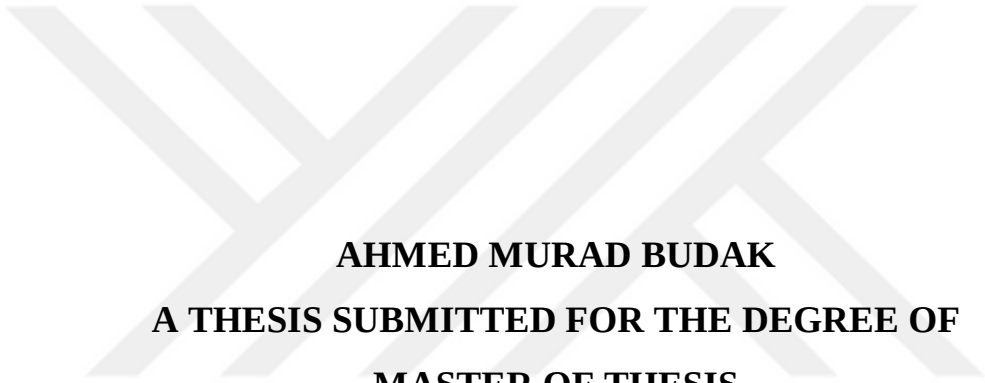


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HIGH GAIN BROADBAND SP6T MATCHED RF SWITCH



AHMED MURAD BUDAK
A THESIS SUBMITTED FOR THE DEGREE OF
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GEBZE

2023



YÜKSEK LİSANS JÜRİ ONAY FORMU

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SUMMARY

This thesis examines a system with high gain, broadband rf amplifiers that are switched based on the operating frequency. Bandpass filters are designed and utilized with rf amplifiers in this system. The purpose of bandpass filters is to suppress the amplification of signals outside the specified frequency range while passing the signals within the selected frequency range. Seven rf low noise amplifiers and one SP6T rf switch were used in the development of this thesis. The rf switch has six input ports and one output port for each distinct frequency range in addition to the control gates. The signal in the desired frequency range can be amplified and measured by controlling the rf switch. The rf switch suppresses the other five frequency ranges while measuring in one. Therefore, measurements are not taken from unnecessary input ports. AWR Microwave Office and Sonnet Software are used in the design of this thesis.

Key Words: Rf Low Noise Amplifier, Bandpass Filter, SP6T Rf Switch

ÖZET

Bu tezde, yüksek kazançlı ve geniş bantlı rf güç yükselticileri ile birlikte bunları çalıştırları frekansa göre anahtarlayan bir sistem üzerinde çalışılmıştır. Bu sistemde rf yükselticiler ile birlikte bant geçiren filtreler de tasarlanmıştır ve kullanılmıştır. Bant geçiren filtrelerin kullanılma amacı seçilen frekans aralığında sinyalin, rf yükselticileri sayesinde yükseltilmesi ve seçilen frekans aralığı dışında kalan sinyallerin bastırılarak yükseltilmesini engellemektir. Bu tezin tasarımında yedi adet rf düşük gürültülü yükseltici ve bir adet SP6T rf anahtar kullanılmıştır. Rf anahtarın kontrol için olan kapılarının dışında her ayrı frekans aralığına göre altı adet giriş kapısı ve bir adet çıkış kapısı bulunmaktadır. Rf anahtar kontrol edilerek istenilen frekans aralığında gelen sinyal yükseltilerek ölçülebilmektedir. Bir frekans aralığında ölçüm yapılırken diğer beş frekans aralığı rf anahtar tarafından bastırılmaktadır. Böylelikle istenilmeyen giriş kapılarından ölçüm yapılmamaktadır. Bu tezin tasarımı sırasında AWR Microwave Office ve Sonnet programları kullanılmıştır.

Anahtar Kelimeler: Rf Düşük Gürültülü Yükseltici, Bant Geçiren Filtre, SP6T Rf Anahtar

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LIST OF ABBREVIATIONS AND ACRONYMS

<u>Abbreviations and Acronysm</u>	<u>Explanations</u>
RF	: Radio frequency
Γ	: Reflection coefficient
VSWR	: Voltage standing wave ratio
Z	: Impedance
Hz	: Hertz
Z_0	: Characteristic impedance
S	: Scattering
P	: Power
DC	: Direct current
λ	: Wavelength
Π	: Pi
e^{γ}	: Propagation constant
Y	: Admittance
T	: Temperature
k	: Boltzman constant
F	: Noise factor
NF	: Noise figure
μ	: μ - factor
dB	: Decibel
dBm	: Decibel miliwatt
DR	: Dynamic range
P1dB	: 1 dB compression point
SFDR	: Spurious free dynamic range
G	: Gain
LNA	: Low noise amplifier
PCB	: Printed circuit board
ε_r	: Dielectric constant
L	: Inductance

C	:	Capacitance
SMD	:	Surface mount devices
H	:	Henry
F	:	Farad
VNA	:	Vector network analyzer
Cad	:	Computer aided design
MHz	:	Mega Hertz
GHz	:	Giga Hertz
BPF	:	Bandpass Filter
IMN	:	Input matching network
OMN	:	Output matching network
EM	:	Electromagnetic
MMIC	:	Monolithic Microwave Integrated Circuit
SNR	:	Signal to noise ratio
UWB	:	Ultra wide band
K	:	Stability factor
P_{out}	:	Output power

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1. INTRODUCTION

All circuits produced in the design part of this thesis are intended to operate in ISM and GPS frequency bands between 900 MHz and 7125 MHz. Since the frequency range of the design is extremely broadband, the application space of the design in this thesis expands significantly.

Seven LNAs and one RF switch were included in the design within the scope of the thesis. LNAs are used as RF amplifiers as the system is designed to operate in receiver mode. Additionally, the RF switch operates as SP6T. The RF switch has six input RF ports and one output RF port. A broadband LNA circuit is designed on the output port of the RF switch. This LNA circuit operates between DC and 8000 MHz. In addition to the broadband LNA circuit, six LNA circuits are designed for the RF input ports of the RF switch. However, these six LNA circuits are designed to operate at specific frequencies and are not wideband. The LNA circuits at the six input ports of the RF switch are referred to as transmission line circuits in accordance with their frequency ranges (RF1, RF2, RF3,... RF6). The performance ranges of the seven LNAs to be used in their frequency bands are as follows:

LNAs	Frequency Range	Gain Range (dB)
RF1	6175 MHz – 7125 MHz	15-10
RF2	5725 MHz – 5875 MHz	12-10
RF3	2400 MHz – 2500 MHz	12-10
RF4	1555 MHz – 1595 MHz	15-13
RF5	1170 MHz – 1278 MHz	18-15
RF6	902 MHz – 928 MHz	20-17
Broadband LNA	DC – 8000 MHz	25-10

Table 1.1 Frequency ranges and gains of LNA circuits

Bandpass filters are designed for six LNA circuits operating at the input of the RF switch based on the frequencies they operate simultaneously. The bandpass filters are produced using microstrip lines and discrete components. Four distributed bandpass filters and two SMD bandpass filters were designed for this thesis.

The LNAs chosen for the design had extremely low noise, but because the manufacturers did not specify the noise figure parameters, simulation could not be performed on these parameters throughout the design. In addition, MMIC LNAs have been selected with a cost of no more than \$6. Moreover, since the product used for the thesis would work at extremely low signal levels, there is no possibility of LNA saturation.

Since it is a large-scale project, one of the requirements of the project is to use MMIC LNA components instead of transistors in LNA circuits. In addition, the usage of LNA MMICs simplified the DC bias section of the project in comparison to the usage of transistors. This is because the matching performances of most LNAs in the desired frequency ranges are below expectations. Moreover, the RF switch circuit must be matched with the LNA circuits in the network as a whole. Placing components on a 50 ohm transmission line without matching would drastically lower the RF performance of the entire system and render the product useless. Therefore, each LNA circuit, RF switch circuit, and bandpass filter was matched as realistically as possible in the simulation.

In chapter 2, noise, stability, linearity, gain in low noise amplifiers and simulation approach will be examined.

In chapter 3, the design and simulations of bandpass filters designed for six frequency bands will be discussed.

In chapter 4, the design and simulations of LNA circuits and RF switch will be discussed, considering the parameters in chapter 2. Afterwards all system simulations will be checked.

In chapter 5, the measurement results of the manufactured PCB will be evaluated and compared with the simulation results.

Chapter 6 will include reviews about the results and future work of this thesis.

2. LOW NOISE AMPLIFIER FUNDAMENTALS AND SIMULATION APPROACH

LNA (low noise amplifier) is one type of RF amplifier having a low noise characteristic. A low SNR (signal to noise ratio) is characteristic of a low noise amplifier. Noise, gain, impedance matching and reflection coefficients are important LNA design factors [1], [2]. In addition, the P1 dB compression point of the LNA should also be considered. Its linearity becomes crucial if the LNA is to be used on higher signals as opposed to at very low signal levels. Figure 2.1 depicts a block diagram of an LNA.

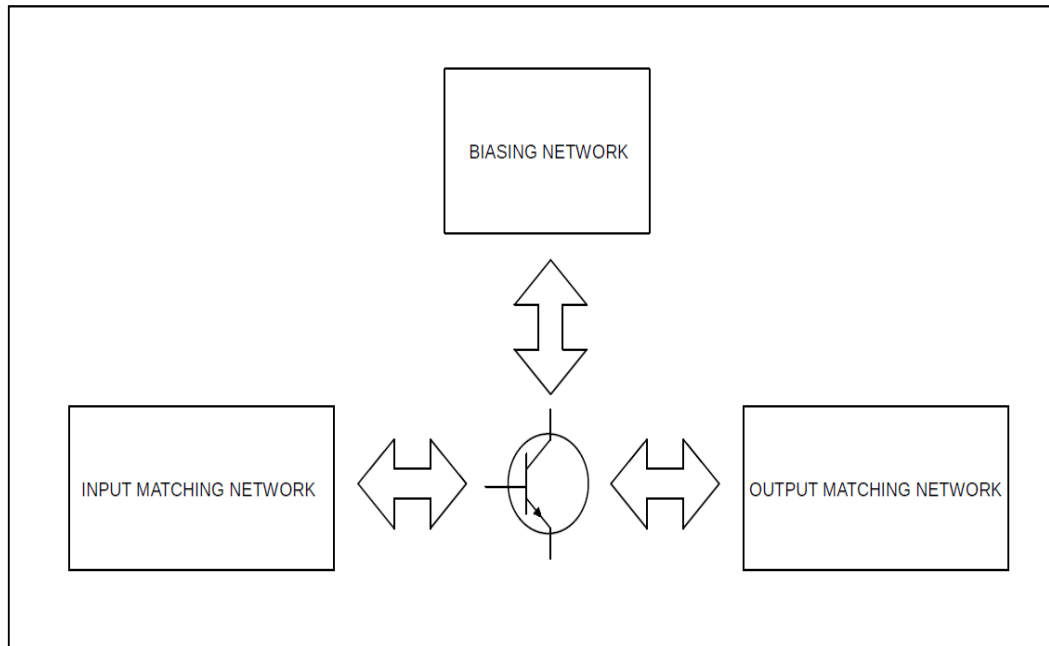


Figure 2.1 Schematic illustration of a simple LNA

LNA design actually contains many challenges. These are to provide adequate gain and good input and output matching with as low noise as possible [3], [4].

2.1 Noise Analysis In Low Noise Amplifier

Noise is any signal other than the desired signal in communication systems and it restricts the minimal signal level that a circuit may handle with acceptable values. There are numerous noise sources that generate various noise types. These are referred to as inherent and interference noise [5].

Electronic devices and electronic circuits in the surroundings generate inherent noise. This noise can only be diminished, never eliminated. Interference noise is produced by interference in the microwave circuit as contrast to inherent noise. Therefore, interference noise can be decreased by making the necessary adjustments to the microwave circuit.

Thermal noise is one of the inherent noises. The voltage measured along a conductor would fluctuate even if there was no current flowing through it. Due to its thermal origin, the thermal noise power spectrum density is exactly proportional to the absolute temperature (T) and is expressed by the equation in (2.1) as the useable noise power [5], [6].

$$P_{NA} = kT \Delta f \quad (2.1)$$

k is the Boltzman constant; its approximate value is 1.380649×10^{-23} J/K. T is the temperature in Kelvin and Δf represents the frequency bandwidth of the noise measurement. [7].

In addition to thermal noise, shot noise and 1/f noise are also inherent noises [8]. The random flow of carriers across the potential barrier causes shot noise. For this noise to occur, a potential barrier must exist between the current flow and charge carriers. On the other hand, nonlinear devices are responsible for this noise. Moreover, shot noise is independent of temperature [7]. 1/f noise is the other sort of inherent noise (flicker noise). 1/f noise is a sort of inherent noise that predominates at low frequencies or low-frequency oscillator offsets.

Typically, the noise performance of a circuit is denoted by a parameter referred as the noise factor (F) and expressed by the equation (2.2).

$$F = \frac{(\frac{S}{N})_{in}}{(\frac{S}{N})_{out}} = \frac{\text{total output noise power}}{\text{output noise power from input source}} \quad (2.2)$$

2.1.1 Noise Factor In Cascade Amplifier Circuits

The input signal travels through multiple stages in a microwave amplifier circuit, each of which decrease the signal to noise ratio. We can determine the noise factor (or noise temperature) at the junction points of the cascade systems if we know the noise factor (or noise temperature) of each stage separately. The noise factor is crucial in the first stage of microwave amplifier circuits [9], [10], [12].

The noise temperature of the composite microwave circuit generated by the gradual connection of two circuits is calculated as follows with two cascade-connected microwave circuits with respective gains G_1 and G_2 and noise temperatures T_{e1} and T_{e2} :

$$T_{cas} = T_{e1} + \frac{1}{G_1} T_{e2} \quad (2.3)$$

The noise factor (or noise temperature) of the first stage microwave circuit has a significant impact on the noise factor of the entire amplifier circuit, as shown by equation (2.3).

2.2 Stability Analysis In Low Noise Amplifier

Stability of a low noise amplifier (LNA) is crucial throughout the design phase and can be determined by the S parameters. Oscillation is possible under specific conditions in a two-port circuit with an amplifier. The reflection coefficients will be $|\Gamma_{IN}| > 1$ or $|\Gamma_{OUT}| > 1$ when negative resistance is generated at the input or output ports of the amplifier [9]. This condition is represented by the symbols $|S_{11}| > 1$ and $|S_{22}| > 1$ when the impedances of the amplifier's input and output ports have no effect on each other's load (unilateral) [13].

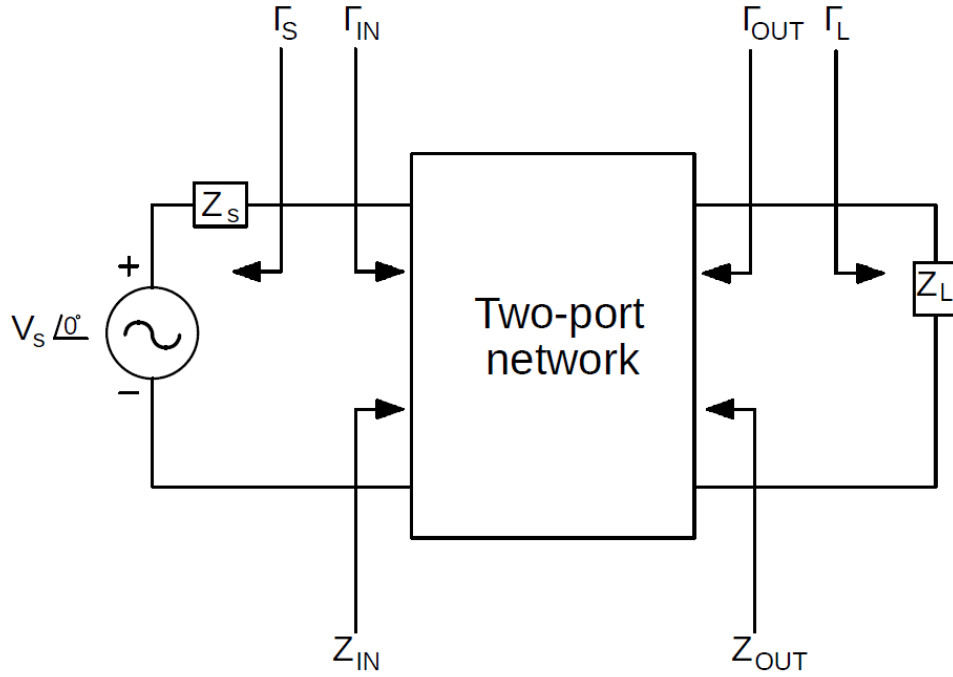


Figure 2.2 Block diagram of a two port amplifier network circuit

Figure 2.2 depicts a two-port network that is unconditionally stable at a given frequency if the real parts of Z_{IN} and Z_{OUT} for all passive load and source impedances are greater than zero. The two-port network is potentially stable if it is not unconditionally stable. So, certain passive load source terminations may generate input and output impedances with a negative real part [9], [11].

