Ansoft Designer.

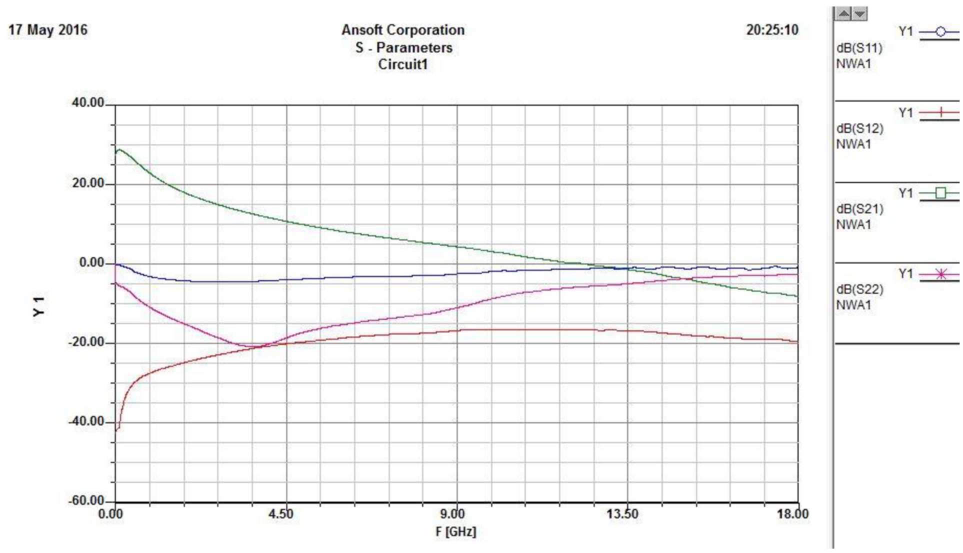


Figure 2.6: S-parameters of the ATF-54143 transistor at the frequency range from 0.1 GHz to 18 GHz.

In Figure 2.7 is shown the Rollett's stability factor (K), which indicates the stability of the transistor for values greater than 1. In this case, it has been observed that transistor is stable at frequency values greater than 3.61 GHz. For frequency values less than 3.61 GHz it is unstable. It is necessary to improve the stability of the transistor at frequency range from 430 MHz to 440 MHz, which is required for the LNA design.

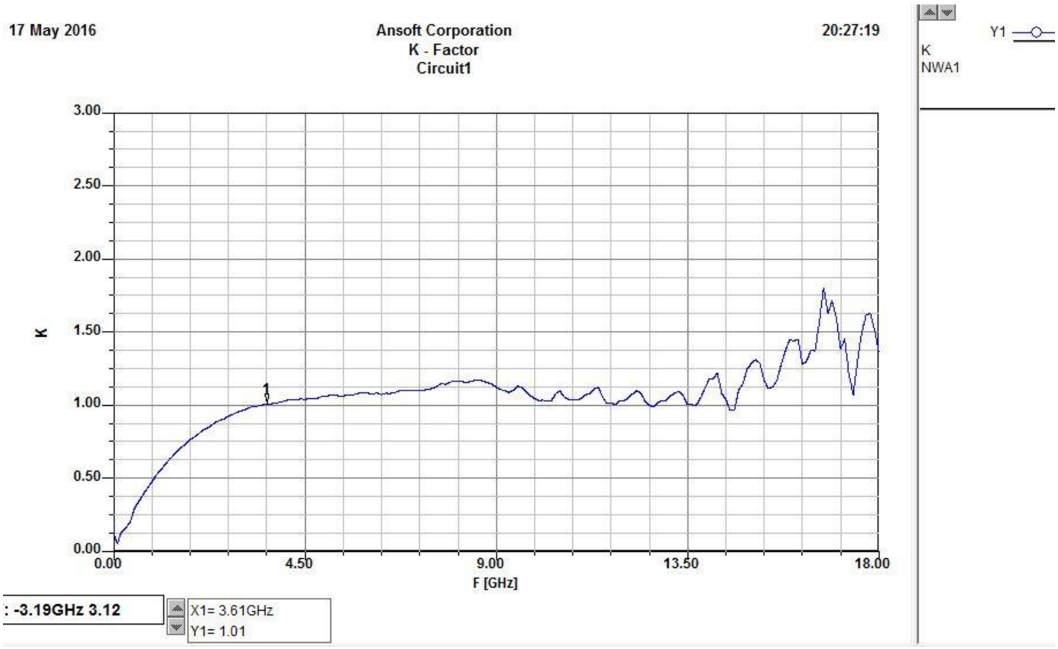


Figure 2.7: Rollett's stability factor (K).

Other way to verify the transistor stability is using the K-factor Circle for Source (KCS) and K-factor Circle for Load (KCL) on the Smith chart, which indicates that there is stability when both circles KCS and KCL are out of the Smith chart. As Figure 2.8 shows, the transistor is unstable at the frequency range from 430 MHz to 440 MHz because both circles KCS and KCL are inside of the Smith chart.

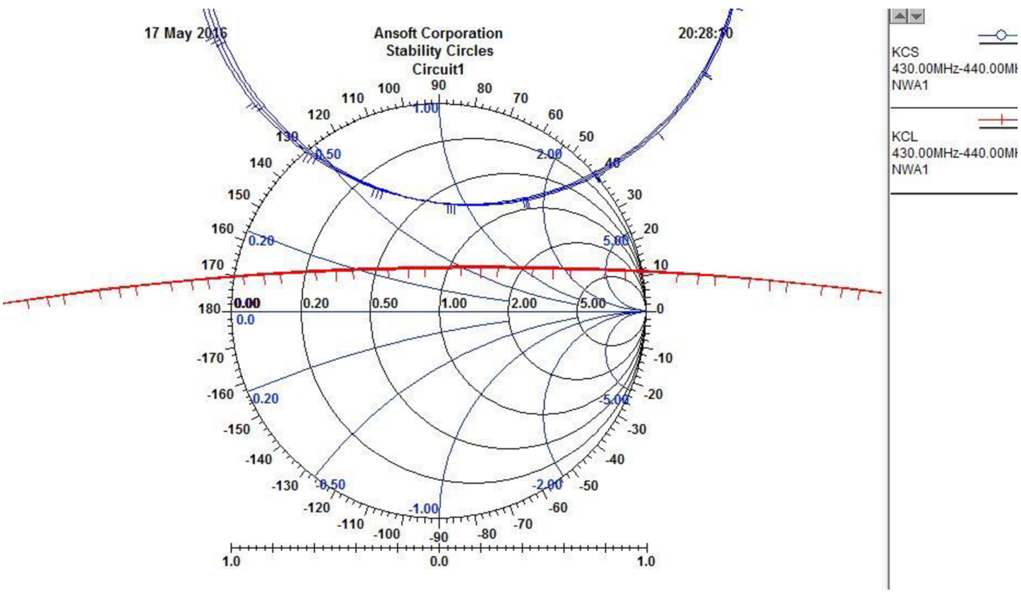


Figure 2.8: KCS and KCL stability circles on the Smith chart.

In order to improve the stability of the transistor, a resistor in series of low value was connected to the drain of the transistor. According to the simulation results, the resistor value of 27 Ω has been chosen because this value does not change the behavior of Scattering parameters at the given frequency range.

Morover, very short transmission lines between each source lead and ground have been used, in this manner the stability has improved without much degeneration of the noise properties. These transmission lines now serve as “negative current feedback” and improve the stability. It has been simulated with the following parameters: length $P = 1$ mm and width $W = 0.3$ mm as shows Figure 2.9.

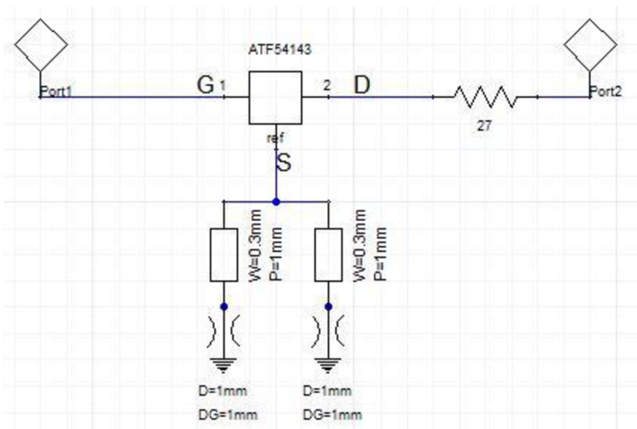


Figure 2.9: Improving the stability of the ATF-54143 transistor.

In figure 2.10 is shown the improving stability of the ATF-54143 transistor, but at frequencies below 0.8GHz the transistor is still unstable.