In terms of reflection coefficients, when it is unconditionally stable at a given frequency it is expressed as:

$$|\Gamma_s| < 1 \quad (2.4)$$

$$|\Gamma_L| < 1 \quad (2.5)$$

$$|\Gamma_{IN}| = \left| S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \right| < 1 \quad (2.6)$$

$$|\Gamma_{OUT}| = \left| S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S} \right| < 1 \quad (2.7)$$

Even though the real parts of Z_{IN} and Z_{OUT} are positive, it is possible that the absolute values of the reflection coefficients Γ_L and Γ_S are greater than 1 if the two-port system is potentially stable. The circles $|\Gamma_{IN}|=1$ and $|\Gamma_{OUT}|=1$ must be determined for potentially stable systems in the smith chart.

$$\text{Provided that } \Delta = S_{11}S_{22} - S_{12}S_{21};$$

The Γ_L value determines the radius of $|\Gamma_{IN}|=1$ (Output Stability Circle).

$$\Gamma_L = \left| \frac{S_{12}S_{21}}{|S_{22}|^2 - |\Delta|^2} \right| \quad (\text{radius}) \quad (2.8)$$

The center point of the output stability circle is obtained as follows:

$$C_L = \frac{(S_{22} - \Delta S_{11}^*)^*}{|S_{22}|^2 - |\Delta|^2} \quad (2.9)$$

The Γ_S value determines the radius of $|\Gamma_{OUT}|=1$ (Input Stability Circle).

$$\Gamma_S = \left| \frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2} \right| \quad (\text{radius}) \quad (2.10)$$

The center point of the input stability circle is obtained as follows:

$$C_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} \quad (2.11)$$

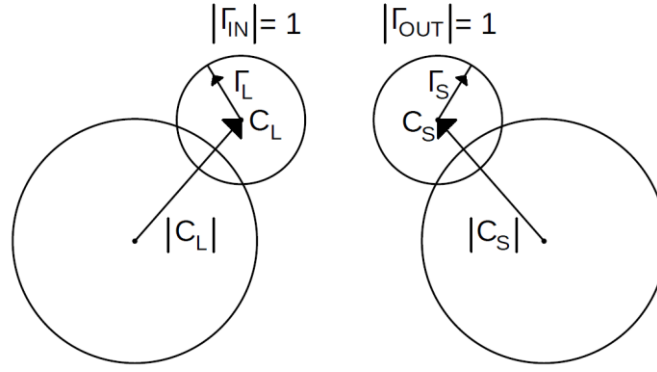


Figure 2.3 Input and output stability circles

2.2.1 Test For Unconditional Stability In Amplifiers

There are other methods for determining whether an amplifier is unconditionally stable or potentially stable besides using stability circles. The Rollet method is one of these methods [20].

$$K \triangleq \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1 \quad (2.12)$$

The amplifier is unconditionally stable, as shown by equation (2.12). Additionally, there is the μ -factor test for unconditional stability and is represented as [18]:

$$\mu \triangleq \frac{1 - |S_{11}|^2}{|S_{22} - S_{11}^* \Delta| + |S_{12}S_{21}|} > 1 \quad (2.13)$$

2.3 Linearity In Low Noise Amplifier

The dynamic range can easily be defined as the power range obtained by dividing the maximum input signal that the RF system can tolerate by the minimum meaningful input signal (P_{SENS}). The maximum input signal is somewhat unclear. This could mean an input signal level where the gain of the RF system is compressed by 1 dB. 1 dB value may vary depending on the characteristics of the amplifier used. The dynamic

range (DR) can be calculated as follows if 1 dB of compression on the system and application is taken into consideration as the upper limit [14]:

$$DR = P_{in-P1db/dBm} - P_{sens/dBm} \quad (2.14)$$

It is also known as the P1dB point in amplifiers where there is 1 dB compression at output power.

However, it is crucial to determine what determines the signal's upper limit in cases in which the data in the amplified signal cannot be carried by the amplitude of the input signal. Here, the focus is on SFDR (Spurious-free dynamic range). The SFDR is the interval between the smallest detectable signal in a system just above the noise floor and the maximum signal level that may be transmitted without perceptible distortion. The third-order IM product's power becomes equal to the noise floor's power when the maximum permitted input signals for a two-tone test are applied to SFDR [14], [17]. To calculate the upper bound, equation (2.15) can be written as:

$$P_{OIP3} = \frac{\Delta P}{2} + P_{out} \quad (2.15)$$

P_{OIP3} represents output third-order intercept point, $\frac{\Delta P}{2}$ represents the difference in dBm between one of the fundamental output signals and the third-order product seen in figure 2.4 [25].

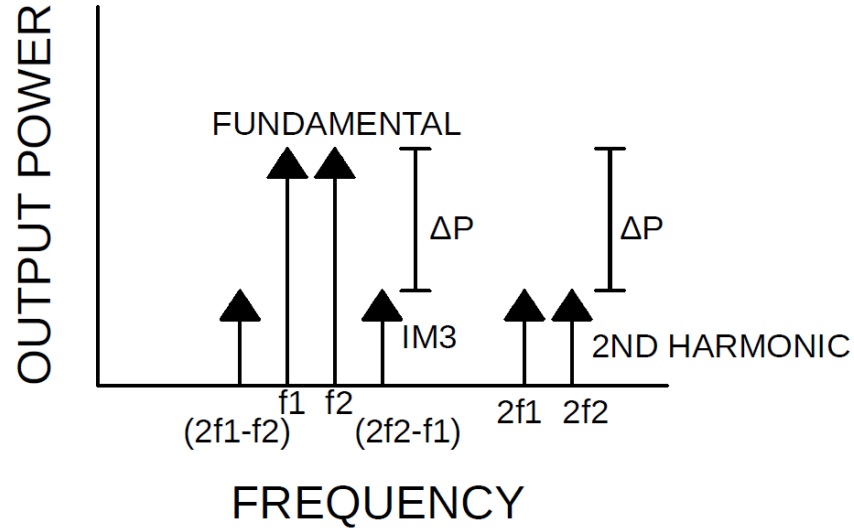


Figure 2.4 The output power levels contain the two amplified tones, IM3 (Intermodulation Distortion), third-order distortion product and harmonic distortions

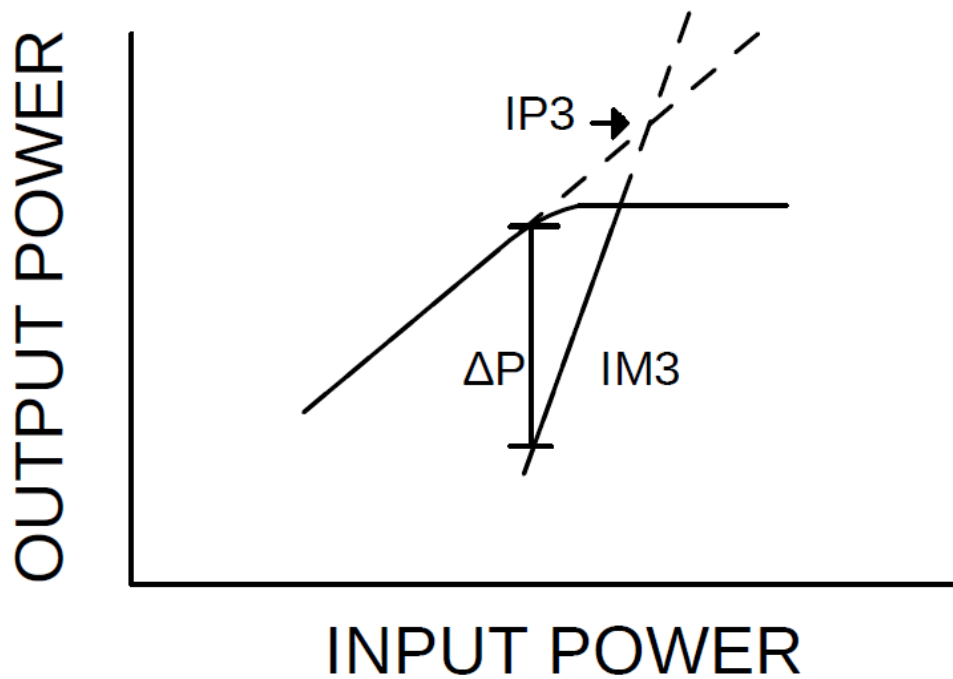


Figure 2.5 Third-order intercept point

P_{IIP3} can be specified input third-order intercept point and G_{dB} represents the gain of the amplifier. From equation (2.15), SFDR can also be calculated as:

$$P_{IIP3} + G_{dB} = \frac{P_{out} - P_{out-IM3}}{2} + P_{in} + G_{dB}$$

$$P_{IIP3} = \frac{P_{in} + G_{dB} - P_{in-IM3} - G_{dB}}{2} + P_{in}$$

$$P_{IIP3} = \frac{3P_{in} + P_{in-IM3}}{2}$$

Thus,

$$P_{in/dBm} = \frac{2P_{IIP3} + P_{in-IM3}}{3}$$

If set to input noise floor related to third order intermodulation,

$$P_{in,max/dBm} = \frac{2P_{IIP3} + (-174dBm + NF(db) + 10 \log(\Delta f))}{3} \quad (2.16)$$

The SFDR is found as in (2.17):

$$SFDR = P_{in,max/dBm} - P_{sens/dBm}$$

$$SFDR = \frac{2(P_{IIP3} + 174dBm - NF - 10 \log(\Delta f))}{3} - SNR_{min/dBm} \quad (2.17)$$

2.4 Gain Analysis In Low Noise Amplifier

Input and output matching networks related to the gain of the circuit are designed after the stability of the amplifier has been determined. G_T (transducer gain) represents the gain, which expresses the energy transfer of the microwave amplifier network.

$$G_T = \frac{|S_{21}|^2 (1 - |\Gamma_S|^2)}{|1 - \Gamma_S \Gamma_{in}|^2 |1 - S_{22} \Gamma_L|^2} \quad (2.18)$$

In case $G_T < G_{Tmax}$, the transducer gain expression is not useful for designing amplifier circuits in terms of gain. [19]. For $G_T = G_{Tmax}$, the condition must be as:

$$\Gamma_{in} = \Gamma_S^* \quad (2.19)$$

$$\Gamma_{out} = \Gamma_L^* \quad (2.20)$$

On the other hand, G_T is also expressed as the ratio of the power supplied to a load to the power supplied from the source [15].

$$G_T = \frac{\text{Power delivered to the load}}{\text{Power available from the source}} \quad (2.21)$$

$$G_T = \frac{|S_{21}|^2}{|1-\Gamma_S|^2} \frac{|1-\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} = \frac{|S_{21}|}{|S_{12}|} (K - \sqrt{K^2 - 1}) \quad (2.22)$$

Other gain units than G_T are also expressed in amplifier networks. These are G_A and G_P . G_A means that maximum available power gain and G_P means that operating power gain [16].

$$G_A = \frac{\text{Power available from the network}}{\text{Power available from the source}} \quad (2.23)$$

$$G_P = \frac{\text{Power delivered to the load}}{\text{Power delivered to the network}} \quad (2.24)$$

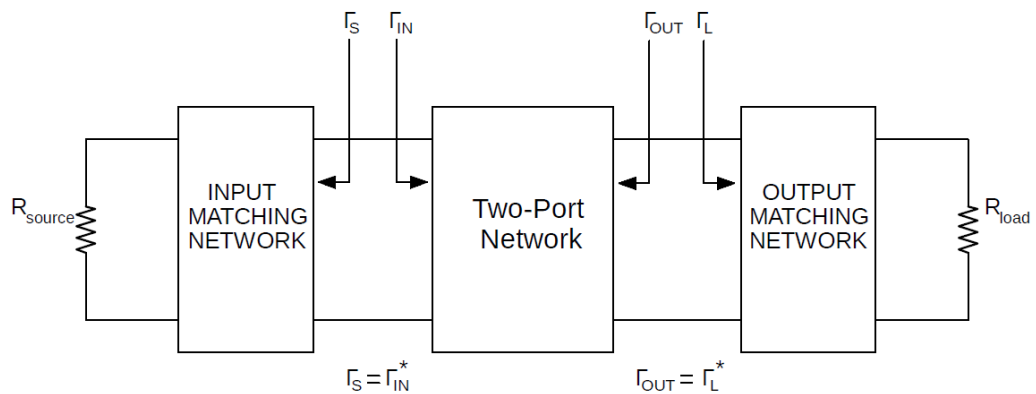


Figure 2.6 Maximum gain transmitting amplifier circuit block diagram

Conjugate matching cannot always be used to achieve optimal matching and maximum gain delivery in LNA networks.. The main goal of LNA circuits, which generally operate in receive mode, is to keep a low noise figure in addition to achieving a high gain. These criteria are considered during the design process and the appropriate matching is done.

2.5 Simulation Approach

Noise, stabilization, gain, and matching techniques in low noise amplifier must be well-understood in order for simulations to be performed accurately. The design was designed using simulation tools NI AWR and Sonnet Software after understanding these techniques.

Simulations were conducted as small signal S parameters simulation. Unfortunately, nonlinear data of LNAs from manufacturers could not be obtained. Therefore, nonlinear simulation could not be performed. However, this is not a problem since the product designed within the scope of thesis will operate at a much lower signal level than the saturation point of LNAs.

Evaluating if the matching requirements were met was the most important aspect of S parameter simulations. No LNA noise data could be acquired. Nonetheless, only LNAs with a low noise figure are selected.

2.5.1 Design Process

The port locations on the PCB were determined at the beginning of the thesis design process in accordance with the project's requirements. Then, bandpass filters for each frequency range were designed. The designed bandpass filters relate to the frequency ranges presented in Table 1.1. The LNA circuits were designed based on six frequency ranges once the bandpass filters were designed. Afterwards, a new broadband LNA circuit was designed. Then, the layouts of the RF switch and the LNA circuits were analyzed and designed. Schematic designs were performed initially, later electromagnetic designs and optimizations were done for each of these designs. Thus, it is desired that the simulation results closely resemble the measurement results. The design phase followed the series in figure 2.5.

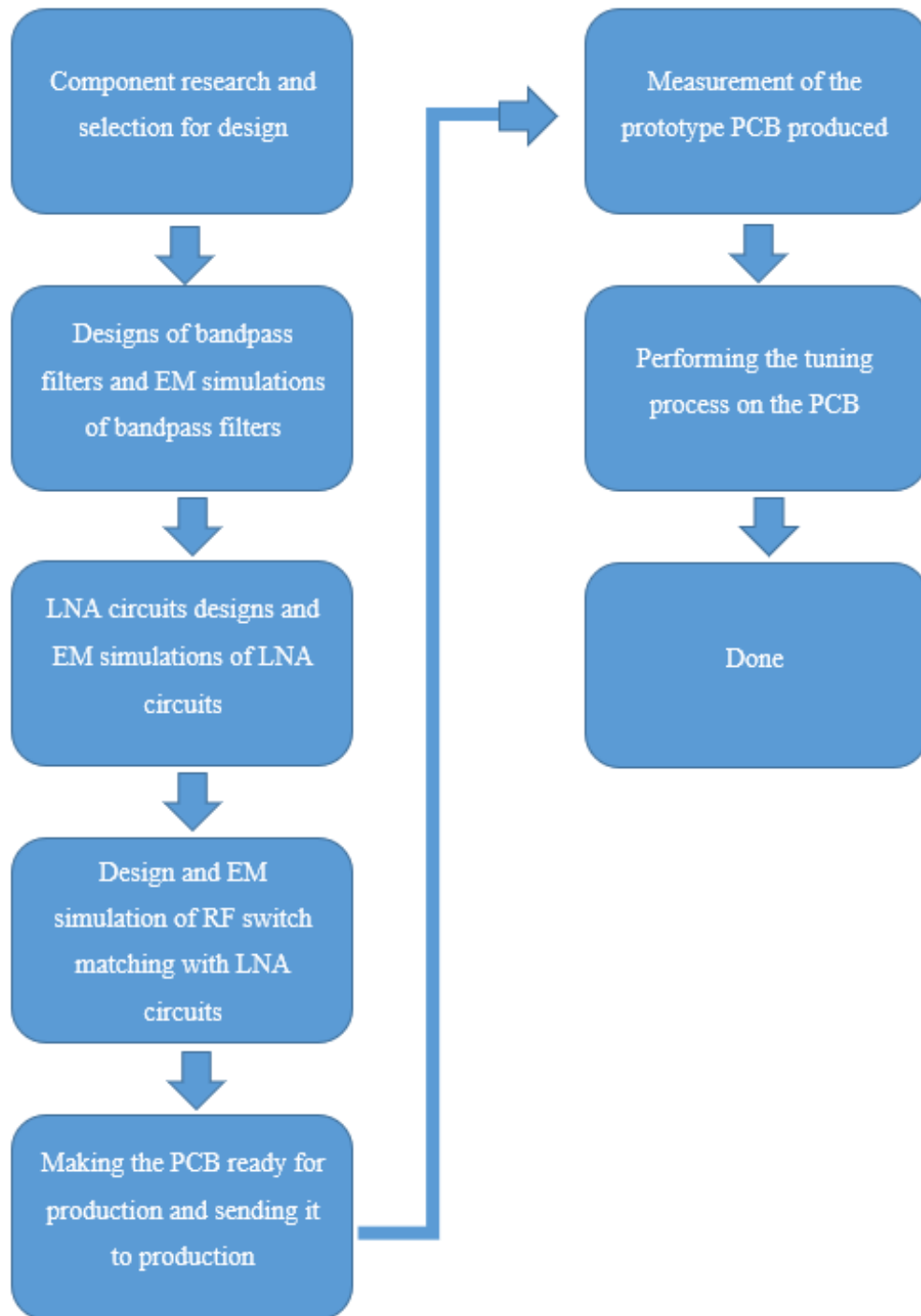


Figure 2.7 Flowchart of design process

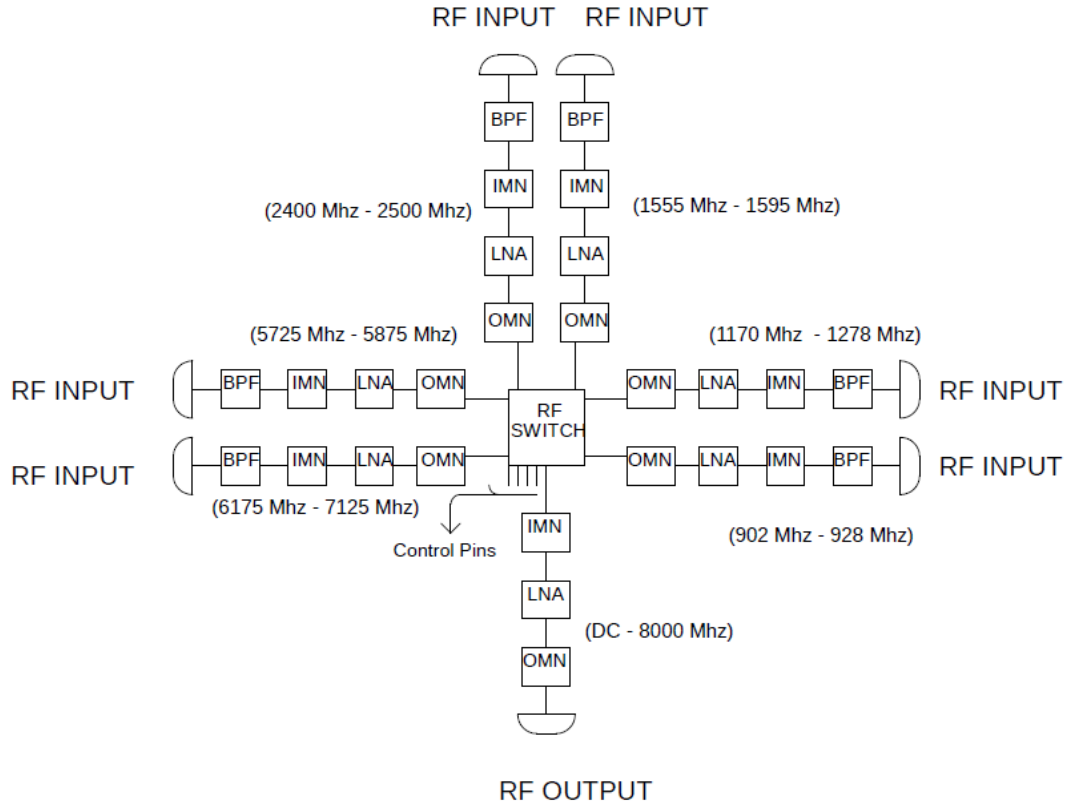


Figure 2.8 High gain broadband SP6T matched switch matrix block diagram

This design is manufactured as a PCB (printed circuit board). The PCB material with a dielectric constant of 2.2 and a thickness of 0.75 millimeters has been selected to facilitate matching procedures. Since the capacitance between the conductor on the dielectric and the conductor under the dielectric is small, the dielectric coefficient was chosen as 2.2. Equations in (2.25) and (2.26) illustrate this situation:

$$Z_0 = \sqrt{\frac{L}{C}} \quad (2.25)$$

$$Z_0 = \frac{30\pi b}{\sqrt{\epsilon_r} \times We + 0.441b} \quad (2.26)$$

Where We is the effective width of the center conductor and b is the height of the dielectric material.

The transmission line's characteristic impedance has been increased by decreasing the transmission line's capacitance in the design. Therefore, the capacitance must be increased in order to decrease the impedance. This requires expanding the transmission line's width. This allowed the system to be adapted to accommodate wider transmission lines. The main benefit of this condition is that the impedance protection tolerance significantly increases as a result of production defects.

In addition, the selected dielectric material has a low loss tangent value. Therefore, it was intended to prevent losses resulting from a dielectric material.



3. DESIGN STAGES AND SIMULATIONS OF BANDPASS FILTERS

Microwave filters, which are two-port networks used in microwave circuits to control the frequency response in a specific frequency range by ensuring transmission at specific frequencies in the passband and attenuation in the stopband, are an essential component of microwave systems [20]. Bandpass filters for microwaves are designed by connecting a lowpass filter and a highpass filter in series in a variety of configurations. These bandpass filters prevent the transmission of incoming signals below a given frequency and above another frequency. Therefore, transmission occurs between the desired frequency values and signals with frequencies outside of this range cannot flow through the system.

It is required to design a highpass filter and a lowpass filter and to match these filters to one another to design a bandpass filter in the typical sense. There are two methods to filter design. These are the image parameter method and the insertion loss method. Bandpass filters are designed utilizing the insertion loss method in this thesis.

The insertion loss filter design method generates a filter with a fully specified frequency response using network synthesis techniques. This method provides greater control over the passband and synthesizes the desired response from the amplitude and phase properties of the stopband. In this case, the filter design should satisfy the application specifications under optimal conditions [21]. The insertion loss method controls the high-order bandpass and bandstop amplitude and phase characteristics [22]. The insertion loss power ratio is explained as follows in the insertion loss method:

$$P_{LR} = \frac{P_s}{P_l} = \frac{1}{1 - |\Gamma(w)|^2} \quad (3.1)$$

P_l means the power reaching the load, P_s refers to the power sent from the power source, and Γ refers to the reflection coefficient at the input port of the network.

The insertion loss method may be applied in various ways, depending on the desired insertion loss characteristic. These alternatives include Butterworth and Chebyshev cases. The Butterworth method is preferred when designing bandpass

filters for this thesis. Simulations show that the Butterworth method is utilized due to the flatter insertion loss response.

The Butterworth method is implemented as a filter that responds by modifying the insertion loss in the desired intervals as little as possible. This system is used for the lowpass filter based on the formula in (3.2):

$$P_{LR}=1+k^2(w/w_c)^{2n} \quad (3.2)$$

W_c shows the frequency that expresses the cutoff frequency of the low pass filter, and n shows the order of the low pass filter in the equation (3.2). The k value is a constant used to express the value in the amount of fluctuation when there is a cut in the bandpass.

Bandpass filters are designed for six frequency bands after the bandpass filter design method is chosen.

3.1 902 MHz - 928 MHz Bandpass Filter Design

The bandpass filter is designed with SMD (surface mount device) components in this frequency range. This is because the wavelength in this frequency range is between 0.33 meters and 0.32 meters. It would occupy a lot of space if the bandpass filter was designed using microstrip transmission lines. Therefore, SMD components are used in the bandpass filter design for convenience in production.

The bandpass filter is designed with 9 degrees of discrete components. The schematic design of the bandpass filter with ideal values is shown in Figure 3.1:

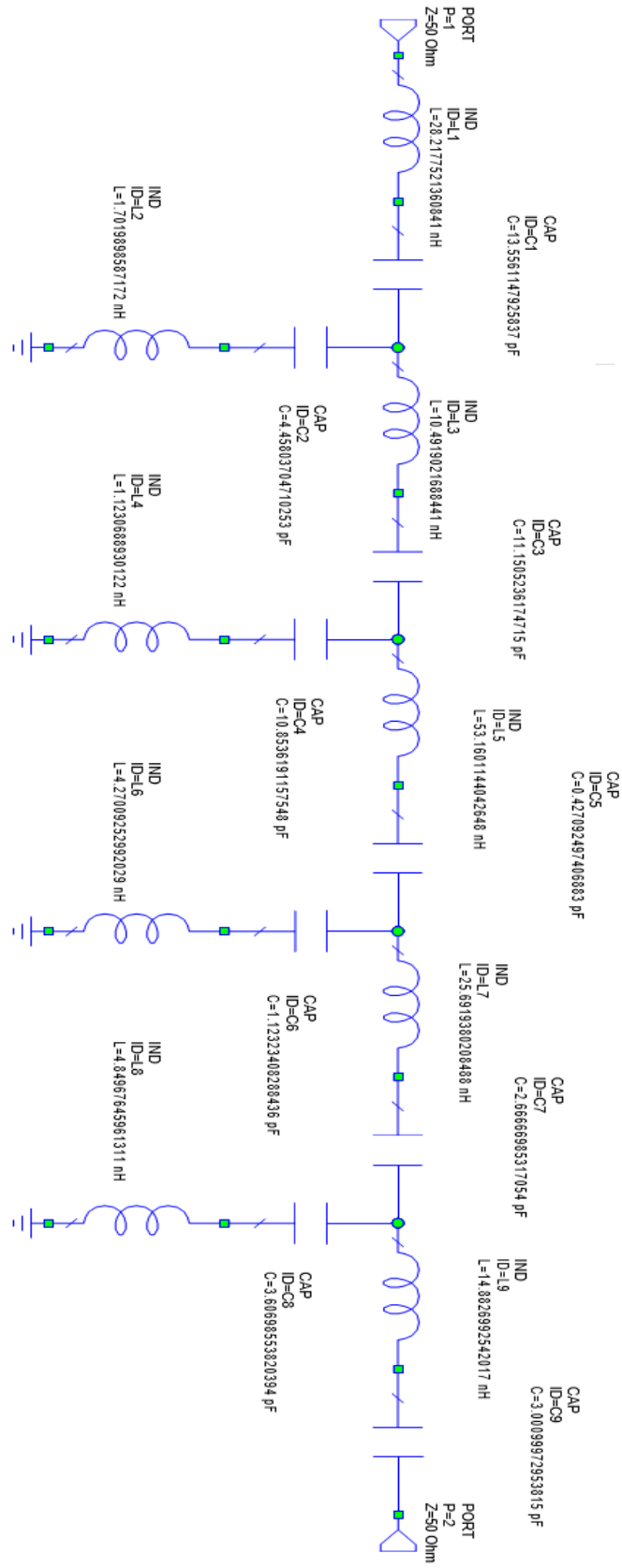


Figure 3.1 Bandpass filter between 902 MHz and 928 MHz designed with ideal elements

The simulation result of the bandpass filter in Figure 3.1 is as follows:

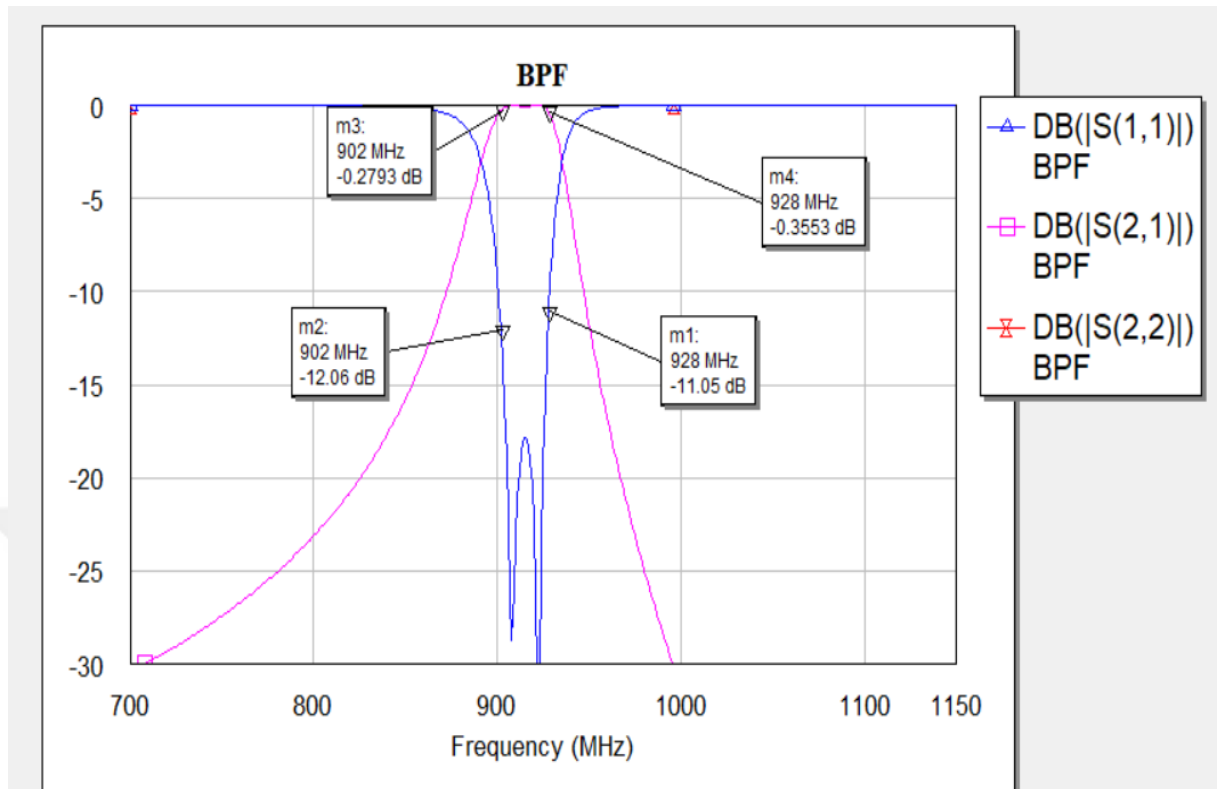


Figure 3.2 Bandpass filter simulation result between 902 MHz - 928 MHz designed with ideal elements

The simulation results have reached the desired level as seen in the Figure 3.2, according to the schematic design. The layout was designed utilizing 0603 standard capacitors and 0805 standard inductors for SMD components after the bandpass filter was designed with ideal values.

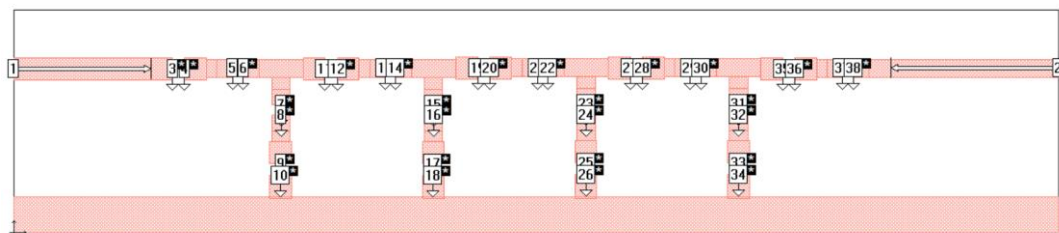


Figure 3.3 Bandpass filter layout between 902 MHz - 928 MHz

The SMD component values were adjusted again after the electromagnetic simulation. The component values were arranged above the standard values.

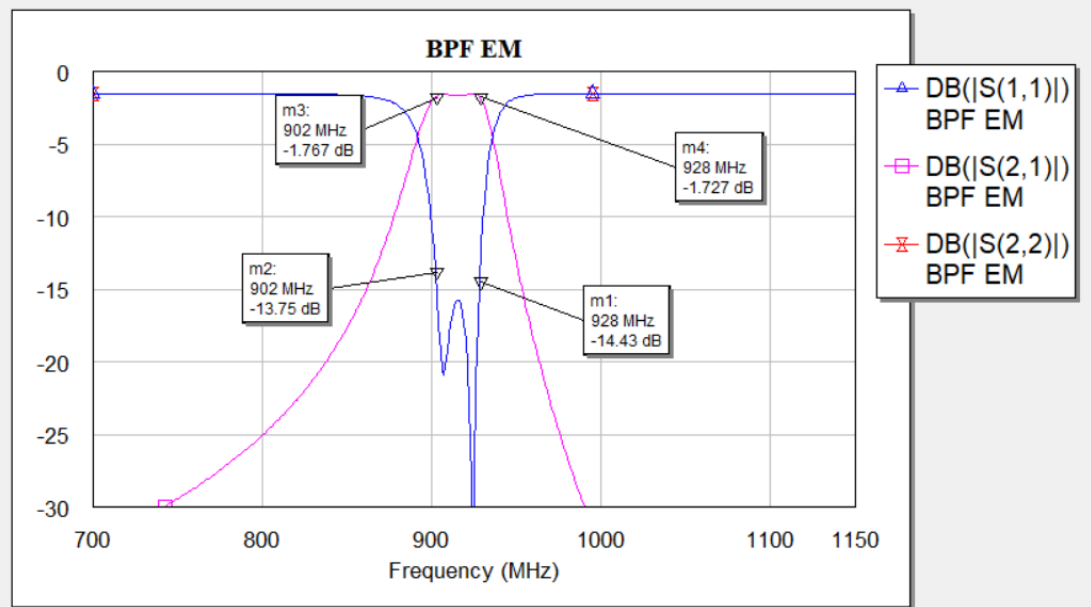


Figure 3.4 Layout simulation result of bandpass filter between 902 MHz - 928 MHz

As seen in the Figure 3.4, the desired result was obtained as a result of the EM simulation. The fact that the S11 and S22 values in the specified frequency range are less than -10 dB and the insertion loss value is as minimal as feasible is the most crucial indicator of this circumstance.

3.2 1170 MHz - 1278 MHz Bandpass Filter Design

This bandpass filter is also designed using discrete components. The schematic design and simulation result of the designed bandpass filter are as follows:

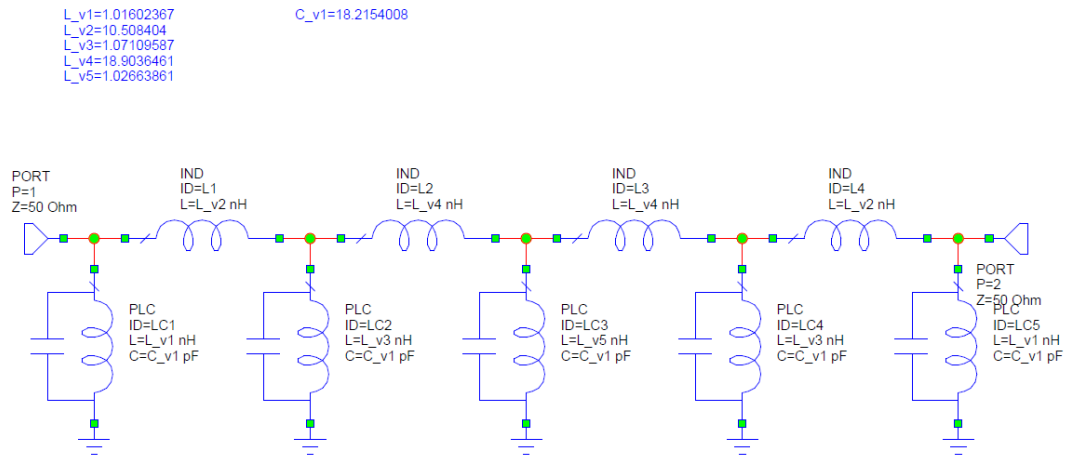


Figure 3.5 Bandpass filter between 1170 MHz - 1278 MHz designed using ideal elements

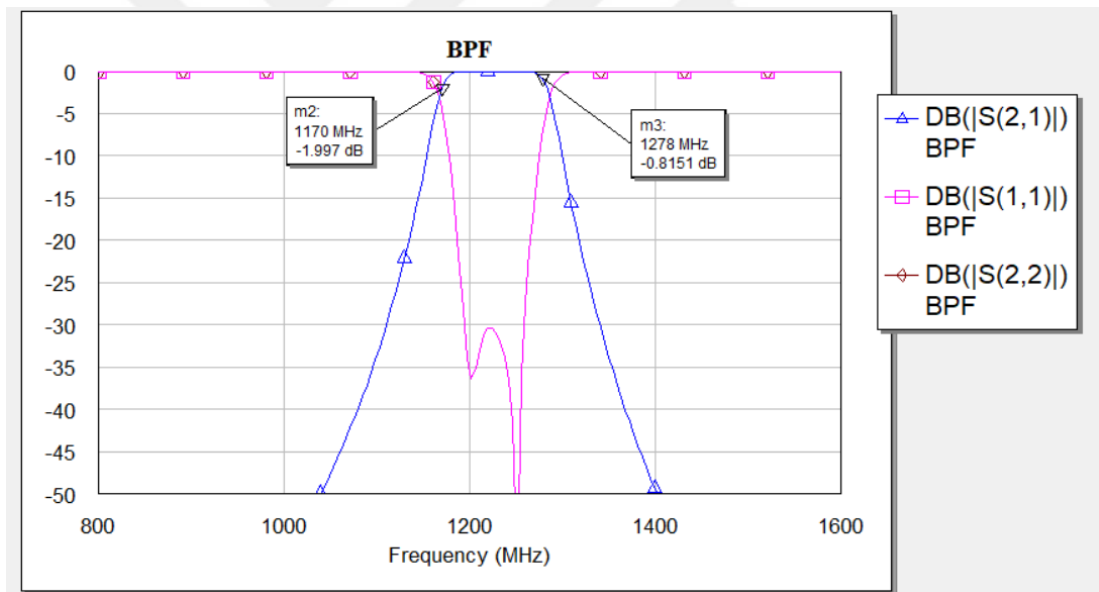


Figure 3.6 Bandpass filter simulation result between 1170 MHz - 1278 MHz designed using ideal elements

The simulation results have reached the desired level as seen in the Figure 3.6, according to the schematic design. After that, the layout was designed using discrete components. EM simulation results are as follows:

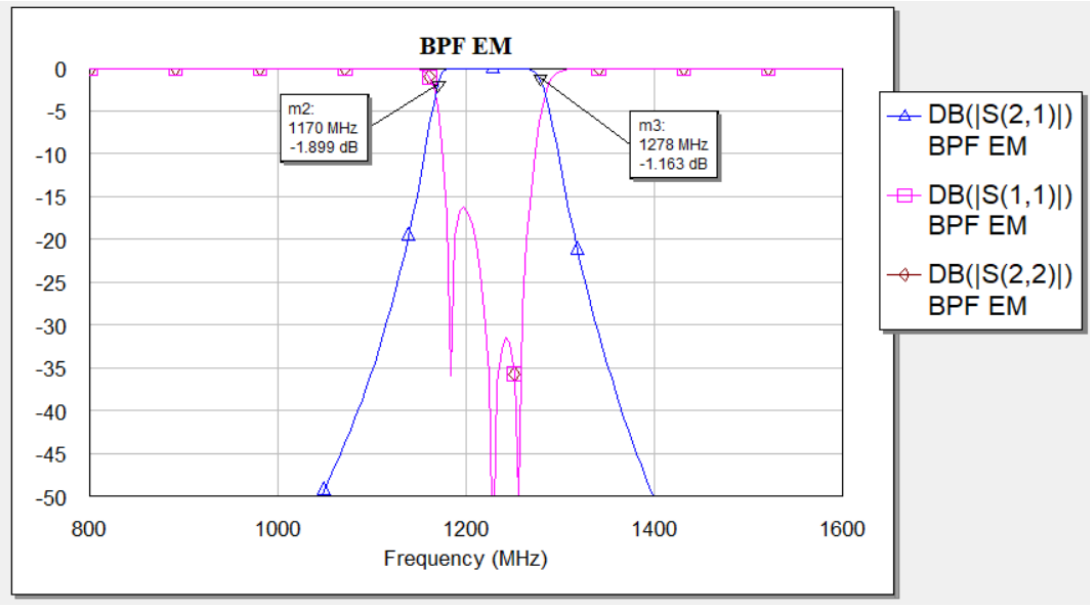


Figure 3.7 Layout simulation result of bandpass filter between 1170 MHz - 1278 MHz

EM simulation results are considered sufficient as seen in the Figure 3.7.