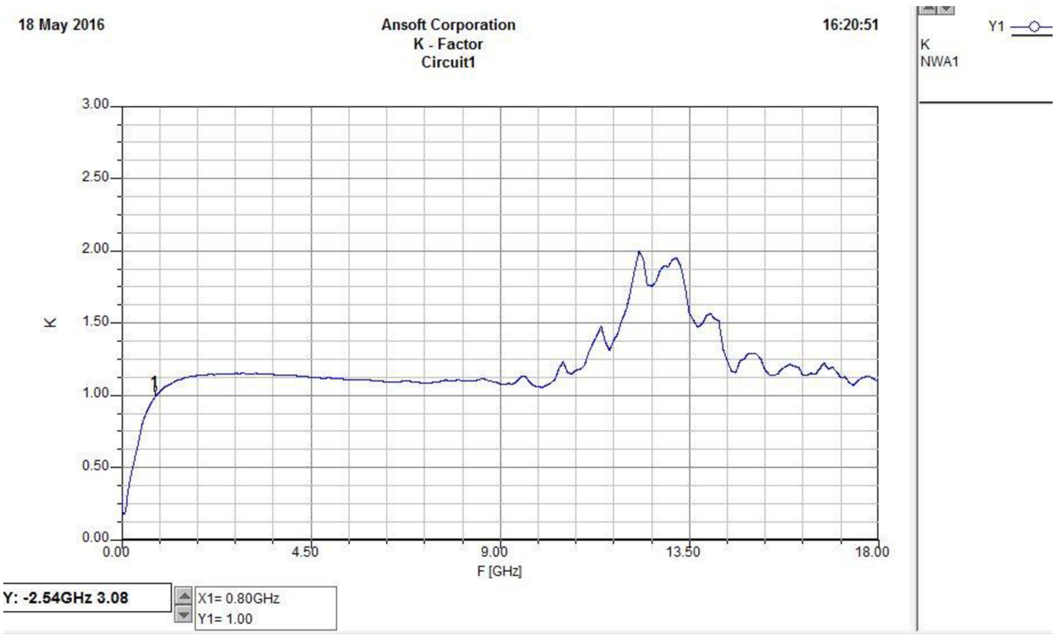


Figure 2.10: Improving stability - Rollett's stability factor (K).

Figure 2.11 shows the connection of biasing networks of ATF-54143 transistor recommended by Avago Technologies. These networks also influence the stability of the transistor. The values used for components were provided by Avago Technology and in some cases they were changed in order to improve the performance for the LNA designing.

The inductor L1 provides low frequency gain reduction, which can minimize the transistor's susceptibility to overload from nearby low frequency transmitters. L1 provides a means of inserting gate voltage for biasing of the transistor. This requires a good bypass capacitor in the form of C7. This network represents a compromise between noise figure, input return loss and gain.

Resistor R6 and capacitor C5 provide a low frequency resistive termination for the device which helps to keep the stability. The same applies to the resistor R7 and capacitor C6. The inductor (L2) provides a means of inserting drain voltage for biasing of the transistor and this also requires a good bypass capacitor in the form of C8.

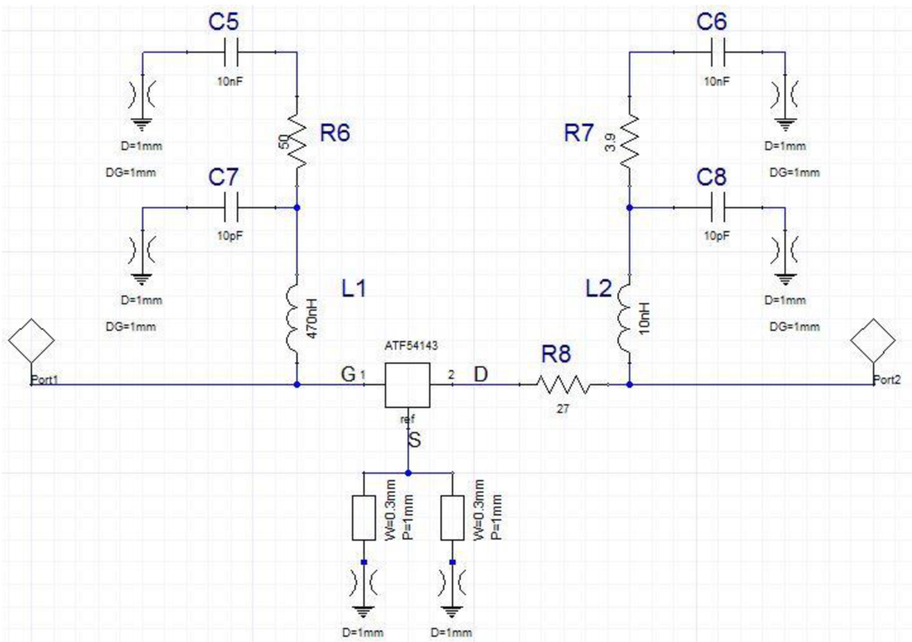


Figure 2.11: Complete scheme of the ATF-54143 transistor's stability.

In Figure 2.12 can be observed the complete stability of the transistor by Rollet stability factor (K) at the shown frequency range where the centre frequency 435 MHz is included.

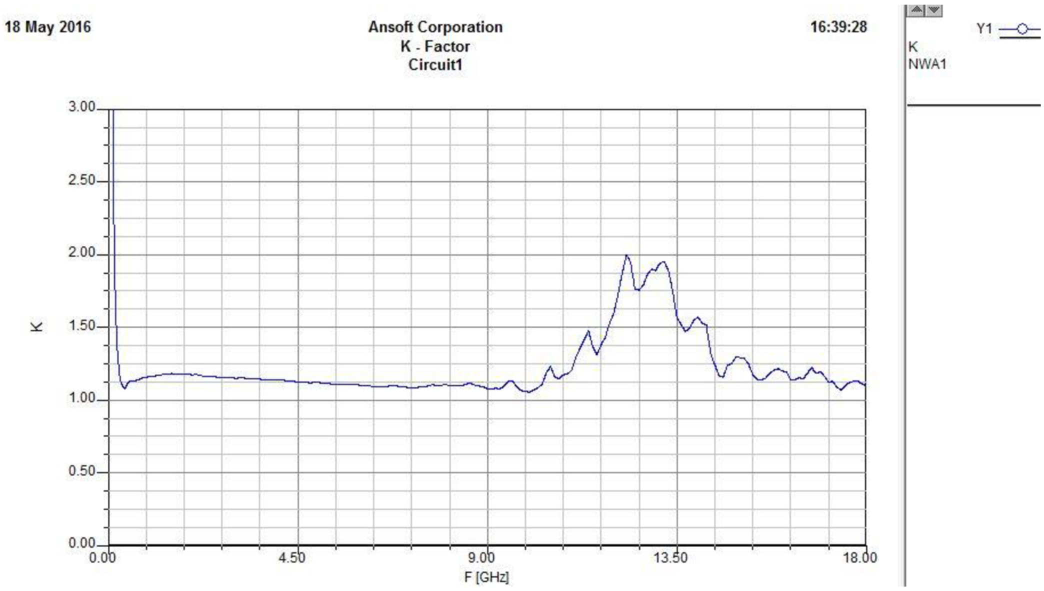


Figure 2.12: Stable transistor - Rollett's stability factor (K).

It is also possible to observe the transistor stability by the stability circles on the Smith chart as it is shown in Figure 2.13, in this case at frequency range from 430 MHz to 440 MHz. The KCS and KCL stability circles are outside of the Smith chart, which are the conditions for the stabilily of the transistor.

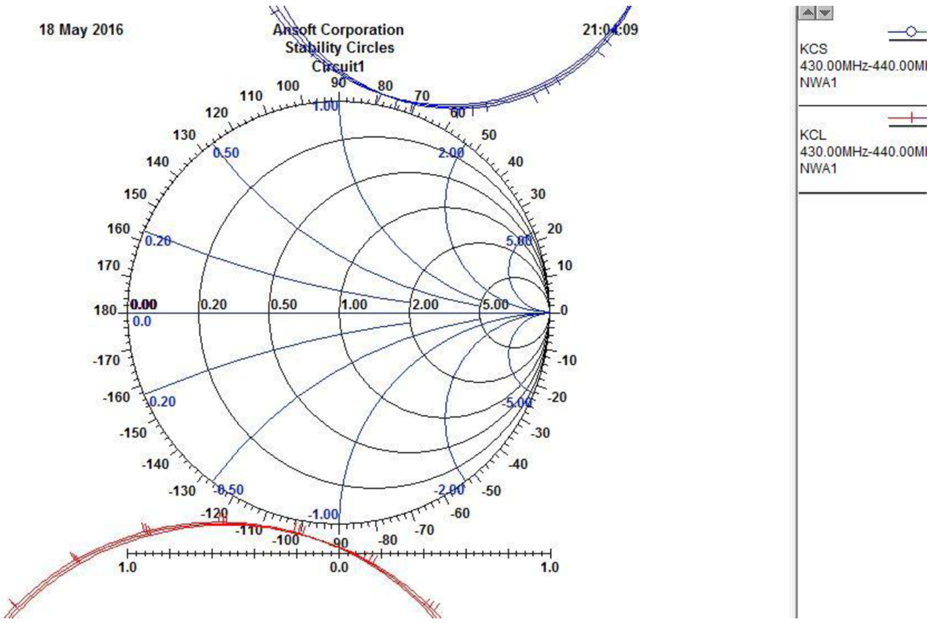


Figure 2.13: Stable transistor - KCS and KCL stability circles on the Smith chart

Other way to find out if the transistor is stable is to verify if Scattering parameters, such as input reflection coefficient (S11) and output reflection coefficient (S22), are below 0 dB at all frequency range as it is shown in Figure 2.14. A different value of these parameters could generate that the transistor behaves as an oscillator causing its inestability.