3.3 1555 MHz - 1595 MHz Bandpass Filter Design

Increasing the operational frequency results in parasitic effects on SMD components. The primary reason for this is that as the frequency of operation increases, the wavelength decreases, causing capacitance and inductance values on the SMD component to deviate from their ideal values. The parasitic effects change depending on the SMD component's package size as the frequency increases, which impacts the microwave circuit's performance [23], [24]. So, the bandpass filter is built with microstrip lines for this frequency range. This design was done as a hairpin bandpass filter to take up the least amount of space possible on the PCB. After that, the bandpass filter was put on the PCB the same way it was designed.

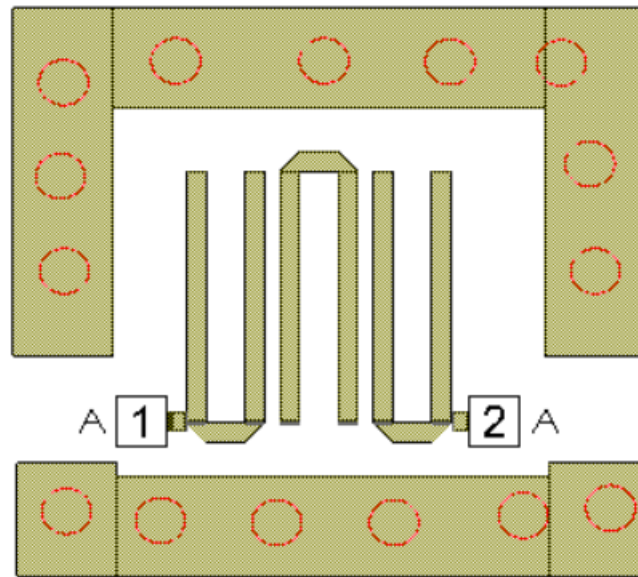


Figure 3.8 Bandpass filter between 1555 MHz - 1595 MHz designed using microstrip transmission lines

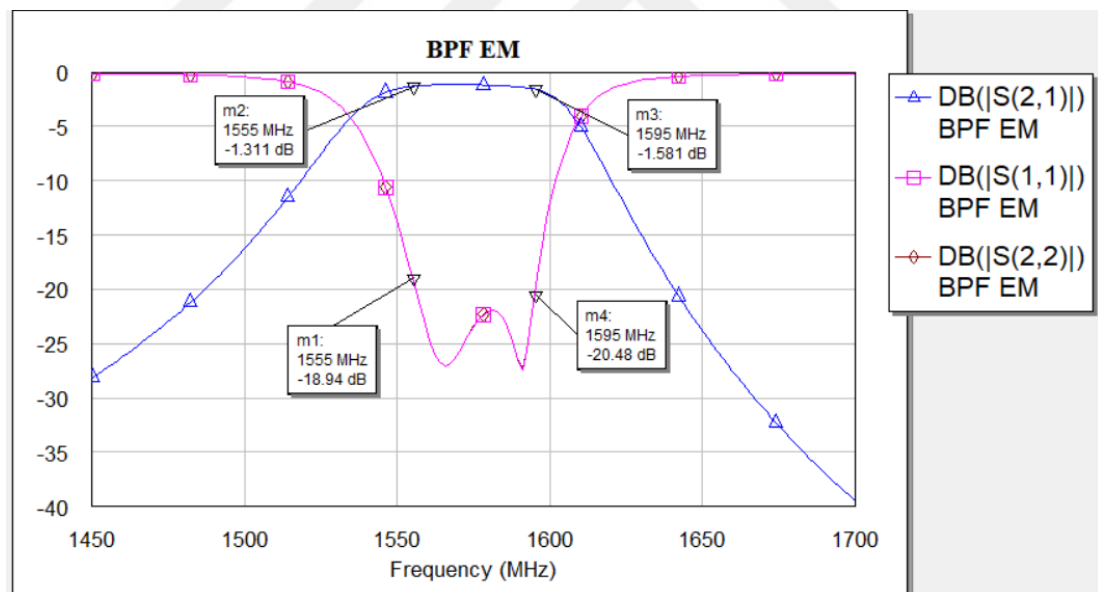


Figure 3.9 Bandpass filter simulation result between 1555 MHz - 1595 MHz using microstrip transmission lines

It is shown in the Figure 3.9 with the insertion loss less than 1.6 dB in the frequency range. Additionally, the input and output return loss values are higher than the design requirements of 10 dB in the frequency range.

3.4 2400 MHz - 2500 MHz Bandpass Filter Design

Microstrip transmission lines were used to design the bandpass filter for this frequency range. It was decided that the design should be an interdigital bandpass filter based on the bandwidth and the amount of space it takes up on the PCB.

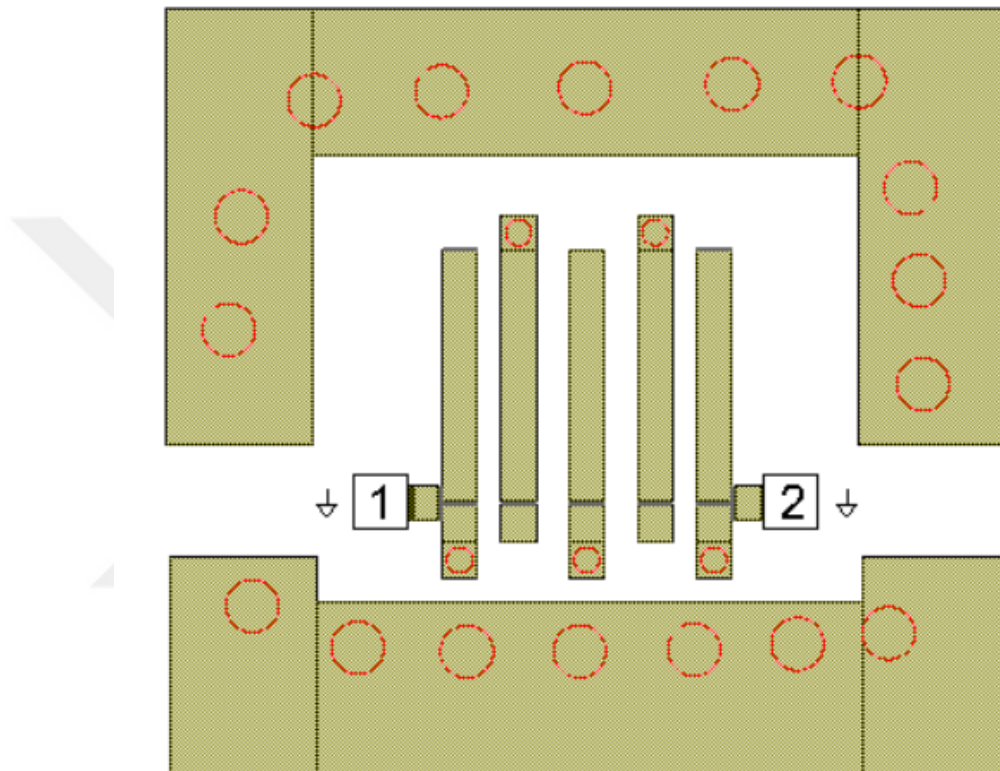


Figure 3.10 Bandpass filter between 2400 MHz - 2500 MHz designed using microstrip transmission lines

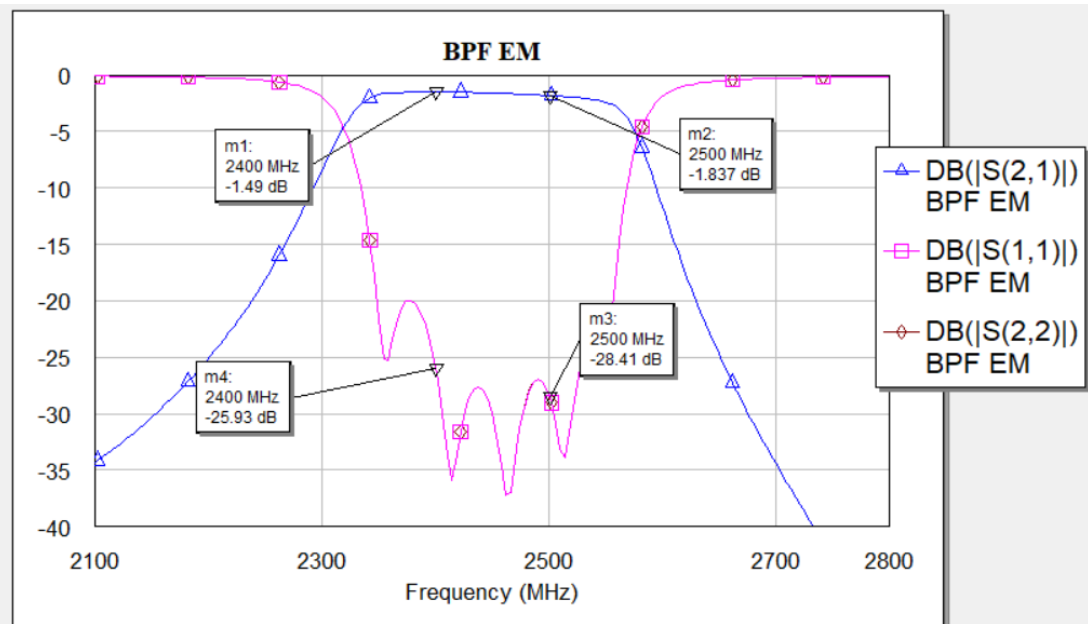


Figure 3.11 Bandpass filter simulation result between 2400 MHz - 2500 MHz using microstrip transmission lines

It is shown in the Figure 3.11 with the insertion loss less than 2 dB in the frequency range. Moreover, the input and output return loss values are much higher than the design requirements of 10 dB in the frequency range.

3.5 5725 MHz - 5875 MHz Bandpass Filter Design

As the frequency increases, the size of the bandpass filters designed with microstrip transmission lines decreases. Thus, the bandpass filter designed for this frequency range can be more simply placed on the PCB. The interdigital bandpass filter is designed for this frequency range because of both the optimum PCB space it occupies and the frequency range's narrowness.

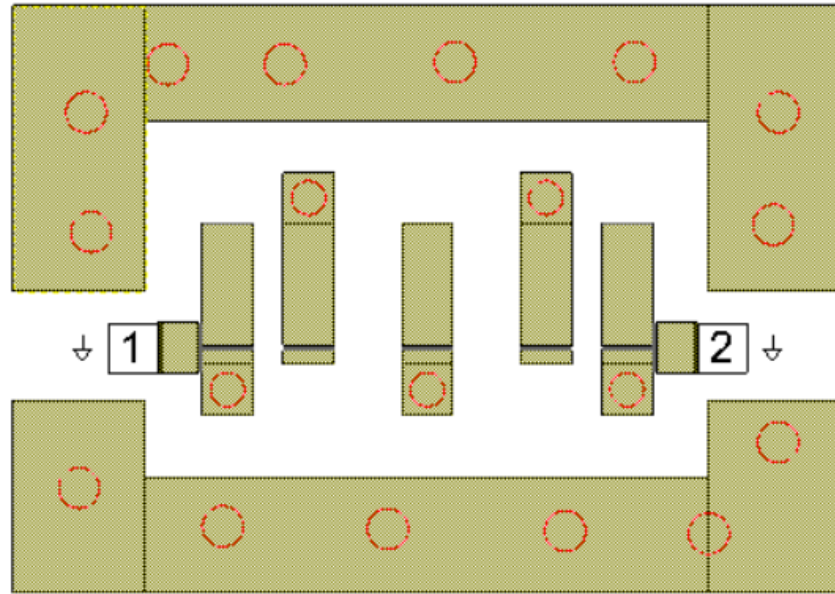


Figure 3.12 Bandpass filter between 5725 MHz - 5875 MHz designed using microstrip transmission lines

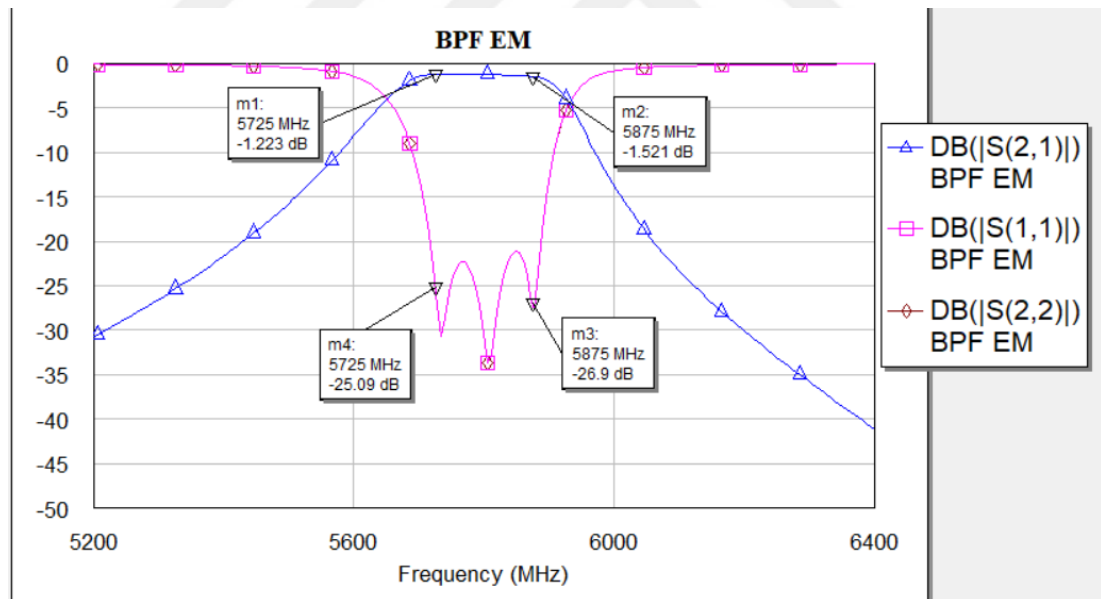


Figure 3.13 Bandpass filter simulation result between 5725 MHz - 5875 MHz using microstrip transmission lines

Figure 3.13 shows EM simulation results having an insertion loss in the frequency range of less than 1.6 dB. In addition, the input and output return loss values are considerably greater than the frequency range design requirement of 10 dB.

3.6 6175 MHz - 7125 MHz Bandpass Filter Design

The edge coupled transmission line approach was determined to be the most effective design for the bandpass filter as a result of the simulations performed in this frequency range.

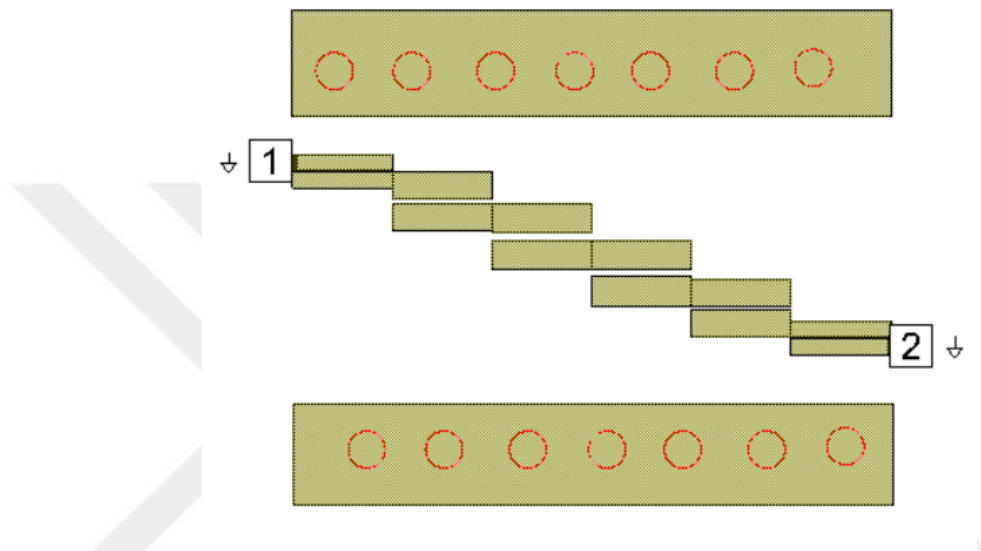


Figure 3.14 Bandpass filter between 6175 MHz - 7125 MHz designed using microstrip transmission lines

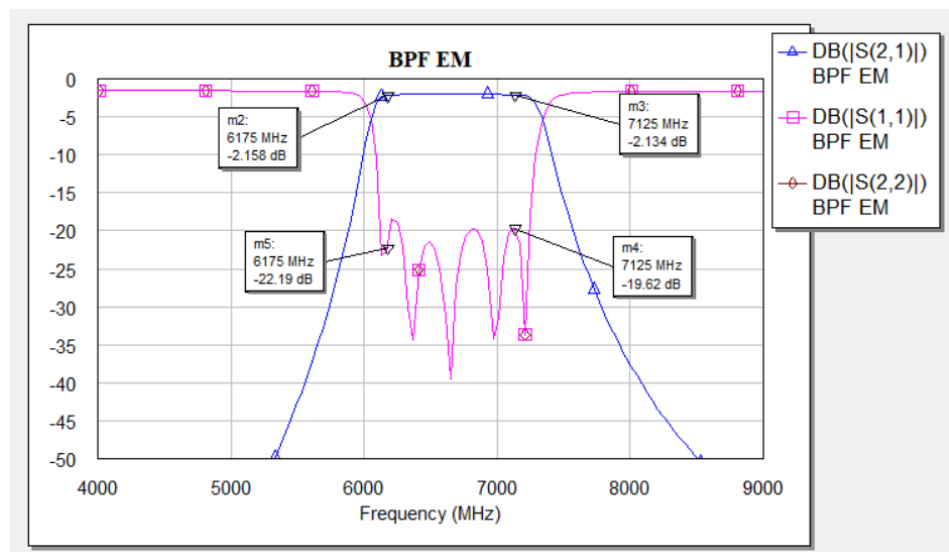


Figure 3.15 Bandpass filter simulation result between 6175 MHz - 7125 MHz using microstrip transmission lines

Figure 3.15 illustrate EM simulation results having an insertion loss in the frequency range of less than 2.5 dB. In addition, the input and output return loss values are considerably greater than the frequency range design requirement of 10 dB.



4. DESIGN STAGES AND SIMULATIONS OF LNA CIRCUITS AND RF SWITCH

Seven LNA circuit configurations for the frequency bands mentioned in Table 1.1 and RF switch circuit configuration are explained in this chapter. Suitable LNAs were chosen and separate LNA circuits were built for each frequency range. The design sequence is RF1 to RF6, according to their pseudonyms listed in the introduction. The broadband LNA circuit was analyzed after analyzing six LNA circuits at the inputs of the RF switch. The RF switch was matched to the LNA circuits after all of the LNA circuits had been designed.

4.1 6175 MHz - 7125 MHz LNA Circuit Design

An unconditionally stable low noise amplifier is used as the LNA in the circuit design. Because it satisfies both the pricing and performance requirements of the project.

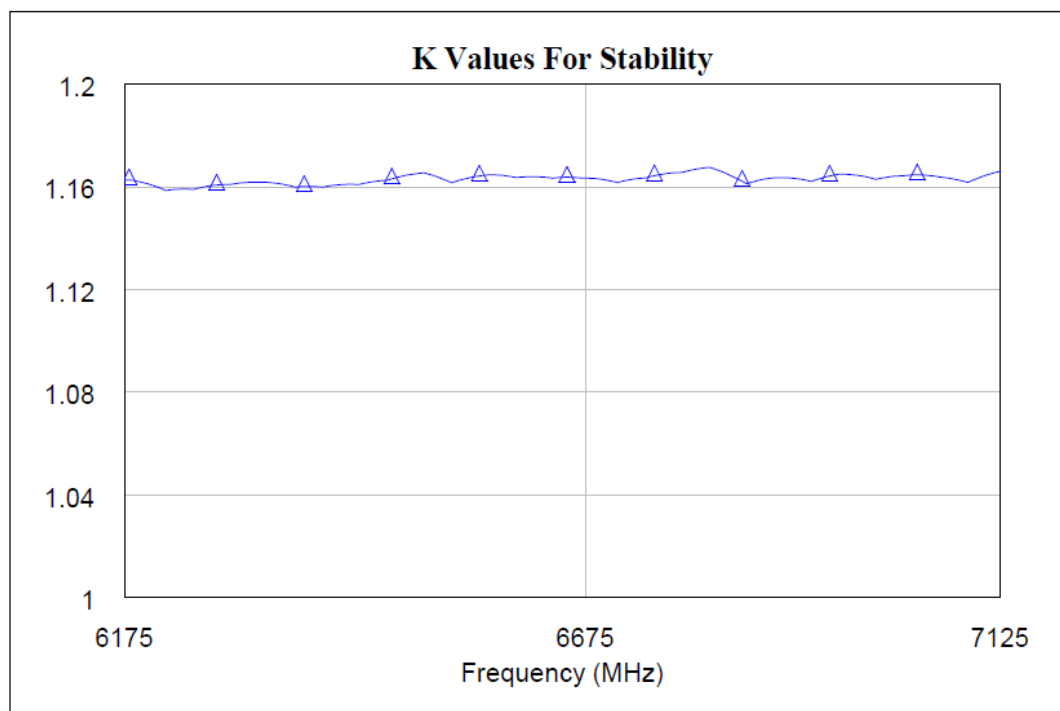


Figure 4.1 LNA stability response from 6175 MHz to 7125 MHz

Figure 4.1 shows that the K value is greater than 1 in the frequency range. This proves that the LNA is unconditionally stable in this circumstance.

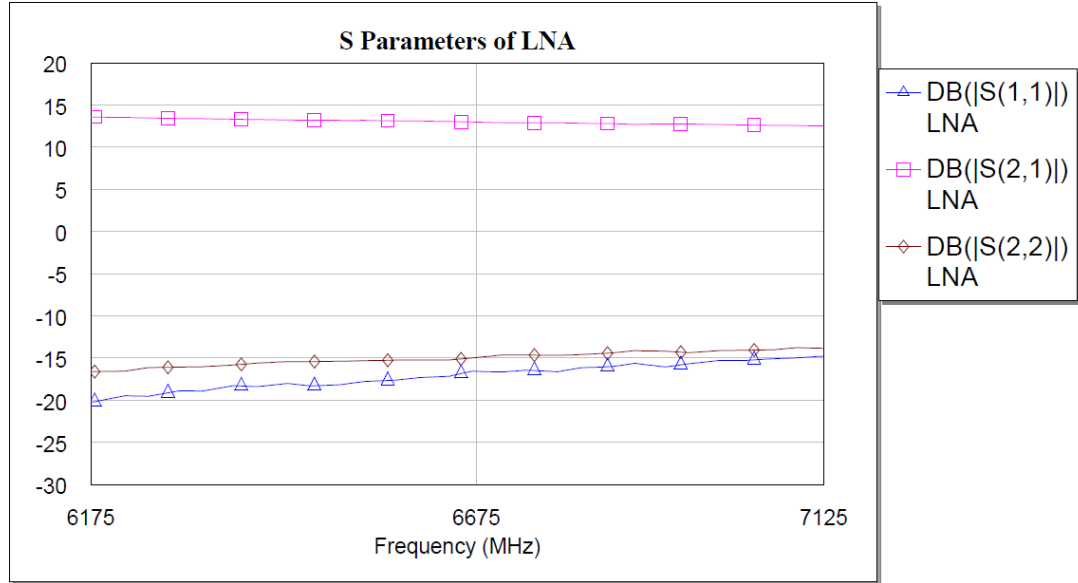


Figure 4.2 LNA S parameters from 6175 MHz to 7125 MHz

Although the LNA's S parameter values indicate that nearly no matching is required, the LNA's small footprint area demonstrates the high impedance transmission line effect at the input and output. Therefore, input matching circuit and output matching circuit are designed on both sides of the LNA.

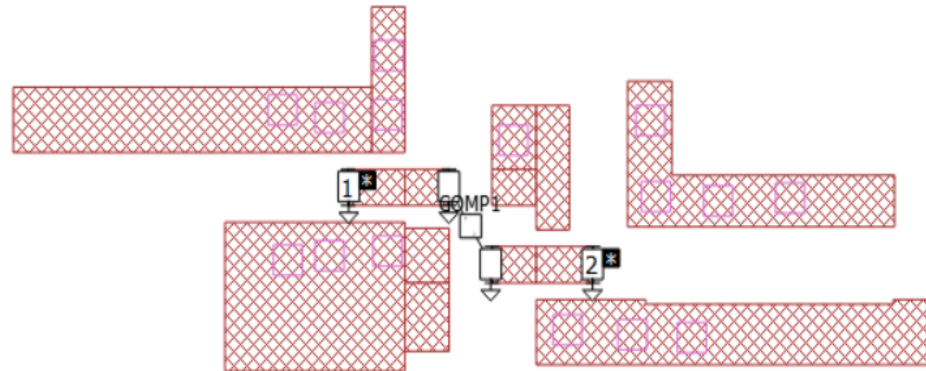


Figure 4.3 LNA footprint layout between 6175 MHz - 7125 MHz

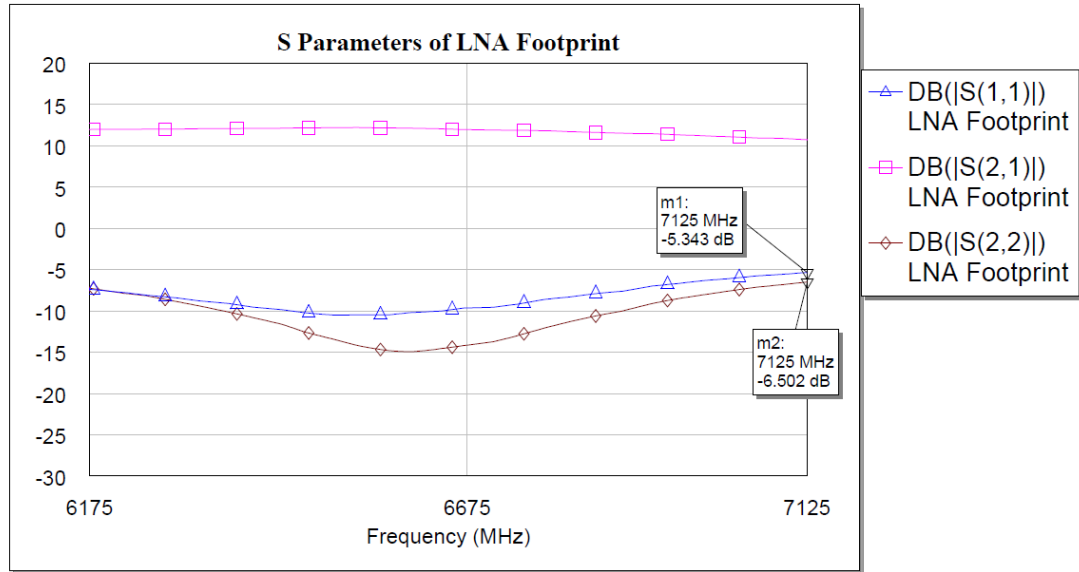


Figure 4.4 LNA footprint simulation result between 6175 MHz - 7125 MHz

It is obvious that the values of S11 and S22 will exceed -6 and -7 dB, respectively, if the matching circuits are not designed. Thus, it was necessary to design matching circuits.

Since the matching process for the input and output matching circuits will be broadband, 3-order ideal values are utilized.

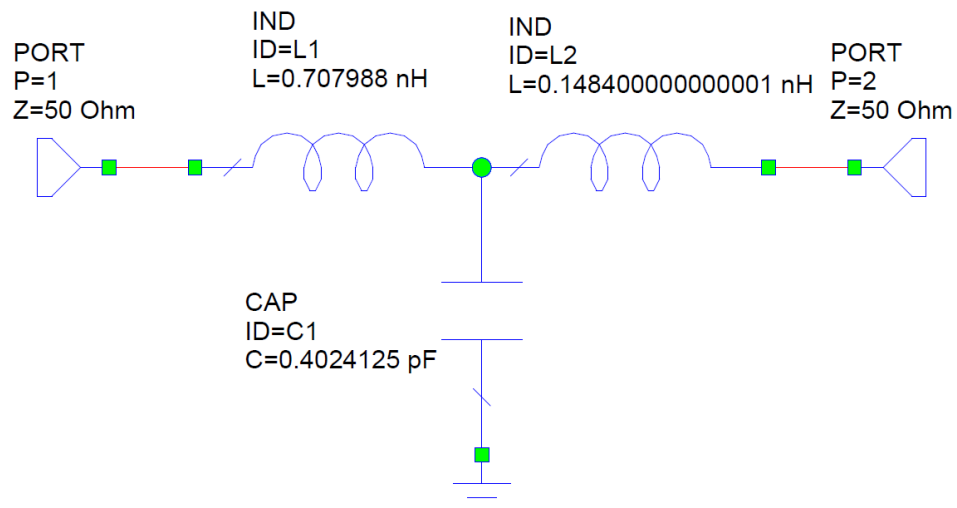


Figure 4.5 LNA input matching network (IMN) between 6175 MHz - 7125 MHz using ideal elements

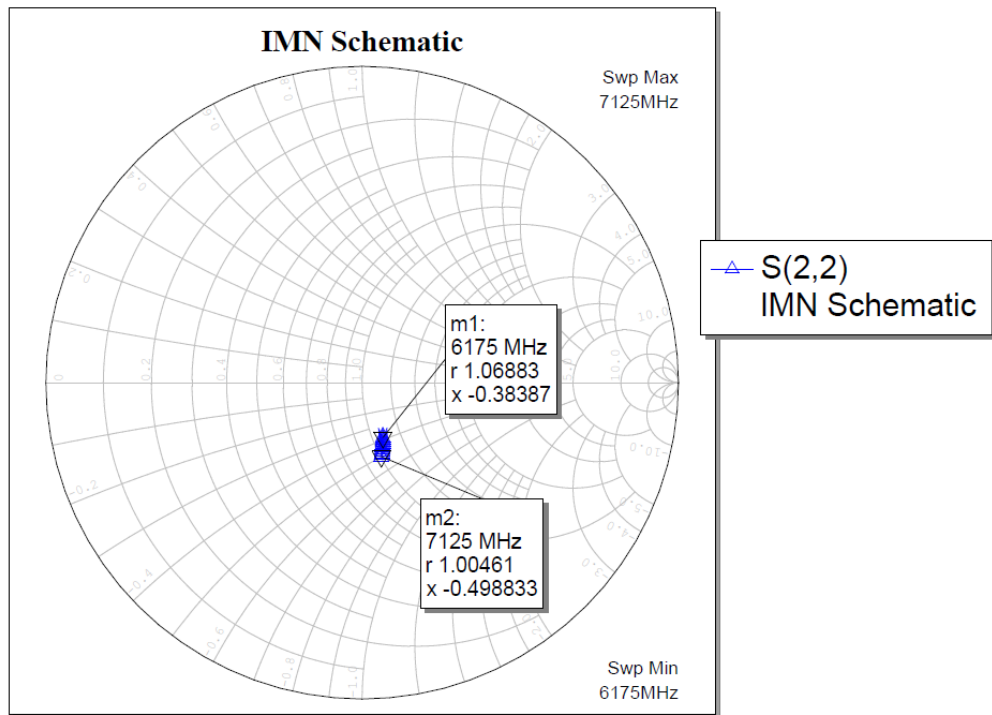


Figure 4.6 LNA input matching network simulation result between 6175 MHz - 7125 MHz using ideal elements

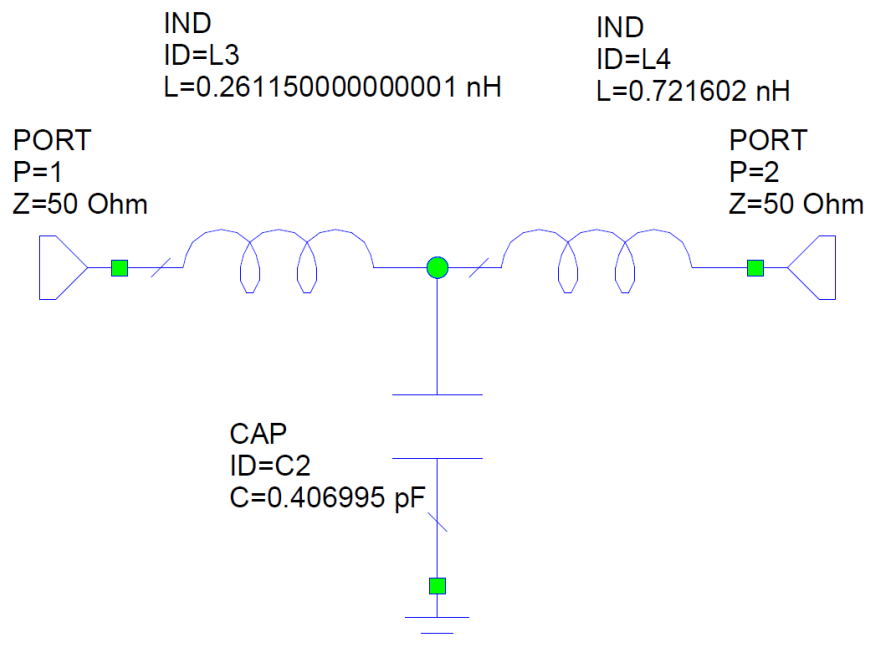


Figure 4.7 LNA output matching network (OMN) between 6175 MHz - 7125 MHz using ideal elements

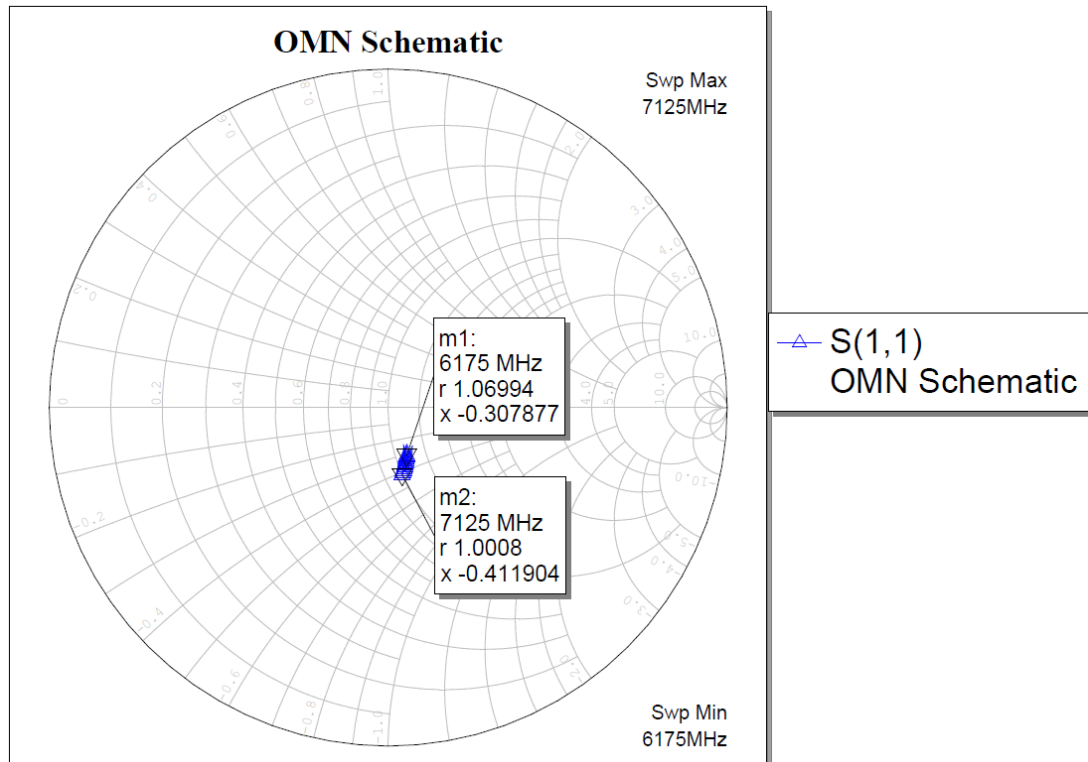


Figure 4.8 LNA output matching network simulation result between 6175 MHz - 7125 MHz using ideal elements

The S parameters reached the desired level as a result of the ideal element-based matching procedure.