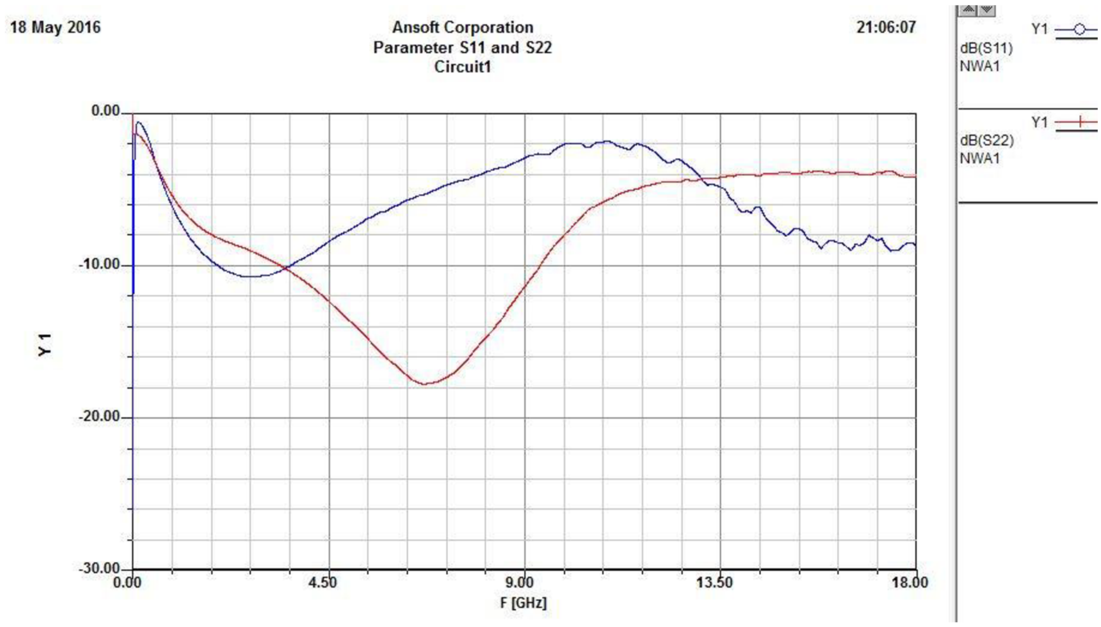


Figure 2.14: Parameters S11 and S22 at the frequency range from 0.1 GHz to 18 GHz.

2.1.6 Calculating Values of the Active DC Bias Circuit

As it was mentioned in chapter 1.2.6.2, an active DC bias circuit is used for setting the operating point (Q-point). This circuit has been chosen because it keeps stability of the Q-point. It means that the circuit is not affected by any influence of the temperature or other factors.

In Figure 2.15 is shown the typical active DC bias circuit for the ATF-54143 transistor recommended by Avago Technologies. BC807 general purpose transistor was chosen for this active DC bias circuit because it fulfils with requirements for this circuit. Equation relationships used to calculate the component values are also provided by Avago Technologies in the catalog of the ATF-54143 transistor.

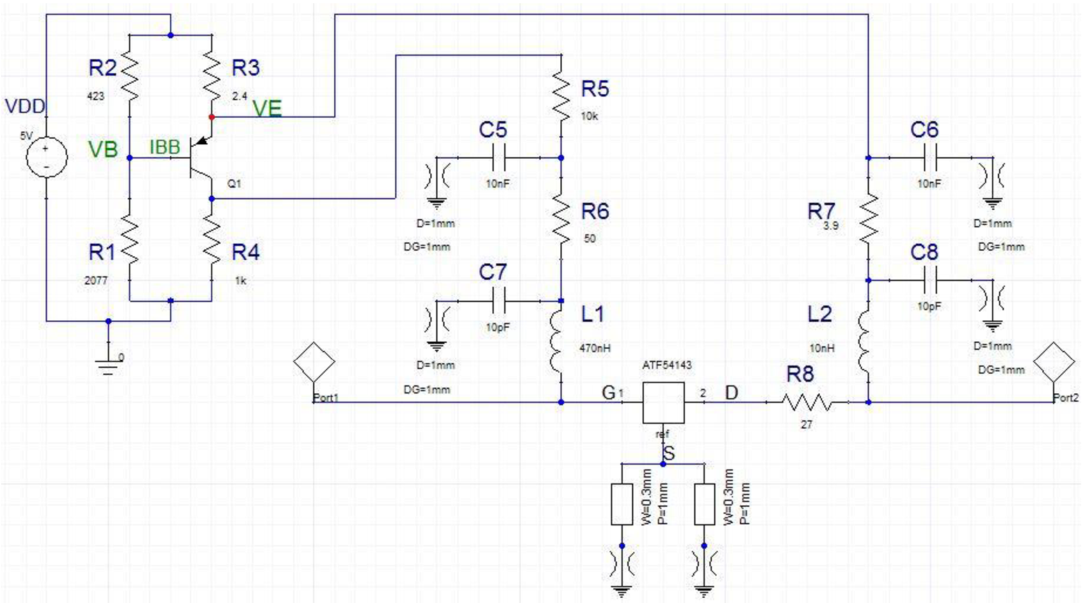


Figure 2.15: Active DC bias circuit for LNA

Before calculating component values of the active bias circuit it is necessary to define parameters, which the circuit should have, such as DC voltage source and the operating point of the ATF-54143 transistor.

In chapter 2.1.4 was chosen the operating point for the ATF-54143 transistor. Therefore, the values considered are:

- $V_{DD} = 5$ V - DC voltage source of the ATF-54143 transistor.
- $V_{DS} = 3$ V - Drain-to-source voltage of the ATF-54143 transistor.
- $I_{DS} = 60$ mA - Drain-to-source current of the ATF-54143 transistor.
- $V_{GS} = 0.59$ V - Gate-to-source voltage of the ATF-54143 transistor.
- $V_{BE} = 0.7$ V - Base-to-emitter voltage of the BC807 transistor.
- $R_7 = 3.9\ \Omega$ - Resistor value used for the stability of ATF-54143 transistor
- $R_8 = 27\ \Omega$ - Resistor value used for the stability of ATF-54143 transistor
- $I_{BB} = 2$ mA - Current flowing through R_1/R_2 resistor voltage divider network (approximately).

The equations that describe the operation of this active DC bias circuit are as follows [9]:

For calculating R_3

$$V_E = V_{DS} + (I_{DS} \cdot (R_7 + R_8)). \tag{2.8}$$

$$3 + \left(60 \cdot 10^{-3} \cdot (3.9 + 27)\right) = 4.854\ \text{V}.$$

Afterwards:

$$R_3 = \frac{V_{DD} - V_E}{I_{DS}}. \tag{2.9}$$

$$\frac{5 + 4.854}{60 \cdot 10^{-3}} = 2.4\Omega$$

For calculating R_1andR_2:

$$V_B = V_E - V_{BE} \tag{2.10}$$

$$4.854 - 0.7 = 4.154\,V.$$

$$V_B = \frac{R_1}{R_1 + R_2} \cdot V_{DD} \tag{2.11}$$

$$V_{DD} = I_{BB} \cdot (R_1 + R_2) \tag{2.12}$$

Rearranging equation (2.12) provides the following formula:

$$R_1 = \frac{V_{DD}}{I_{BB} \cdot \left(1 + \frac{V_{DD} - V_B}{V_B}\right)}. \tag{2.13}$$

$$\frac{5}{2 \cdot 10^{-3} \cdot \left(1 + \frac{5 - 4.154}{4.154}\right)} = 2077\ \Omega$$

And rearranging equation (2.11) provides the following formula:

$$R_2 = \frac{R_1 \cdot (V_{DD} - V_B)}{V_B}. \tag{2.14}$$

$$\frac{2077 \cdot (5 - 4.154)}{4.154} = 423\ \Omega$$

Resistor R4 is chosen to be 1 KΩ. This resistor keeps a small amount of current flowing through of the operating point of the BC807 transistor to help maintain bias stability. Resistor R5 is chosen to be 10 KΩ. This value of resistance is necessary to limit gate current of the ATF-54143 transistor in the presence of high signal of radio frequency (RF).

2.1.7 Design of Matching Networks

As it was mentioned in the chapter 1.2.5, the impedance matching is a method, which transforms given impedance into desired impedance in order to achieve a particular performance. For obtaining matching networks for the input and output of the LNA is used the tool that provides Ansoft Designer as known as “Smith Tool“, which is necessary to configure in advance.

To achieve the input impedance matching with impedance of 50 Ω was necessary to set the LNA center frecuency of operation at 435MHz. After that, it was selected the noise parameter with reference value of Minimum noise figure (Fmin) equal to 0.13dB.