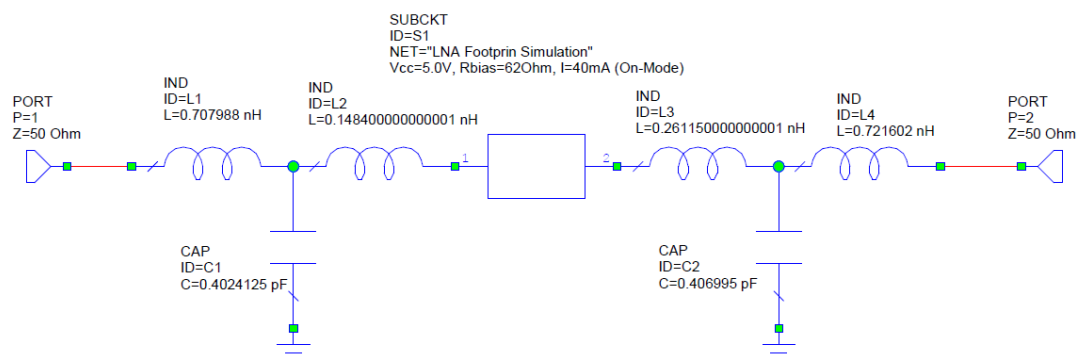


Figure 4.9 Matching networks usign ideal elements between 6175 MHz - 7125 MHz LNA

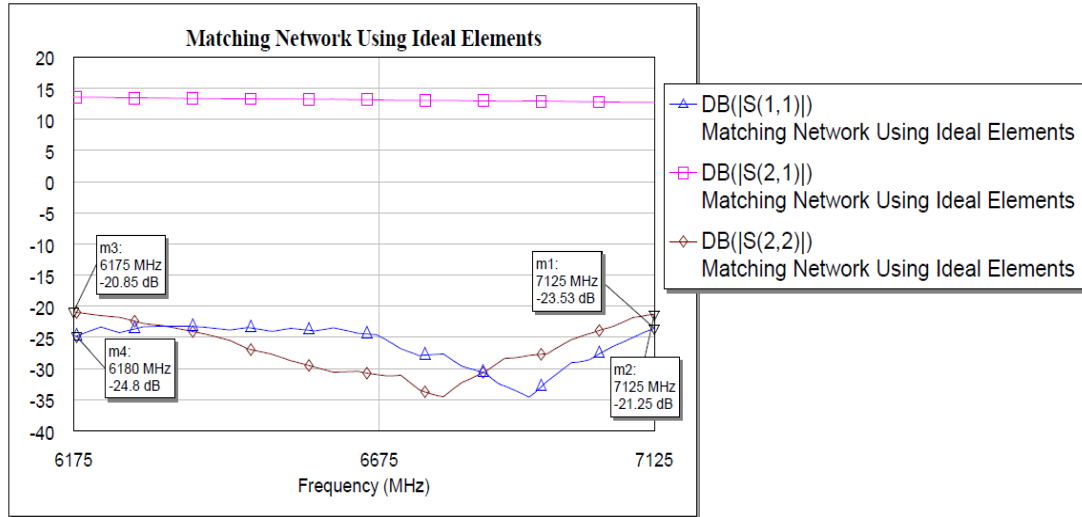


Figure 4.10 LNA matching networks simulation result using ideal elements between 6175 MHz - 7125 MHz

Matching networks were designed by converting ideal elements to microstrip transmission lines after matching with ideal elements.

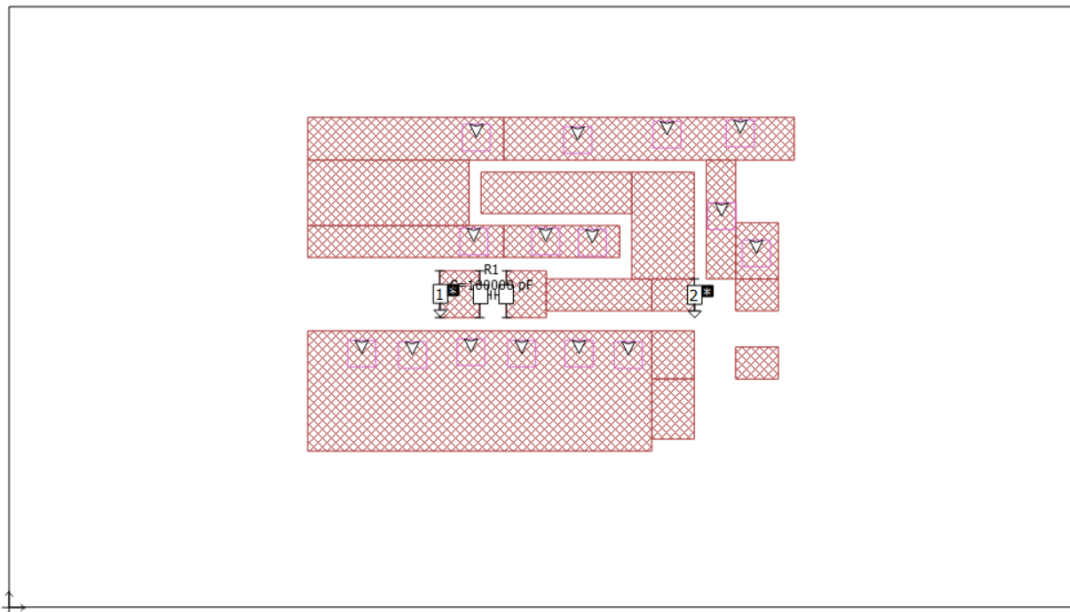


Figure 4.11 LNA input matching network using microstrip transmission lines between 6175 MHz - 7125 MHz

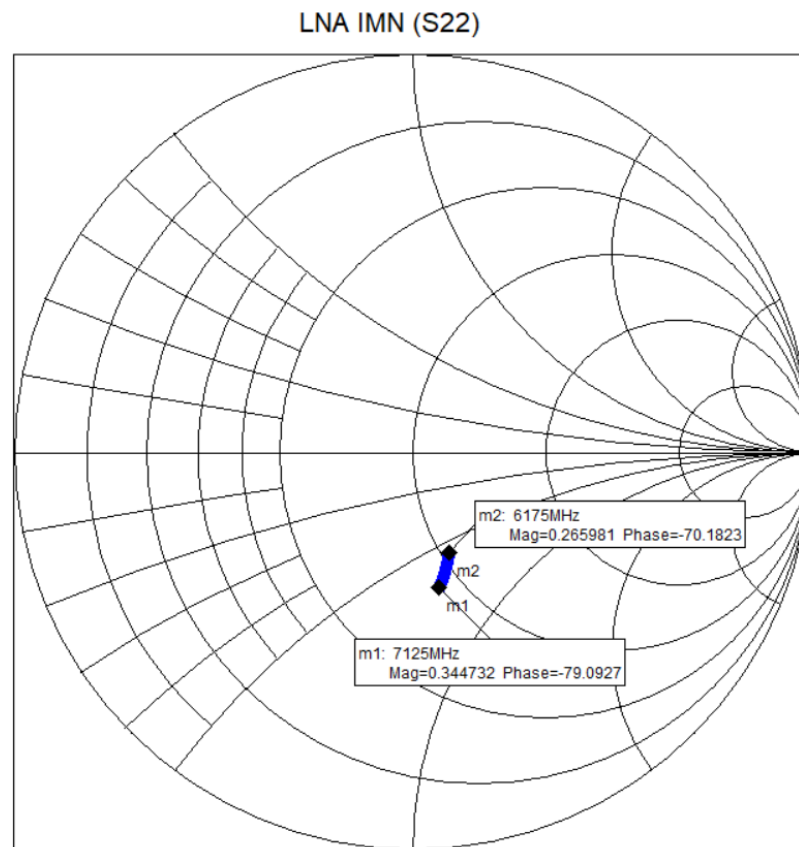


Figure 4.12 LNA input matching network simulation result using microstrip transmission lines between 6175 MHz - 7125 MHz

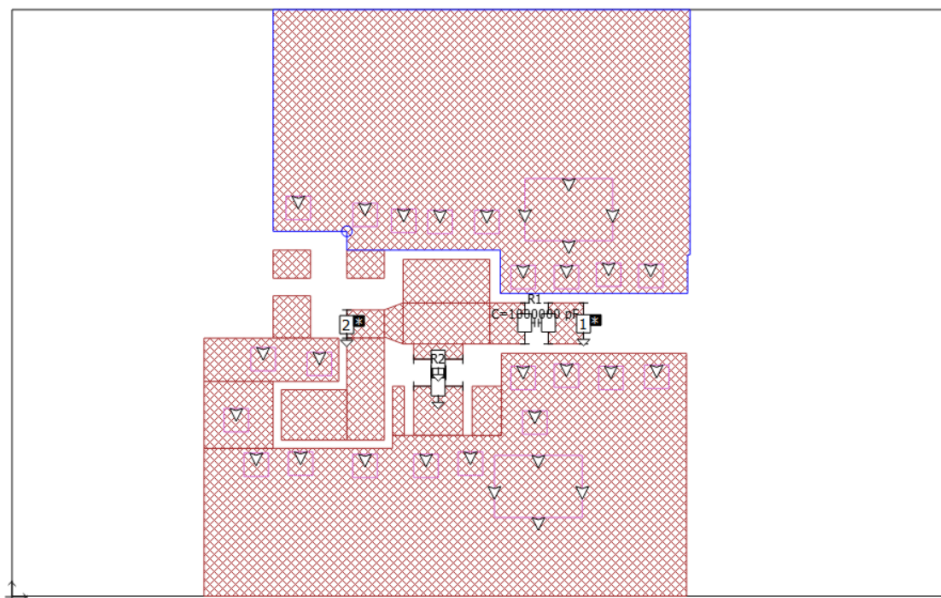


Figure 4.13 LNA output matching network using microstrip transmission lines between 6175 MHz - 7125 MHz

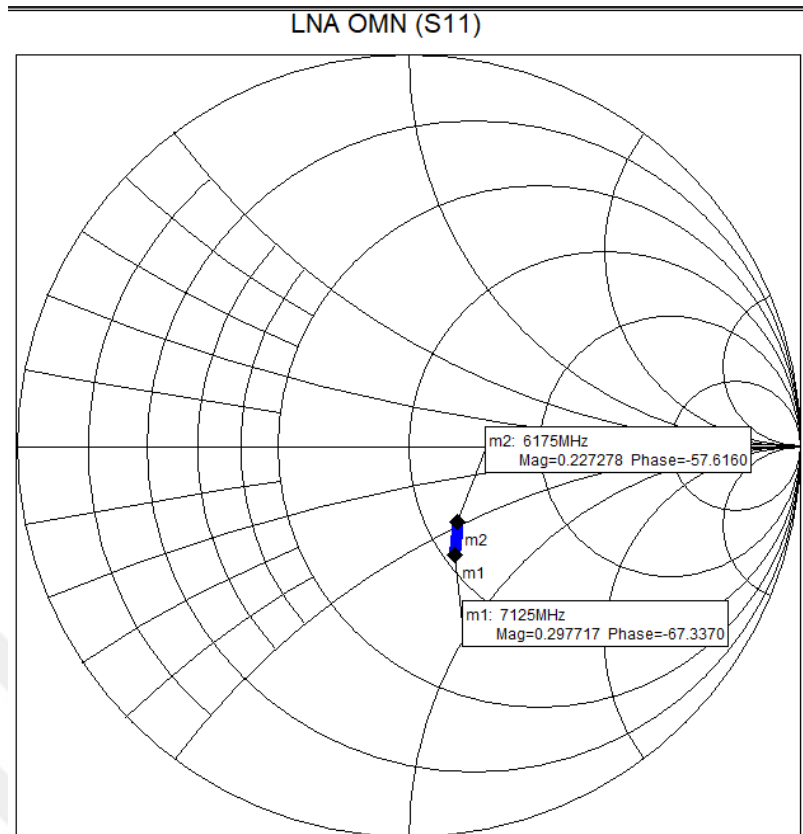


Figure 4.14 LNA output matching network simulation result using microstrip transmission lines between 6175 MHz - 7125 MHz

The integrated state of the LNA with the designed matching circuits and the result are as follows:

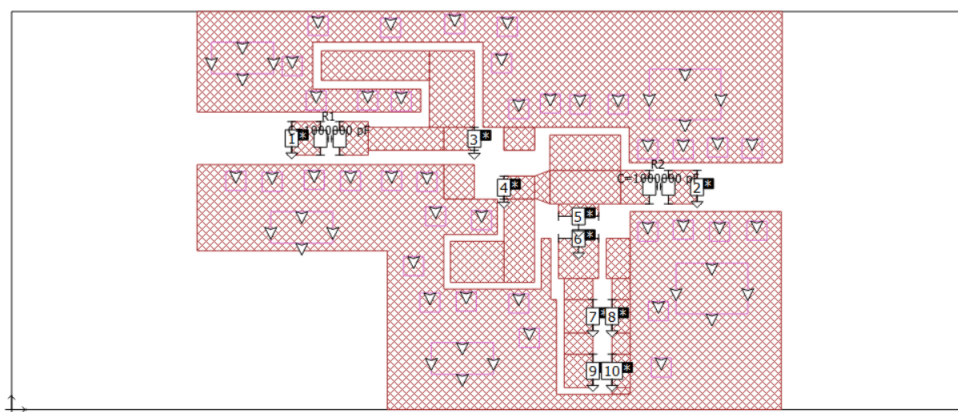


Figure 4.15 LNA circuit between 6175 MHz - 7125 MHz

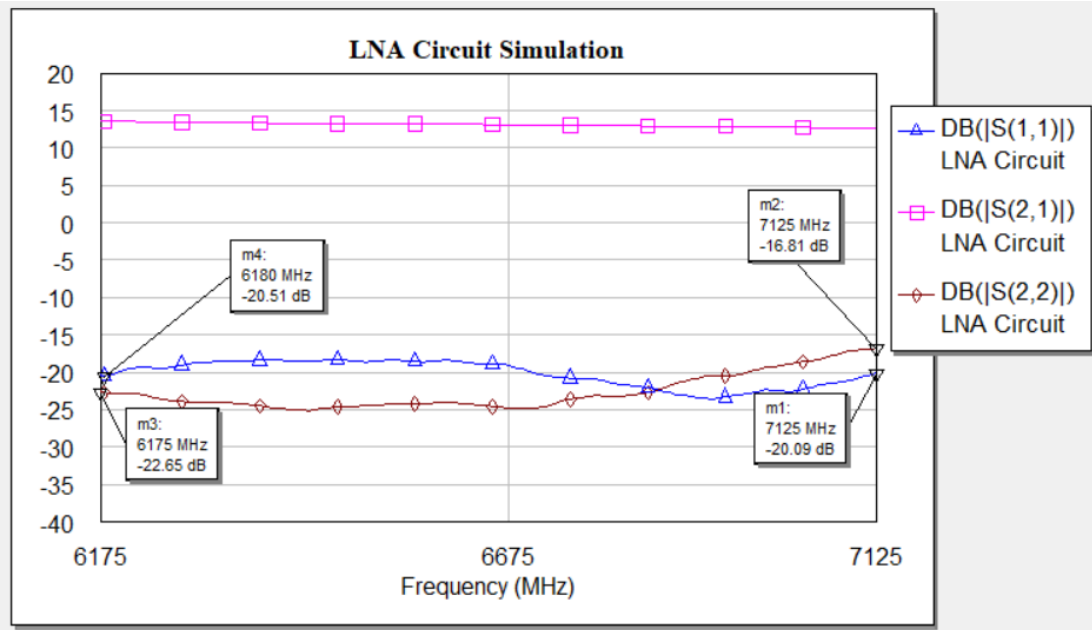


Figure 4.16 LNA circuit simulation result between 6175 MHz - 7125 MHz

The matching process with microstrip transmission lines has been designed at the desired level, as can be seen in Figure 4.16.

4.2 5725 MHz - 5875 MHz LNA Circuit Design

An unconditionally stable low noise amplifier is preferred in the LNA circuit designed for frequencies between 5725 MHz and 5875 MHz. This amplifier is chosen due to its exceptionally low noise figure. The noise figure of the amplifier varies between 0.5 dB and 0.8 dB in the frequency range.

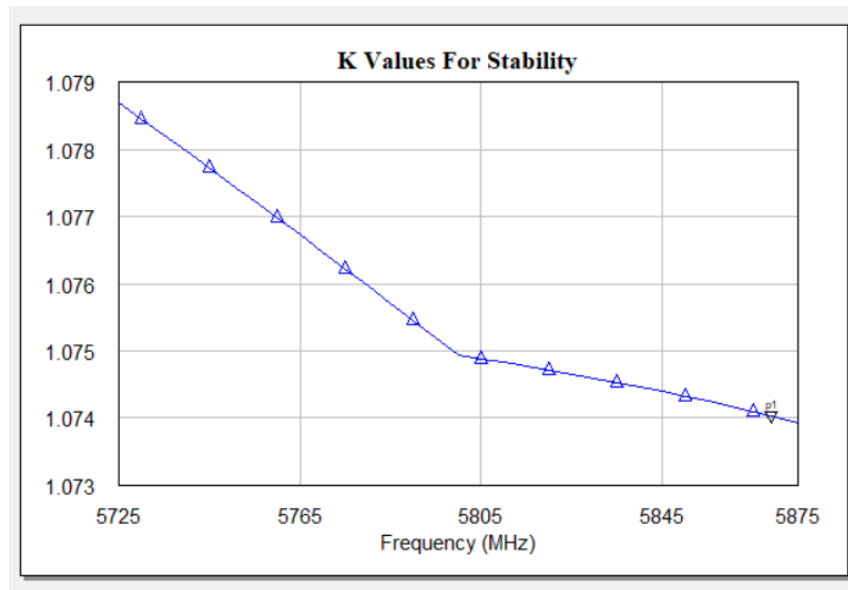


Figure 4.17 LNA stability response from 5725 MHz to 5875 MHz

Figure 4.17 illustrates that throughout the frequency range, K is higher than 1. This indicates that the LNA is unconditionally stable.