The next step was to select the Available Gain Circle for Source (GACS) with reference value minor than the maximum gain, the reference value selected was equal to 25.64dB. In this way, the circles of noise and GACS were obtained on the Smith chart. The value of GACS was set to in order to obtain an intersection between the circle of GACS and the center of the circle of the noise, the value set was 22.89 dB.

Subsequently, it was possible to choose the suitable components for the input impedance matching, taking into account the center point of the Smith chart and the intersection between the circle of GACS and the center point of the circle of noise as the beginning and final of the selection of components respectively. The result of the steps mentioned can be observed on the Smith chart in Figure 2.16.

The input impedance matching obtained was an inductor equal to 13.69 nH connected in series to microstrip line with lenght P=27.64 mm and width W=2.89 mm. The conection of input impedance matching network is shown in Figure 2.18.

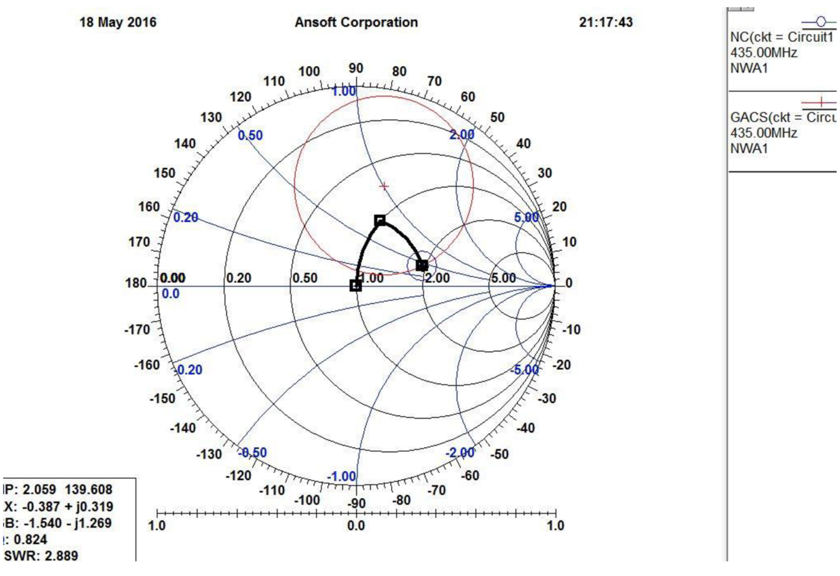


Figure 2.16: Smith Tool - Input impedance matching

To achieve the output impedance matching with impedance of 50 Ω was executed the same steps as for the input impedance matching but adding a new configuration for obtaining the Available Gain Circle for Load (GACL), which had as reference value the same value of the GACS equal to 22.89 dB.

After the obtaining of the GACL, it was selected the intersection between the GACS circle and the center point of the circle of noise in order to obtain a reference point into the GACL circle. Finally, after these steps, it was possible to obtain the conjugate of the previous point obtained from the GACL circle.

This new point was the initial reference for the suitable components choosing for the output impedance matching and the middle of the Smith chart was the final reference. The result of the steps mentioned can be observed on the Smith chart in Figure 2.17.

The output impedance matching obtained was a capacitor equal to 4.12 pF connected in series to microstrip line with lenght P=39.41 mm and width W=2.89 mm. The conection of output impedance matching network is shown in the figure 2.18.

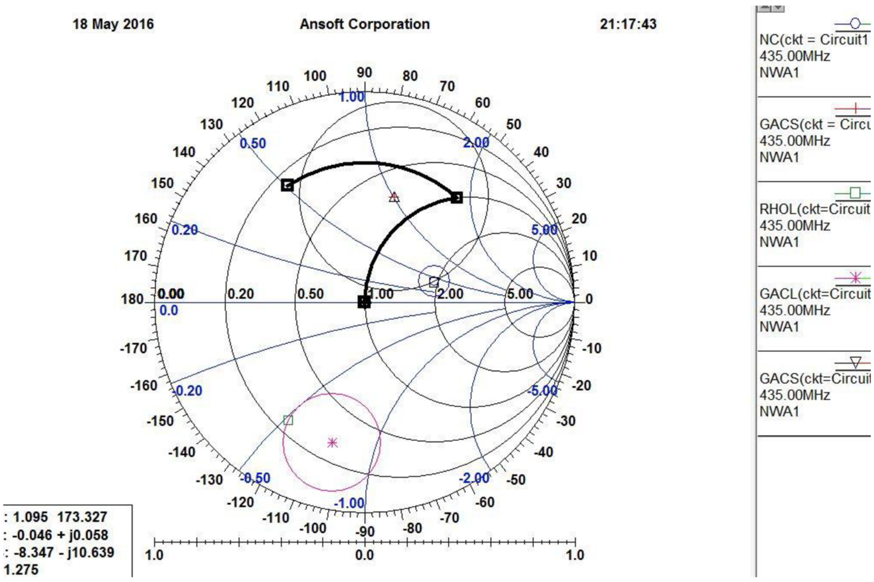


Figure 2.17: Smith Tool - Output impedance matching

In Figure 2.18 is shown the schema of LNA with impedance matching networks obtained for the input and output. The dimensions of microstrip lines were obtained by tool “Microstrip single“ of the Ansoft Designer, in which it was necessary to set the value of characteristic impedance, the electrical length in degrees, the center frequency and the parameters of substrate. The substrate used for the LNA designing is FR-4. The dimensions of microstrip lines obtained are shown in the schema below. For implementation purposes on the PCB due to space limitations, which were not suitable, it was decided to use an inductor 26.5 nH equivalent to the microstrip line of the output impedance matching (See figure 2.23).

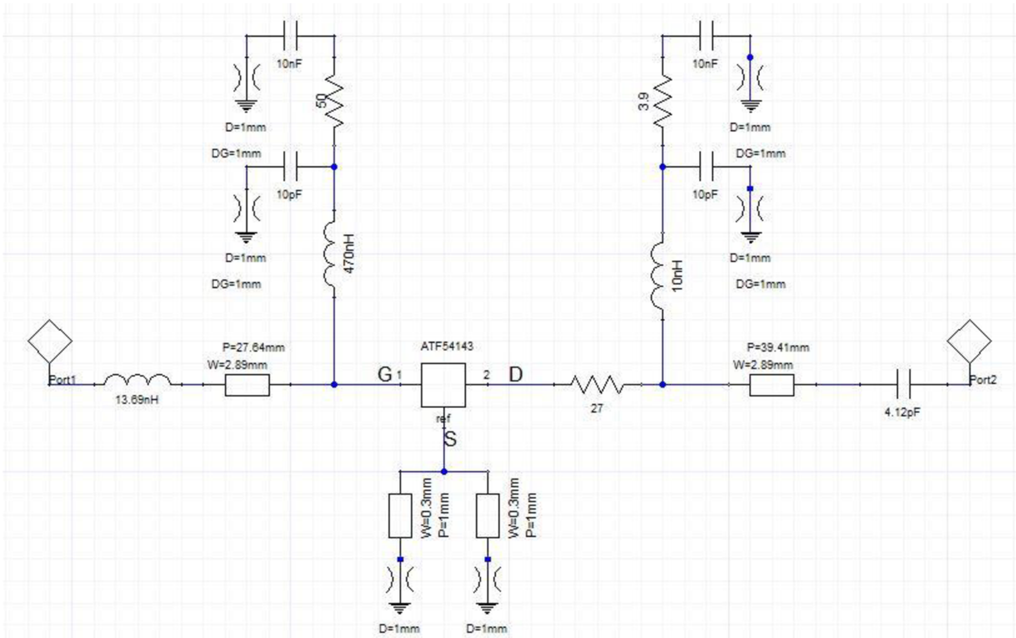


Figure 2.18: Scheme of the LNA with impedance matching networks

The figure 2.19 shows the value of the noise figure (NF) and the value of the minimun noise figure (Fmin) for the LNA at center frequency of 435MHz after set the impedance matching networks. As the graph shows, the value of the noise figure is equal to the value of the mininum noise figure: NF = Fmin = 0.18dB, which indicates that the impedance matching networks were properly designed.

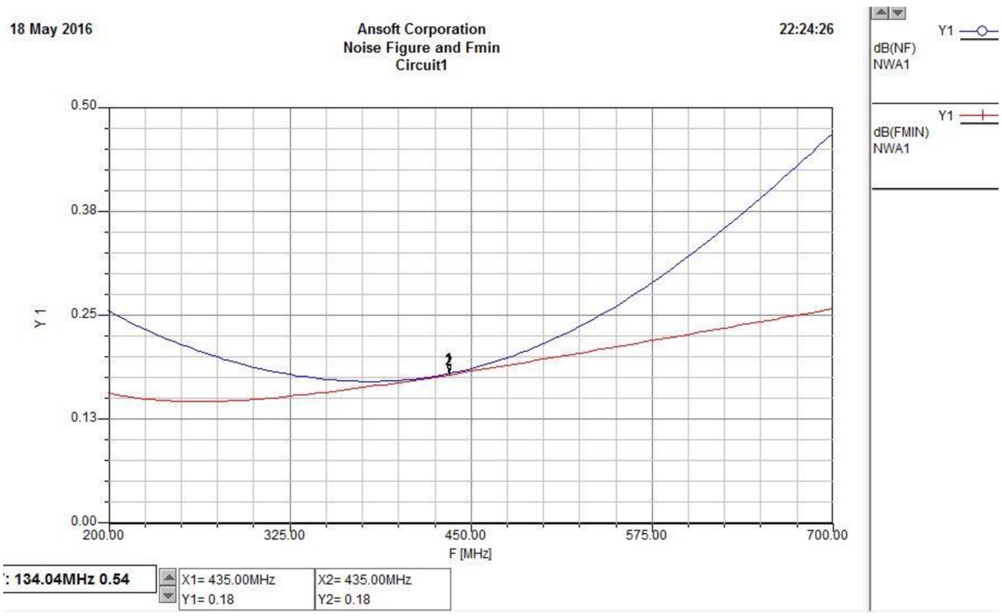


Figure 2.19: Noise figure and Minimum noise figure (Fmin).

In the next figure 2.20 are shown the Scatering parameters and the gain of the LNA for the center frequency of 435 MHz.

The gain obtained: 22.7 dB.

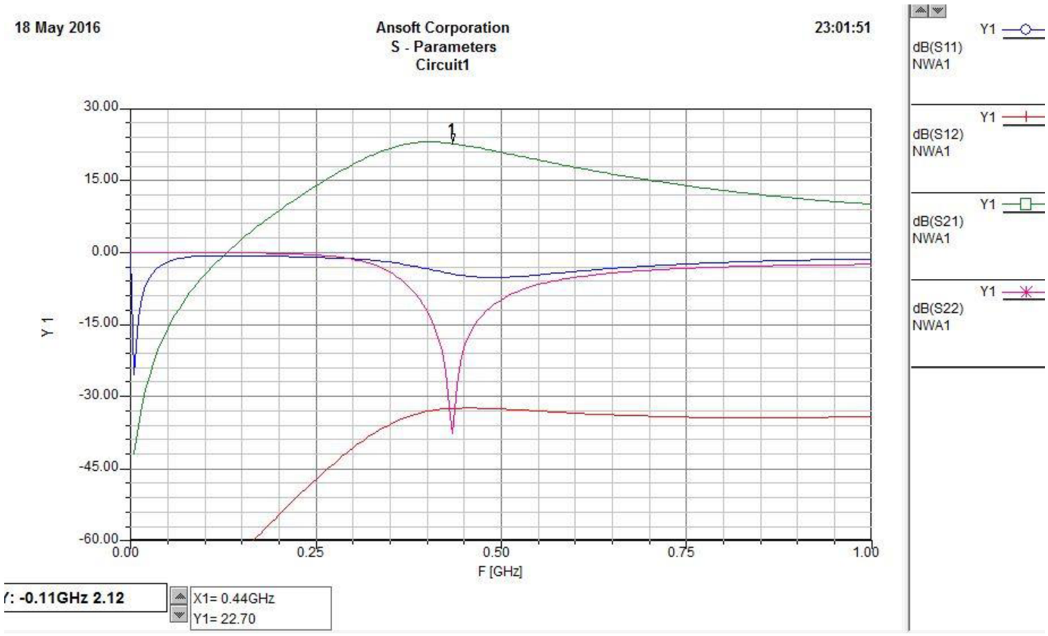


Figure 2.20: Gain of the LNA.

2.2 Design and simulation of Band Pass Filter

2.2.1 Designing Band Pass Filter

As it was mentioned in the chapter 1.3, a quarter-wave (λ/4-wave) coaxial resonator is used for the designing of band pass filter.

Firstly, it was necessary to calculate the approximate length that will have the quarter-wave (λ/4-wave) coaxial resonator for the center frequency required, the relation used for this calculation is defined by [2]

$$\frac{\lambda}{4} = \frac{\frac{c}{v}}{4}. \tag{2.15}$$

λ/4: Quarter-wavelength (m)
c: Speed of light ($3 \cdot 10^8 \mathrm{m/s}$)
v: Center frequency of the LNA ($435 \cdot 10^6 \mathrm{Hz}$)

The value of the quarter-wavelength of resonator obtained is

$$\frac{\lambda}{4} = \frac{\frac{3 \cdot 10^8}{435 \cdot 10^6}}{4} = 0.172\ m$$

The value obtained of the aproximation of the quarter-wavelength of coaxial resonator was considered long for this design of the LNA due to disadvantages of portability and non-easy adapting to small places. For this reason, it was decided to use a coaxial resonator of approximate lenght 60mm. The disadvantage of changing the length of the coaxial resonator is the variation of the center frequency or resonance frequency. In order to keep the center frequency of the LNA design, it was necessary to connect a variable capacitor (Trimmer) in parallel to the coaxial resonator.

It was used the coaxial cable LDF-2-50 (HELIAX) as resonator for this design, which has a good performance in RF signals. The parameters and dimensions of this coaxial cable are:

- Dimension of the inner conductor: 2.74 mm
- Dimension of the outer conductor: 10.29 mm
- Dielectric coefficient: 1.3-1.65
- Impedance: 50 Ω

2.2.2 Simulating Band Pass Filter

The next figure 2.21 shows the connection of the coaxial resonator in parallel with the variable capacitor C2 equal to 5pF. Additionally, the capacitors, C11 for the input and C10 for the output of the BPF, has been connected in order to improve the Q-factor quality of this filter.

Moreover, the capacitor C11 is also intended for decoupling the DC signal from the RF signal.

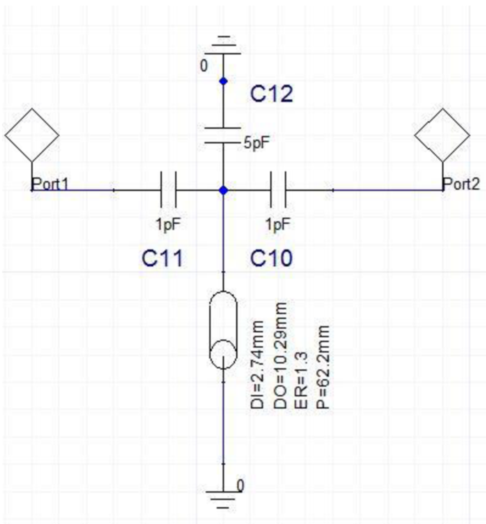


Figure 2.21: Band pass filtr with coaxial resonator

The parameter S21 shows the form, which the band pass filter is supposed to have. The parameters S11 and S22 demonstrate the stability of the filter for the frequency of 435 MHz. These parameters indicate the favorable stability of this band pass filter.

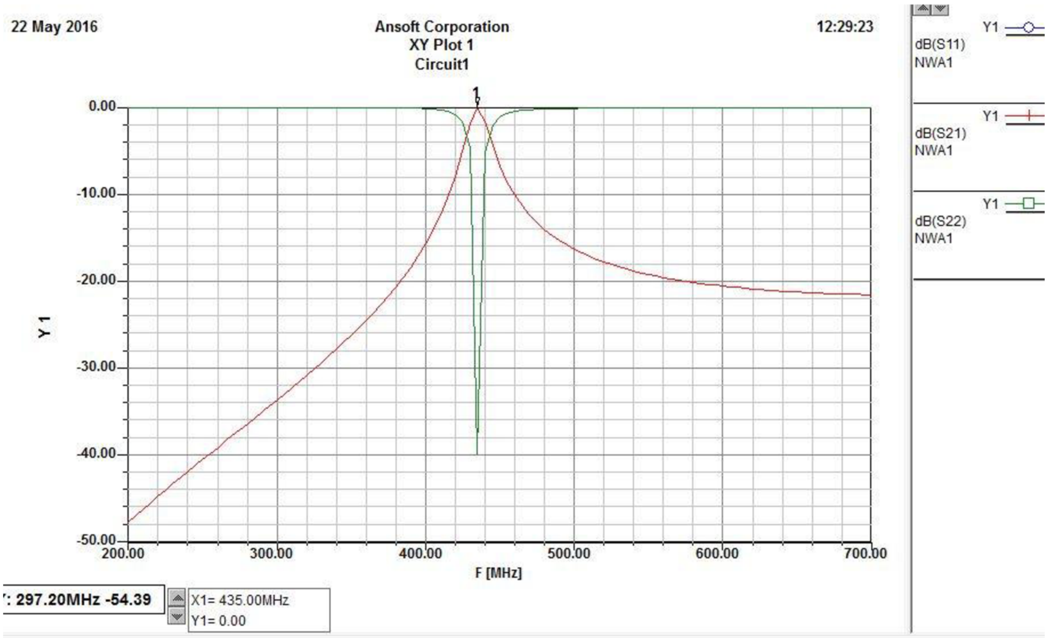


Figure 2.22: Simulation of the band pass filter

2.3 Simulation of the LNA with the Band Pass Filter

In this chapter was executed the simulation of the LNA with impedance matching networks and the Band pass filter both together as a principal circuit of this LNA design shown below in the figure 2.23.