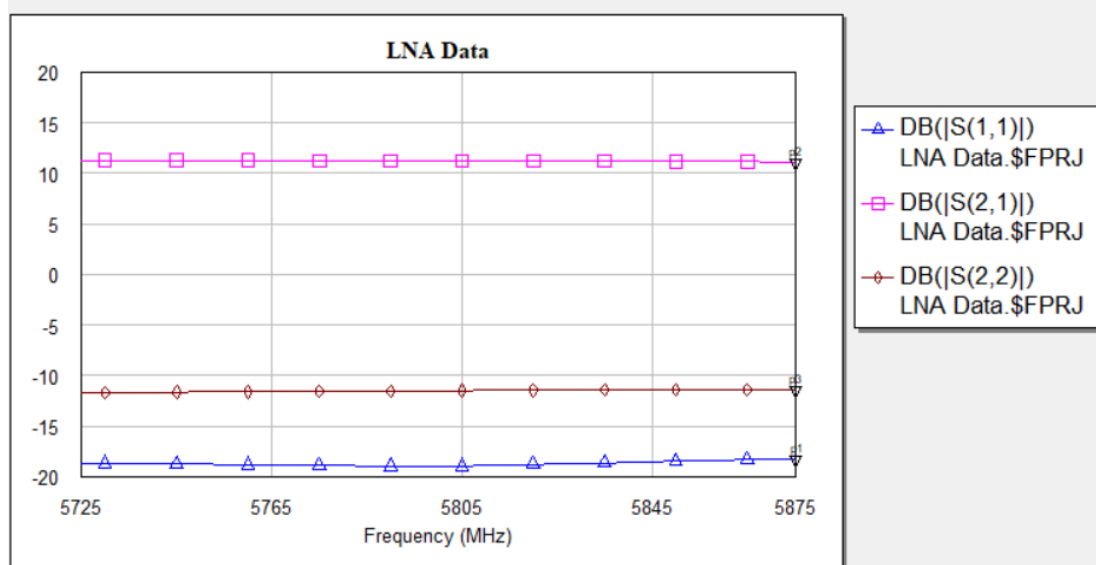


Figure 4.18 LNA S parameters from 5725 MHz to 5875 MHz

The LNA's narrow footprint area shows the high impedance transmission line effect at the input and output even if the S parameter values show that almost no matching is necessary. As a result, input matching and output matching circuits are used to build the LNA's both sides.

The layout image and simulation result are as follows when the LNA is placed in the layout under the appropriate conditions::

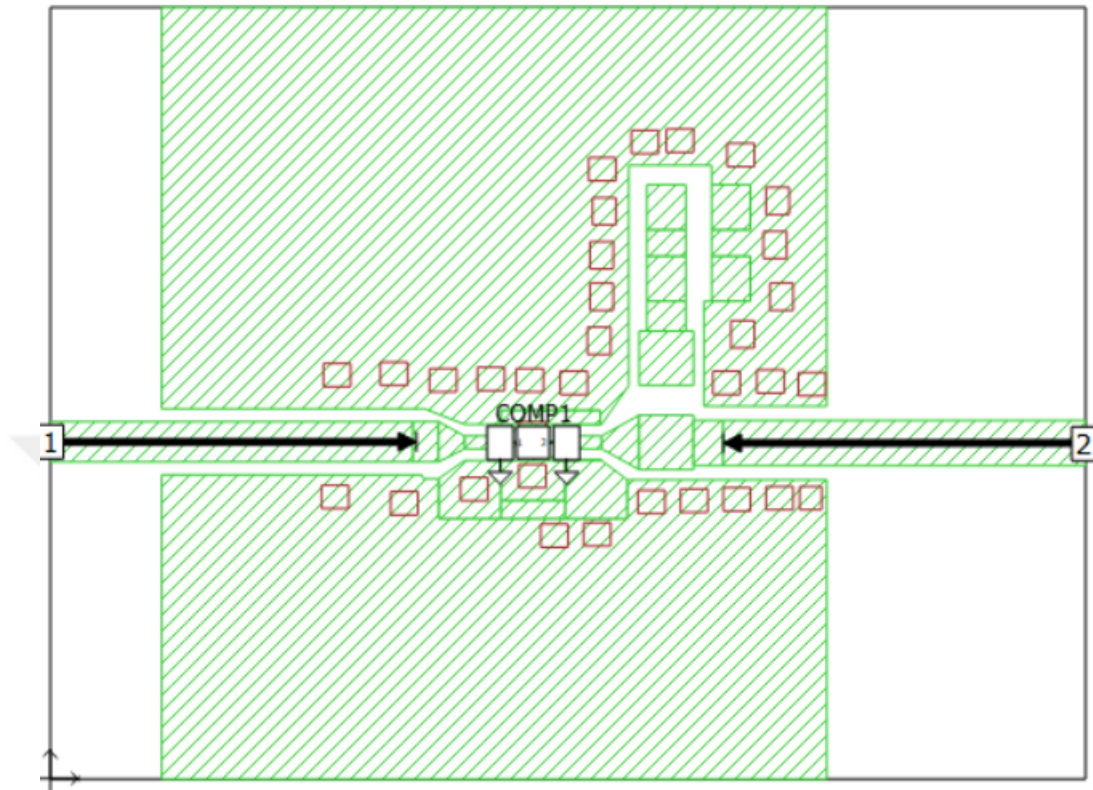


Figure 4.19 LNA footprint layout between 5725 MHz - 5875 MHz

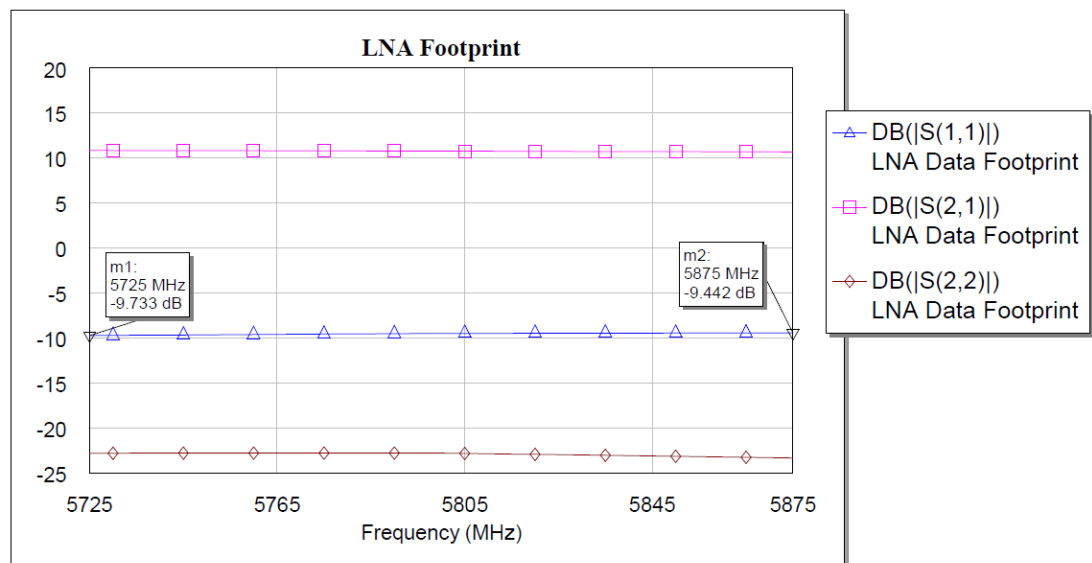


Figure 4.20 LNA footprint simulation result

As seen in the Figure 4.20, the S11 value has exceeded -10 dB. Hence, the input matching circuit has been designed and implemented to correct this.

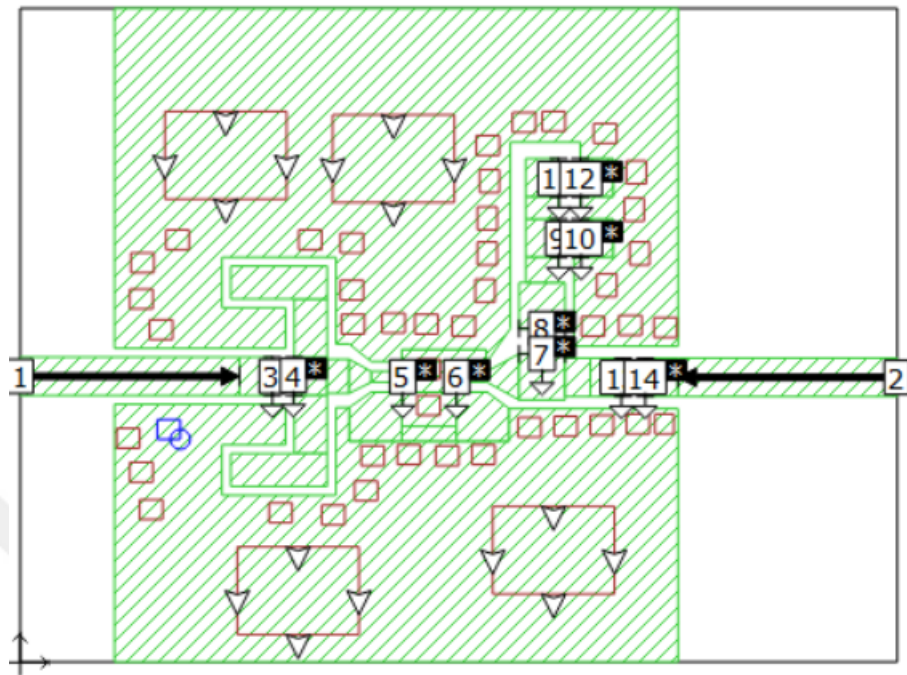


Figure 4.21 5725 MHz - 5875 MHz LNA circuit

In Figure 4.22, it is shown that the S11 value is better after creating the input matching network.

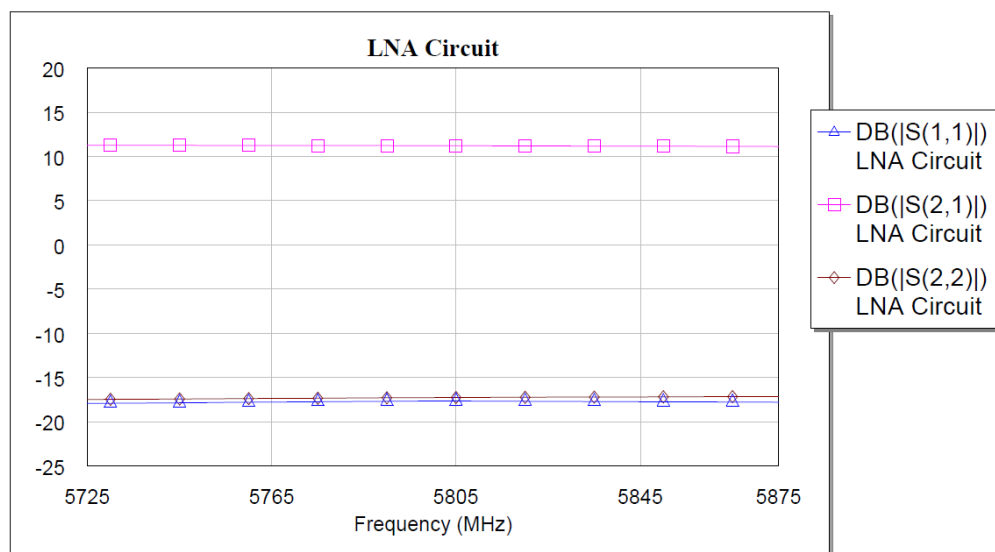


Figure 4.22 LNA circuit simulation result between 5725 MHz - 5875 MHz

4.3 2400 MHz - 2500 MHz LNA Circuit Design

As with other circuits, an unconditionally stable low noise amplifier is preferred in the LNA circuit designed for frequencies between 2400 MHz and 2500 MHz.

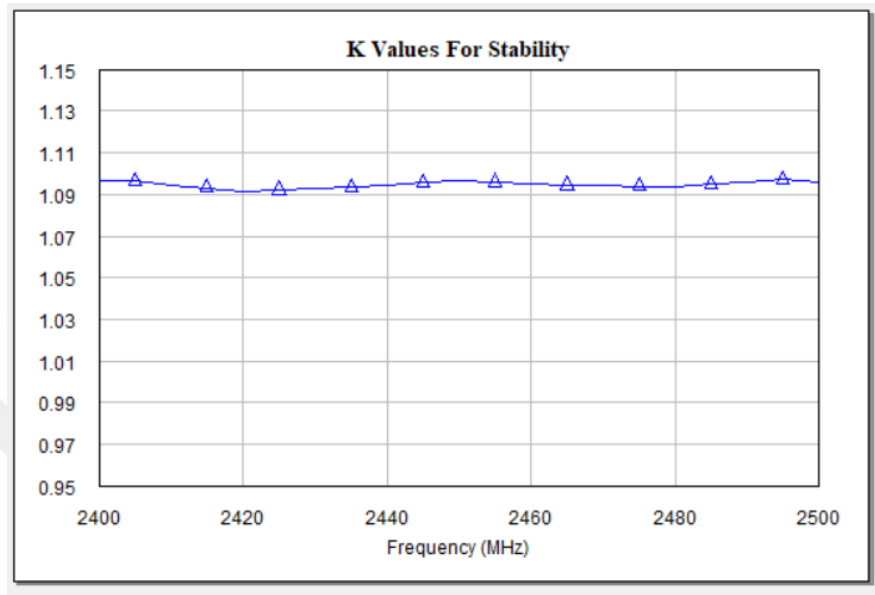


Figure 4.23 LNA stability response from 2400 MHz to 2500 MHz

Figure 4.23 illustrates that the K value is greater than 1 in the frequency range. This proves that the LNA is unconditionally stable in this circumstance.

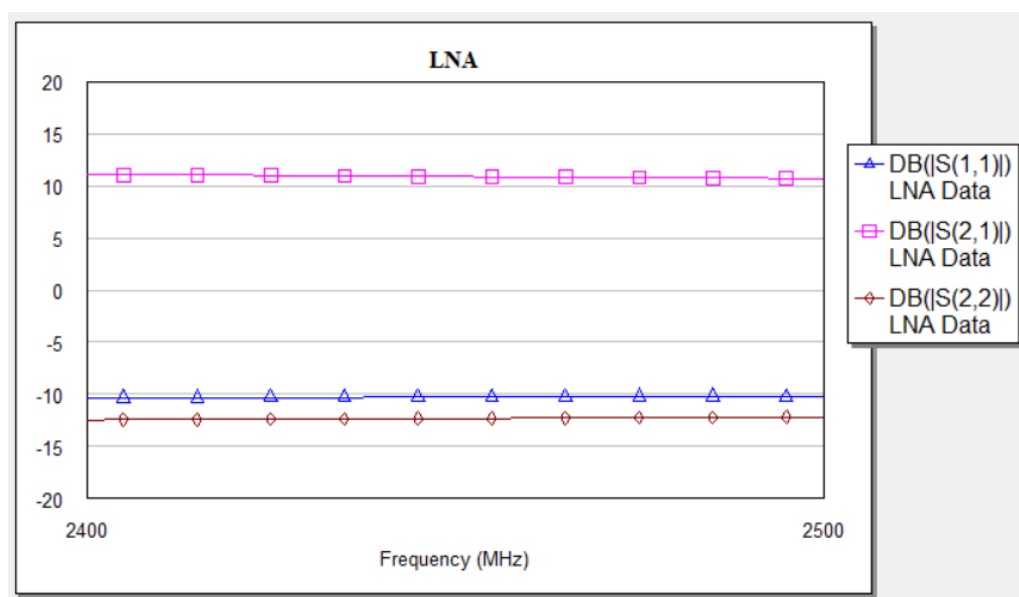


Figure 4.24 LNA S parameters from 2400 MHz to 2500 MHz

SMD components were matched to improve the frequency range performance and further decrease the S11 value at the input section of the LNA. A shunt-connected 0.5 pF capacitor and then a series-connected 1.5 nH inductor is used before the capacitor used for DC current blocking in the input section of the LNA circuit. Modelithic values of SMD components are utilized in the simulation.

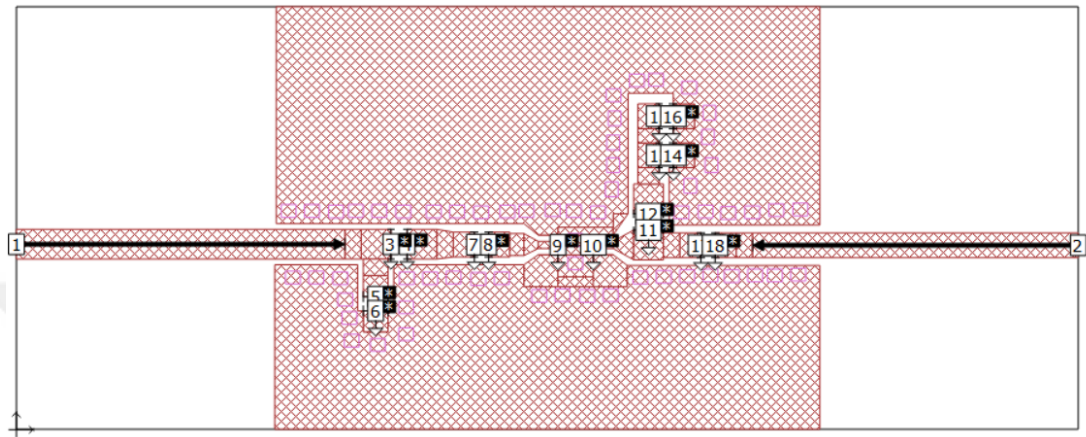


Figure 4.25 LNA circuit layout between 2400 MHz - 2500 MHz

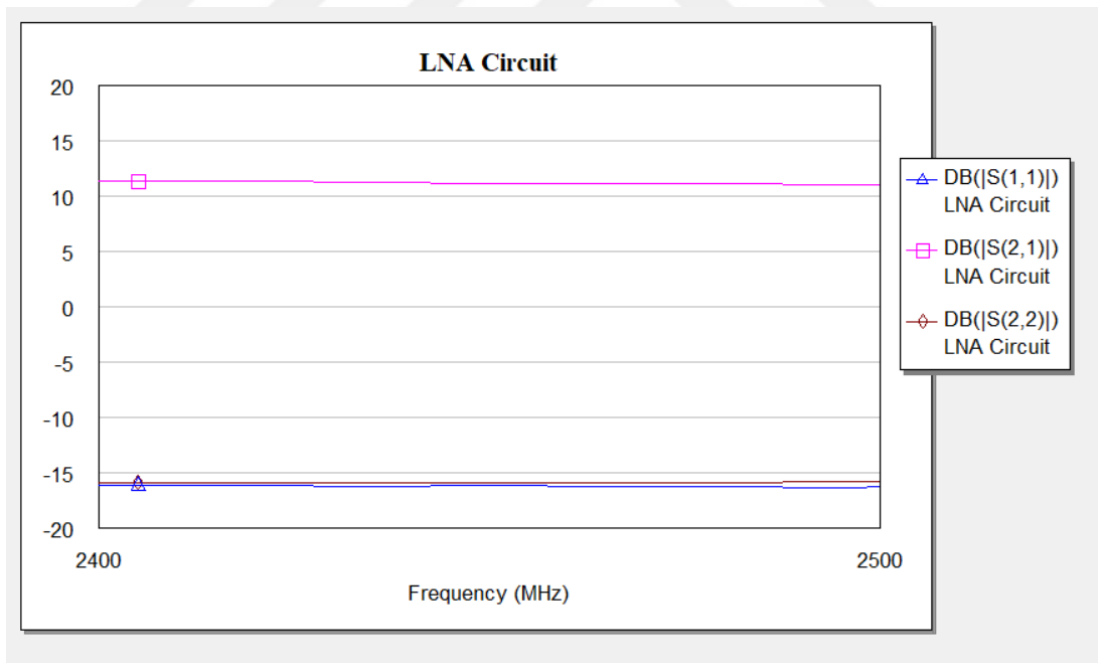


Figure 4.26 LNA circuit simulation result between 2400 MHz - 2500 MHz

The desired level of return loss values has been reached after the matching process with SMD components.