The objective of this simulation was the obtaining of reference values of gain and noise figure of the LNA, which were taken into consideration during the realization.

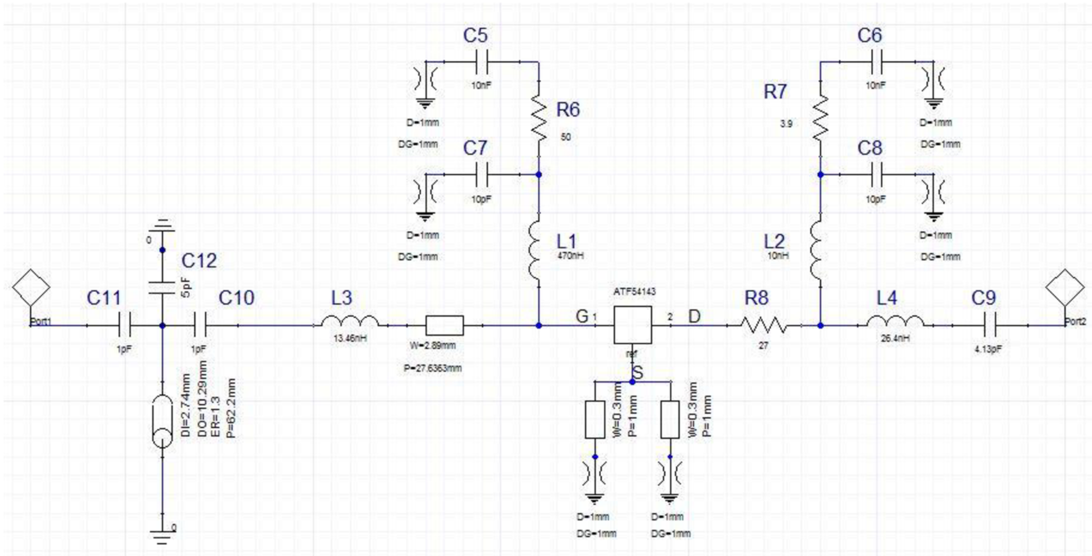


Figure 2.23: Principal scheme of the LNA

The figure 2.24 shows the results of gain and noise figure of the LNA obtained at the center frequency of this design.

From the figure the results are as follow:
- Gain=20.90 dB
- Noise figure=0.18 dB

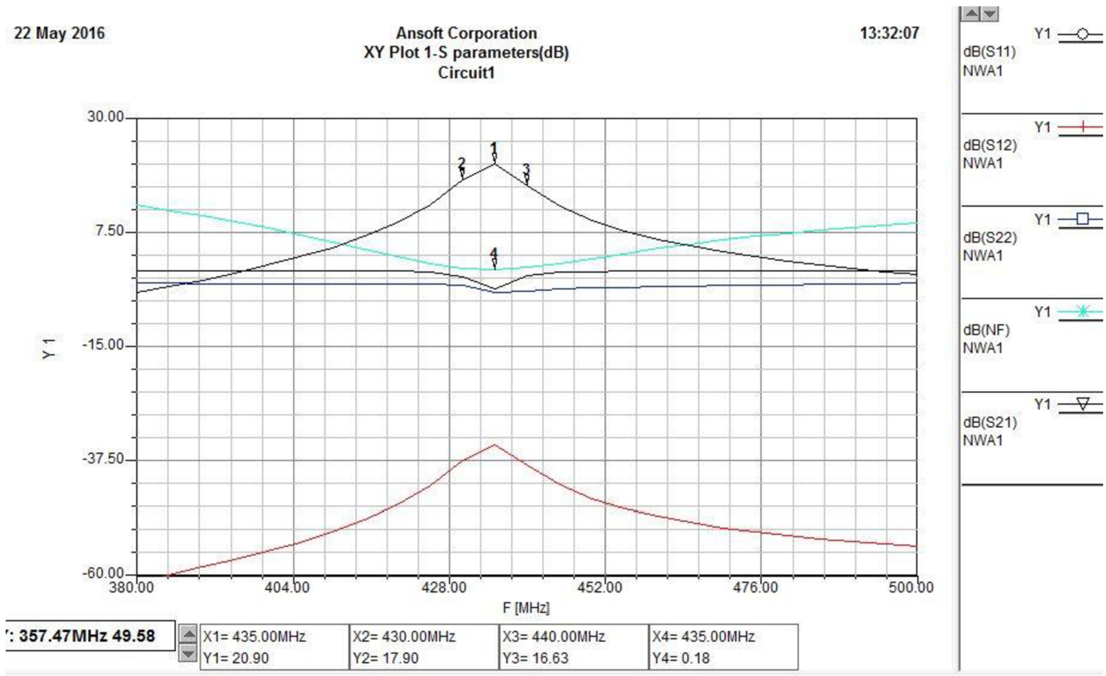


Figure 2.24: Simulation of the LNA

2.4 Realization and Measurement Results of the LNA

2.4.1 Realization Results of the LNA

For the realization of the Printed Circuit Board (PCB) of the LNA it was used EAGLE PCB design software version 7.5.0. The choice of this software as designing tool for the PCB of the LNA was because of flexibility in its capabilities and workflow compatibility, an extensive fully-open component libraries to work as well as a rich variety of tools and also its simplicity to use.

In the design of this LNA it has been considered the appropiate distribution of the spaces for the components, which have been implemented on the PCB. The location of the ATF-54143 transistor was defined carefully in order to obtain the lowest possible noise figure. To achieve this goal, it was considered during the design the shortest possible distance between the components of the input impedance matching network and the pin gate of the ATF-54143 transistor. The final design of the PCB with its real dimensions such as lenght 66mm, widht 45mm and height 1.5mm can be observed in Attachments A.1 and A.2, and the electronical scheme in A.3.

The PCB material used was FR-4 substrate, which was provided by the Faculty of Electrical Engineering and Communication of Brno University of Technology. The principal advantage of this material is its good performance in radio frequencies signals, which was suitable for this LNA design due to its work frequency in UHF band.

After the fabrication of the PCB previously designed, the next step was its implementation with the electronical components, which have been choosen and calculated for this design plan presented. They were carefully soldered on the PCB in order to avoid any damage on them and within the conductor lines of the PCB. On the other hand, it is important to know that the component values used were not ideal, but possible closest comercial ones. However, there were some components, which were not neither found with comercial values, so in this case it was necesary to elaborate them manually such as the inductors L2, L3 and L4 as shows Attachment A.1. Furthermore, it was decided to incorporate capacitors in parallel to the components of the DC bias circuit in order to avoid the possible not wanted influence of RF signals over them as illustrates Attachment A.1.

Figure 2.25: Implementation of the LNA with FR-4 substrate.

Figure 2.26: Low Noise Amplifier for UHF Band (430-440 MHz).

2.4.2 Measurement Results of the LNA

Once the implementation of the LNA was realized, the next and last step was its respective measurement with purpose of comparing of real and theorical values.

This measurement took action by 2 steps: Firstly an operating point measurement and consequently a gain and noise figure measurement done as last.

The values obtained from the first step can be observed in Table 2.4. The difference between the measured value and the theorical value of the operating point of the ATF-54143 transistor for V_{DS}, I_{DS} and V_{GS} respectively has been due to non-ideal values of the bias circuit components. However, they are considered within acceptable values for this LNA design.

Table 2.4: Measured value and calculated value of the operating point of the ATF-54143 transistor.

-	Measured value	Theorical value
V_{DS} (V)	2.7	3
I_{DS} (mA)	69	60
V_{GS} (V)	0.61	0.59

The second step was the measurement of the gain and the noise figure of the LNA respectively. For the measurement of the gain was used the FSP-Scalar Network Analyzer (9 KHz-7 GHz) - Rohde & Schwarz and two signal attenuators, which both were connected to the LNA as illustrates Figure 2.27. The FSP was set to supply -20 dBm of power signal, after that the measured input signal at the center frequency of 432 MHz was -38.27 dBm. Consequently, it was possible to obtain the gain value of the LNA of $\mathrm{G} = 21.73$ dB at the center frequency by the calculation between the output power signal, the input power signal and the attenuations origined in the circuit.

Following relation shows the calculation mentioned for the obtaining of the gain of the LNA:

$$G = C - B \tag{2.16}$$

$$B = A - 30dB = -20dBm - 30dB = -50dBm$$

$$C = D + 10dB = -38.27dBm + 10dB = -28.27dBm$$

The gain is:

$$G = -28.27dBm - (-50dBm) = 21.73dB$$

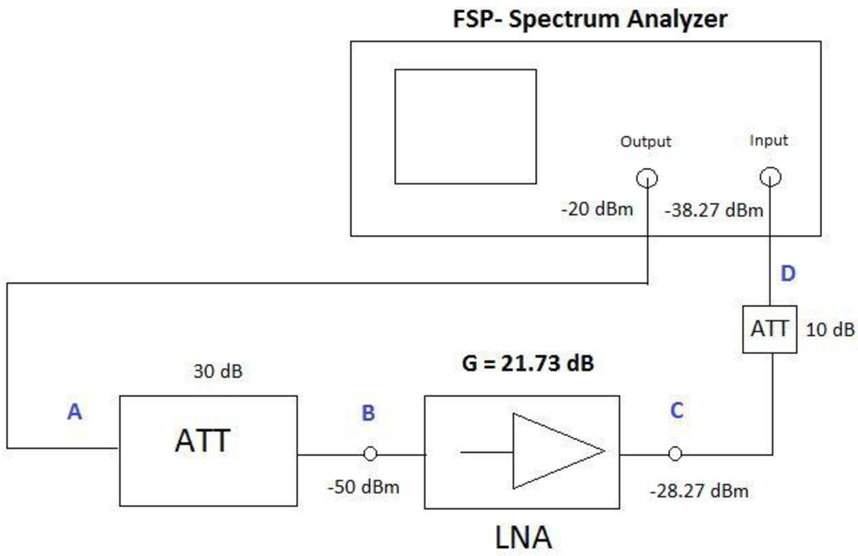


Figure 2.27 Measurement of the gain of the LNA.

Figure 2.28 shows the result in the FSP-Scalar Network Analyzer of the Signal Spectrum generated in the LNA at the center frequency of 432MHz. Moreover, it is possible to observe the optimal Q-Factor Quality of approximately 10 Mhz bandwidth of the amplifier.

The result of the gain of the LNA has been considered successful due to its obtained value, which was greater than the one obtained in the simulation part as shows Table 2.5. This result has been more than satisfactory within the range of expected values for this LNA design, which had a gain expectation within 15-20 dB.

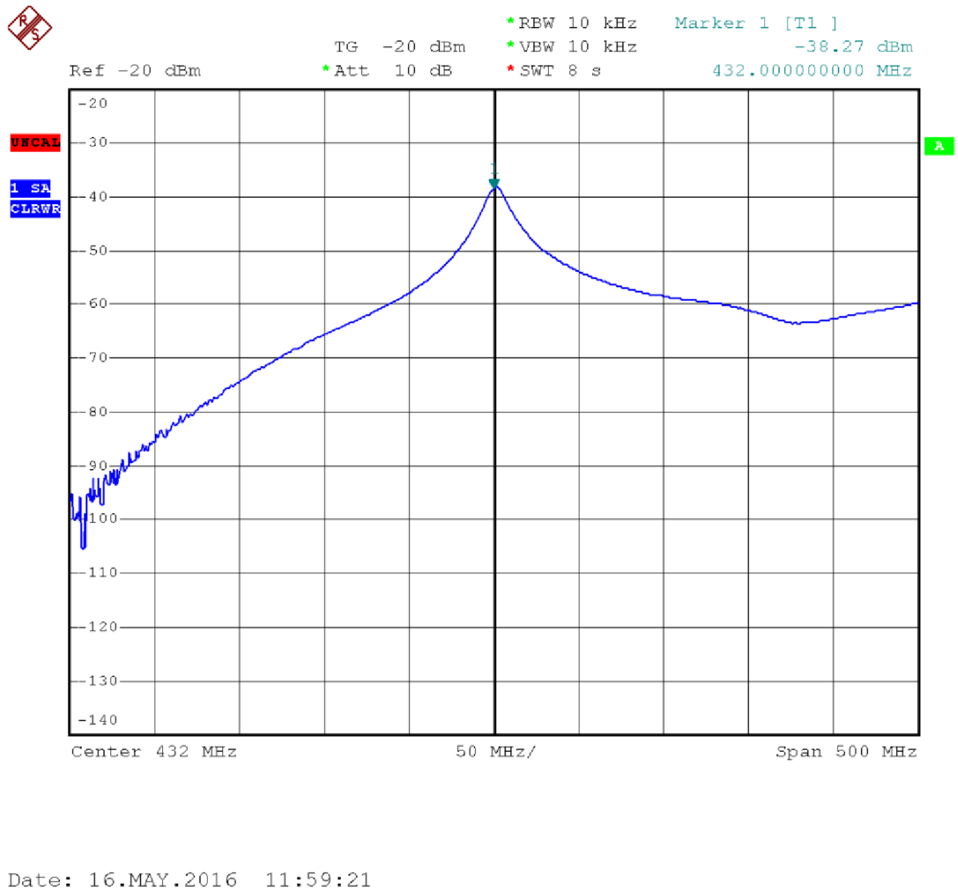


Figure 2.28: Measurement result of the gain.

After the measurement of the gain, it was realized the measurement of the noise figure. For this purpose it was used the NFA-Noise Figure Analyzer (10 MHz-26.5 GHz) - Agilent N8975A. This measurement was under room temperature conditions of 296.5 Kelvin (K), obtaining a result NF=2.049dB at the center frequency. The obtained value differs from the one simulated as it can be observed in Table 2.5. It has been concluded that some factors such as parameters and non-ideal values of the implemented components could be the cause, which affected and conveyed an influence over the result.

Figure 2.29 shows the value of noise figure obtained at the center frequency of 432 MHz.

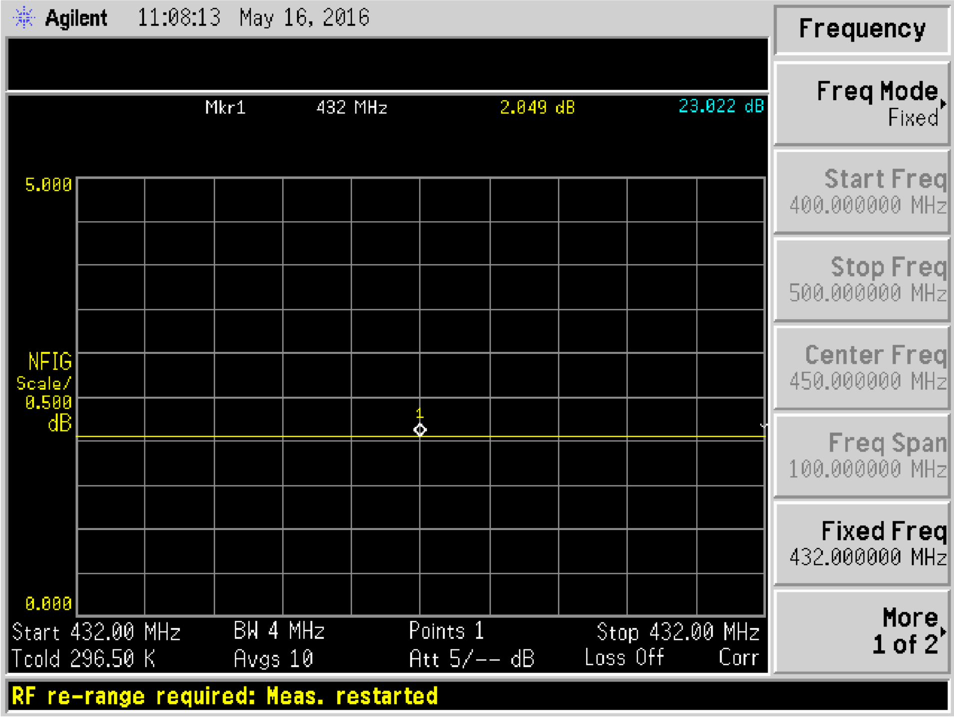


Figure 2.29: Measurement of the noise figure of the LNA.

Table 2.5: Measured value and simulation value of the gain and noise figure of the ATF-54143 transistor.

-	Measured value	Simulation value
G (dB)	21.73	20.90
NF (dB)	2.049	0.18

3 CONCLUSION

The thesis presented the design of a Low Noise Amplifier (LNA) for UHF band 430-440MHz. The main objectives of this design were to obtain a gain of 15-20dB, a minimum insertion noise and input and output impedance of 50 ohms. A very important tool used for designing and improving the LNA has been the Ansoft Designer SV2 simulation software. The ATF-54143 transistor has been chosen as a main device due to its good characteristics in the electromagnetic spectrum of the RF, such as high linearity, low noise figure, and power gain.

The results gained due to the implementation of the LNA at the operating center frequency of 432MHz and with the DC power supply of 5V for the bias circuits are: The gain of 21.73dB with the bandwidth quality factor Q of approximately 10MHz, and the noise figure of 2.049dB. The value of the gain may be considered as very satisfactory with respect to the requirements, and the bandwidth quality factor Q is within the required range. The value of the noise figure is considered as acceptable despite the fact that it does not match the value previously obtained in the design simulation. Considering the noise as a disturbance affecting all the circuits, there are some factors that could influence the value results of the noise figure - for example the component implementation on the PCB board, low quality parameters of the used substratum, and components without perfect parameters and values.

The Low Noise Amplifier design developed in this Bachelor´s thesis has met the established requirements. It has considered aspects, which would improve the performance of this LNA, such as the utilization of the substratum of better quality, utilization of discrete components, and new modeling technologies.

BIBLIOGRAPHY

[1] RADMANESH, M. *RF and Microwave Circuit Design Essentials*. Author House, 2007. ISBN 978-1-4259-7241-7.

[2] VIZMULLER, P. RF *Design Guide, Systems, Circuits, and Equations*. Norwood: Artech House, 1995. ISBN 0-89006-754-6.

[3] WHITE, J.F. *High Frequency Techniques:an introduction to RF and microwave engineering.* New Yersey: John Wiley & Sons,2004. ISBN 471-45591-1.

[4] *"IRE standards on methods of measuring noise in linear twoports, 1959,"* Proceedings of the IRE, vol. 48, no. 1, January 1960.

[5] Pozar, D.M., *Microwave Engineering*. John Wiley and Sons, Inc., 2000.

[6] Kurokawa, K. *"Power waves and the scattering matrix,"* IEEE Transactions on Microwave Theory and Techniques, vol. 13, no. 2, 1965.

[7] Gonzalez, G. *Microwave Transistor Ampliers*. Prentice Hall, 2000.

[8] R. Ludwig and P. Bretchko, *RF Circuit Design: theory and applications*. Pearson Education, 2000.

[9] AVAGO *Technologies ATF 54143 Data Sheet 2012*, [online] 11.04.2016. Available from URL: <http://www.avagotech.com/docs/AV02-2026EN>.

[10] RHODE, U. L., NEWKIRK, D. P. RF/*Microwave Circuit Design for Wireless Applications*. New York: John Wiley & Sons, 2000.

[11] HANUS, S., SVAČINA, J. *Vysokofrekvenční a mikrovlnná technika*. Skriptum. Brno: FEKT, VUT v Brně, 2002.

INDEX OF SIMBOLS, QUANTITIES AND ABBREVIATIONS

BPF Band Pass Filter.

dB Decibel.

DC Direct Current.

E-pHEMT Enhancement pseudomorphic High Electron Mobility Transistor.

FET Field Efect Transistor.

FR-4 Flame retardant type 4

Fmin Minimum Noise Figure.

LNA Low Noise Amplifier.

GACS Available gain circle for source.

GACL Available gain circle for load.

KCS K-factor circle for source.

KCL K-factor circle for load.

MAG Maximum Available Gain.

MSG Maximum Stable Gain.

NF Noise Figure.

PCB Printed Circuit Board

RF Radio Frequency.

RL Return Loss.

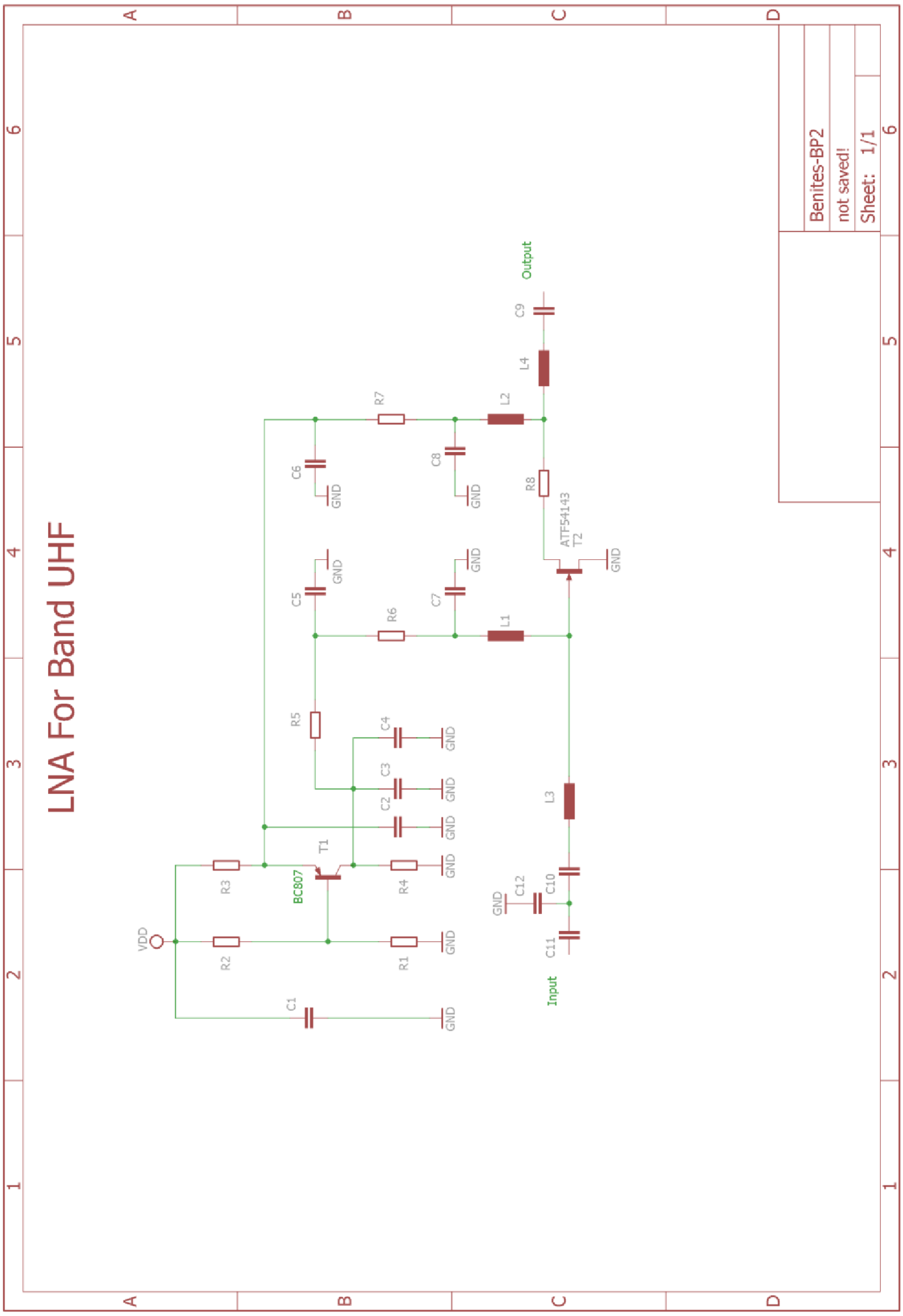
SMD Surface Mount Devices.

SOT Small Outline Transistor.

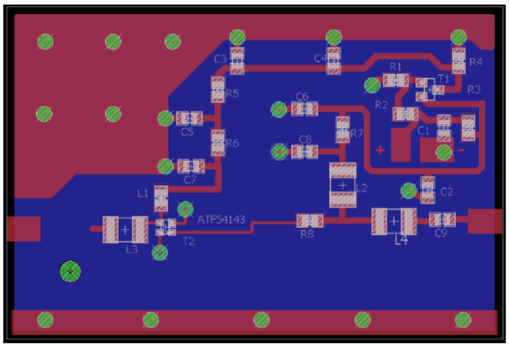
UHF Ultra High Frequency.

A LNA DESIGN

A.1 Circuit Diagram of the LNA

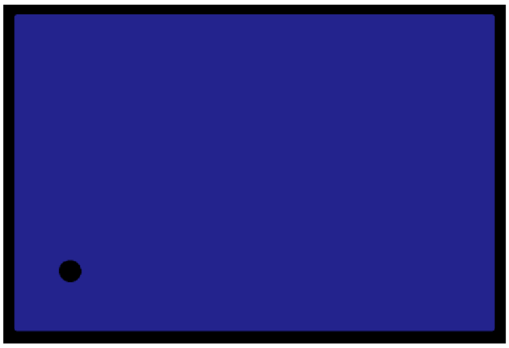


A.2 Printed Circuit Board (PCB) – Top (Component side)



Dimensions of board 45 x 66 [mm], scale M1:1

A.3 Printed Circuit Board (PCB) – Bottom (Connection side)



Dimensions of board 45 x 66 [mm], scale M1:1

B LIST OF COMPONENTS

Label	Value	Case	Description
C1	10nF	0805	Ceramic capacitor
C2	10nF	0805	Ceramic capacitor
C3	10nF	0805	Ceramic capacitor
C4	10nF	0805	Ceramic capacitor
C5	10nF	0805	Ceramic capacitor
C6	10nF	0805	Ceramic capacitor
C7	10pF	0805	Ceramic capacitor
C8	10pF	0805	Ceramic capacitor
C9	5pF	0805	Ceramic capacitor
C10	1pF	-	Ceramic disc capacitor
C11	1pF	-	Ceramic disc capacitor
C12	0.8-10pF	5201 piston	Johanson Technology
L1	470nH	0805	Johanson Technology
L2	10nH	-	Designed
L3	13.5nH	-	Designed
L4	26.5nH	-	Designed
R1	2kΩ	0805	Multicomp
R2	390Ω	0805	Multicomp
R3	2.7 Ω	0805	Multicomp
R4	1kΩ	0805	Multicomp
R5	10kΩ	0805	Multicomp
R6	47Ω	0805	Multicomp
R7	3.9Ω	0805	Multicomp
R8	27Ω	0805	Multicomp
T1	BC807	SOT-23	Philips
T2	ATF-54143	SOT-343	Avago Technologies
X1	50Ω	SMA	Connector
X2	50Ω	SMA	Connector