4.4 1555 MHz - 1595 MHz LNA Circuit Design

As with other circuits, an unconditionally stable low noise amplifier is preferred in the LNA circuit designed for frequencies between 1555 MHz and 1590 MHz. In addition to being stable, this amplifier offers good return losses in the frequency range, although the S11 value might be improved.

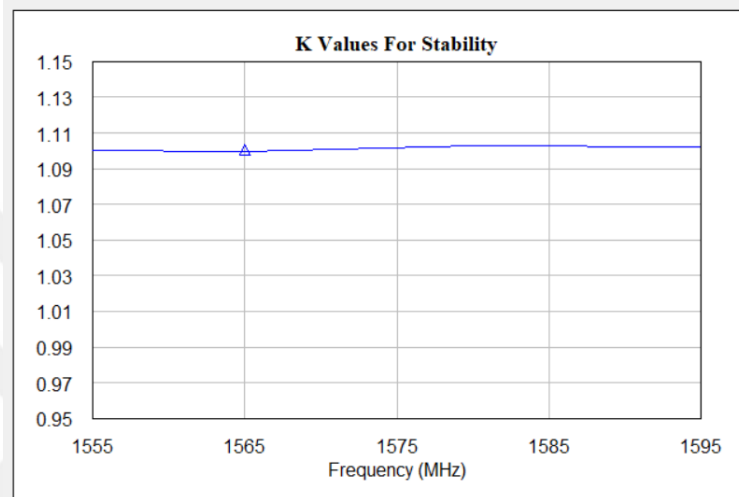


Figure 4.27 LNA stability response from 1555 MHz to 1595 MHz

The K value is greater than 1 in the frequency range, as shown in Figure 4.27. This demonstrates that the LNA is unconditionally stable, like the other LNAs.

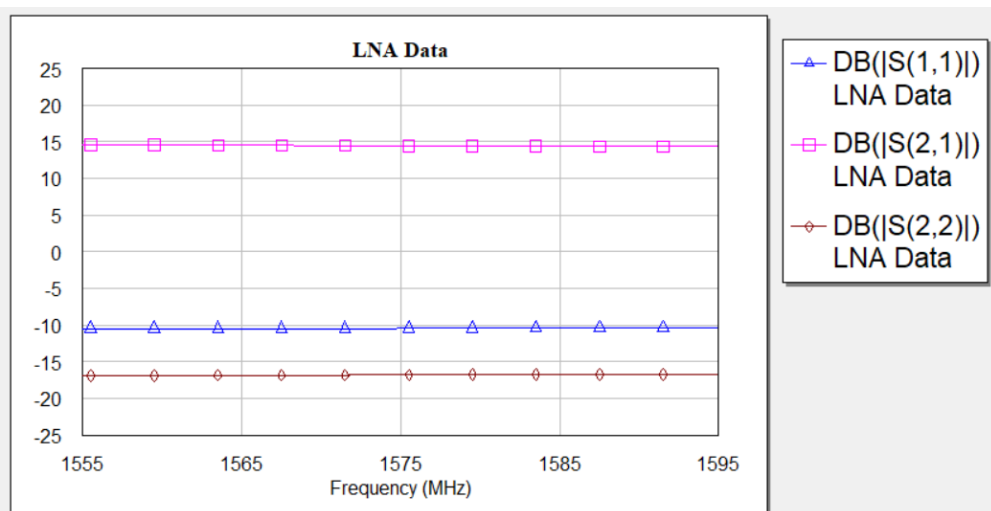


Figure 4.28 LNA S parameters from 1555 MHz to 1595 MHz

It was necessary to shunt an 4.7 nH inductor to the ground before the DC blocking capacitor at the LNA's input to further minimize the simulation's S11 value.

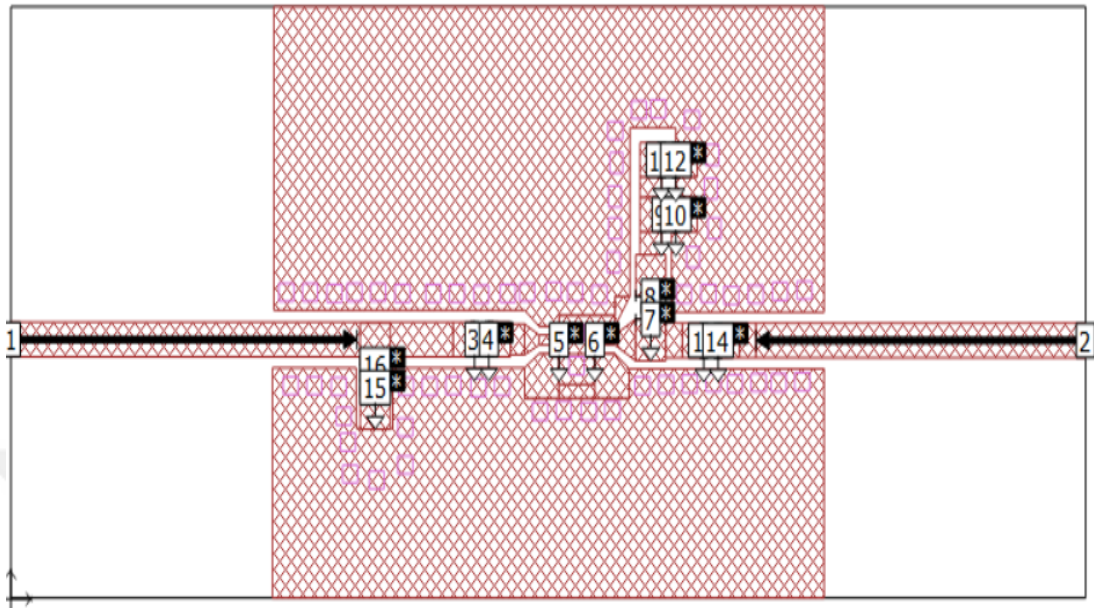


Figure 4.29 LNA circuit layout between 1555 MHz - 1595 MHz

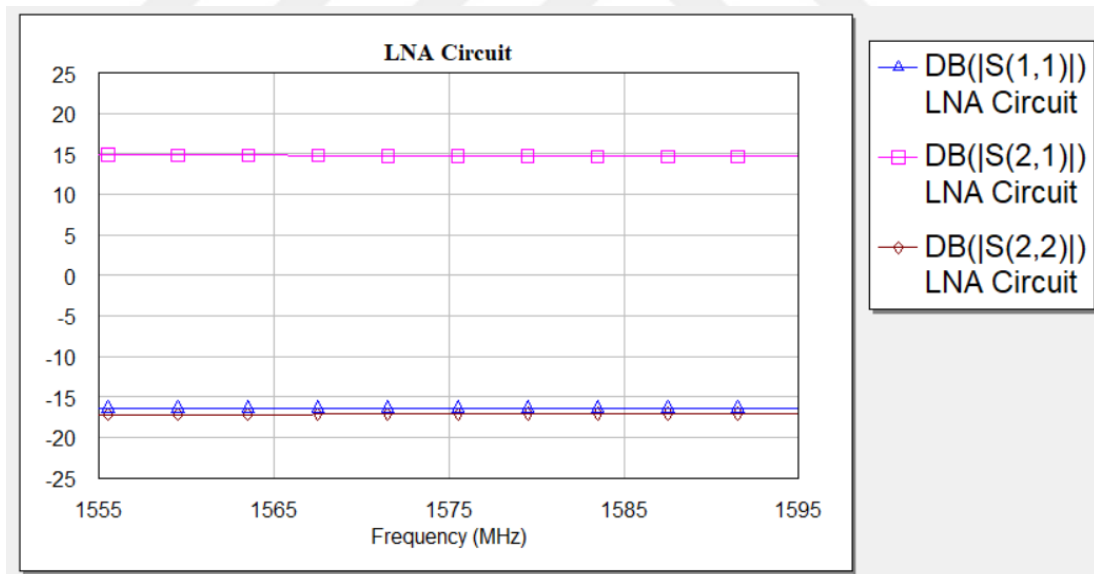


Figure 4.30 LNA circuit simulation result between 1555 MHz - 1595 MHz

The desired S11 value was achieved after the shunt 4.7 nH value was applied.

4.5 1170 MHz - 1278 MHz LNA Circuit Design

As with other circuits, an unconditionally stable low noise amplifier is preferred in the LNA circuit designed for frequencies between 1170 MHz and 1278 MHz. This amplifier is stable and has good return losses in the frequency range, but the S11 value can be improved.

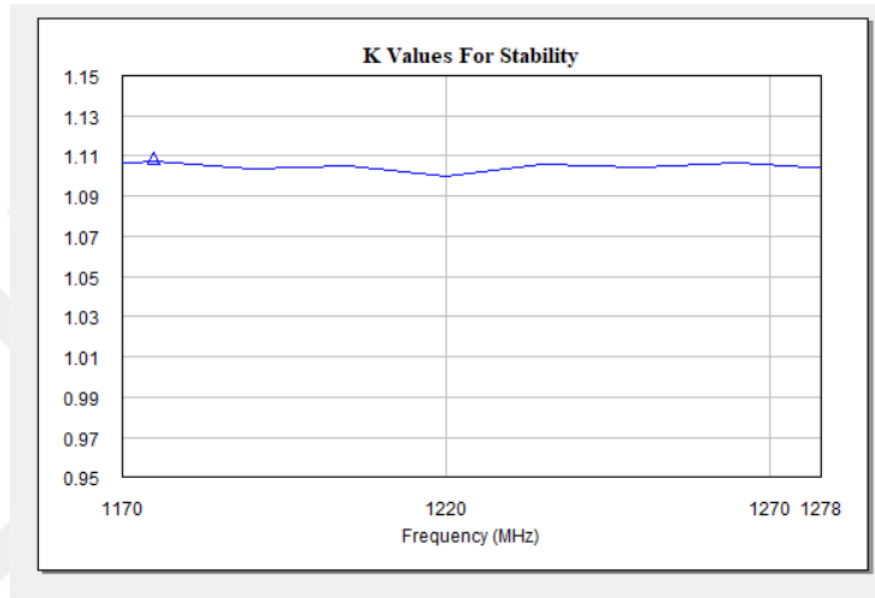


Figure 4.31 LNA stability response from 1170 MHz to 1278 MHz

The K value is greater than 1 in the frequency range, as shown in Figure 4.31. This demonstrates that the LNA is unconditionally stable, like the other LNAs.

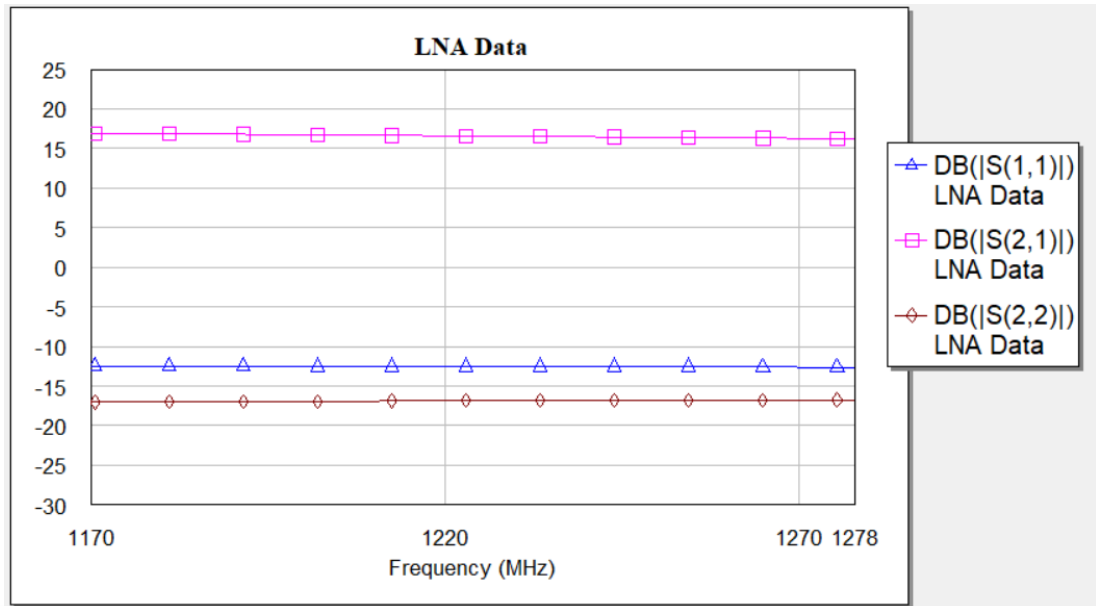


Figure 4.32 LNA S parameters from 1170 MHz to 1278 MHz

It was necessary to add an 0.8 pF capacitor as a shunt to the ground before the DC blocking capacitor at the input section of the LNA to improve the S11 value in simulation. Additionally, a 1 nH inductor was put in series to the transmission line later on. The series inductor is placed as the first component on the circuit's input side.

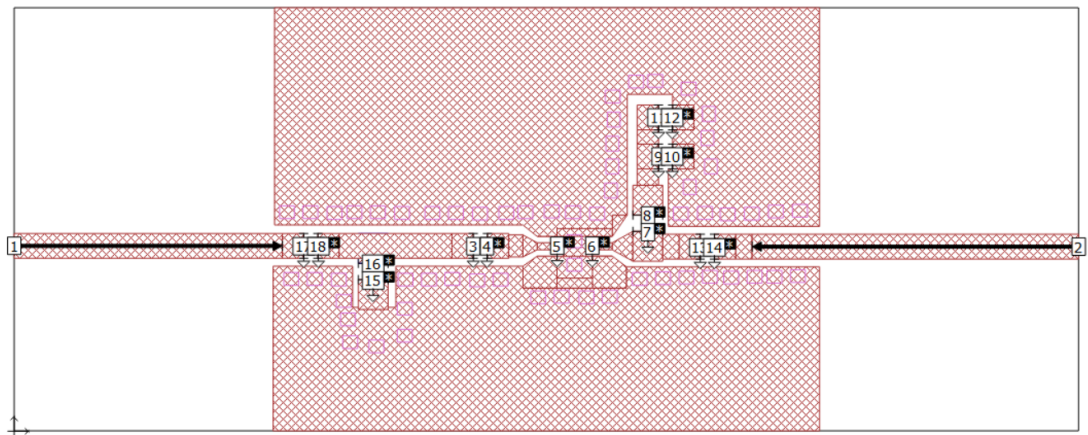


Figure 4.33 LNA circuit layout between 1170 MHz - 1278 MHz

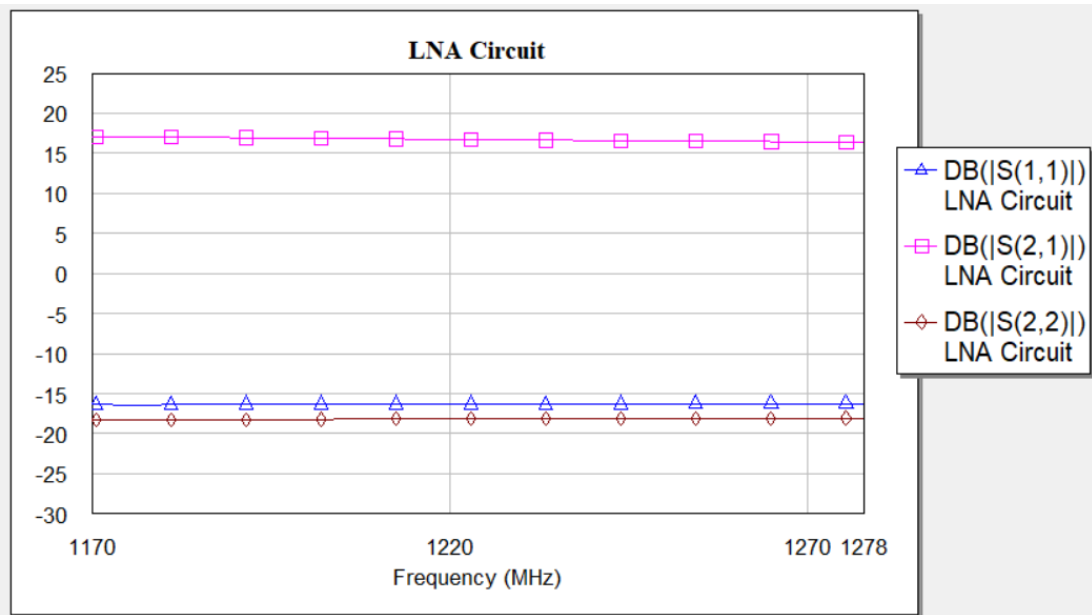


Figure 4.34 LNA circuit simulation result between 1170 MHz - 1278 MHz

Results were obtained in the simulation with return losses better than 15 dB after the matching processes were applied.

4.6 902 MHz – 928 MHz LNA Circuit Design

As with other circuits, an unconditionally stable low noise amplifier is preferred in the LNA circuit designed for frequencies between 902 MHz and 928 MHz.

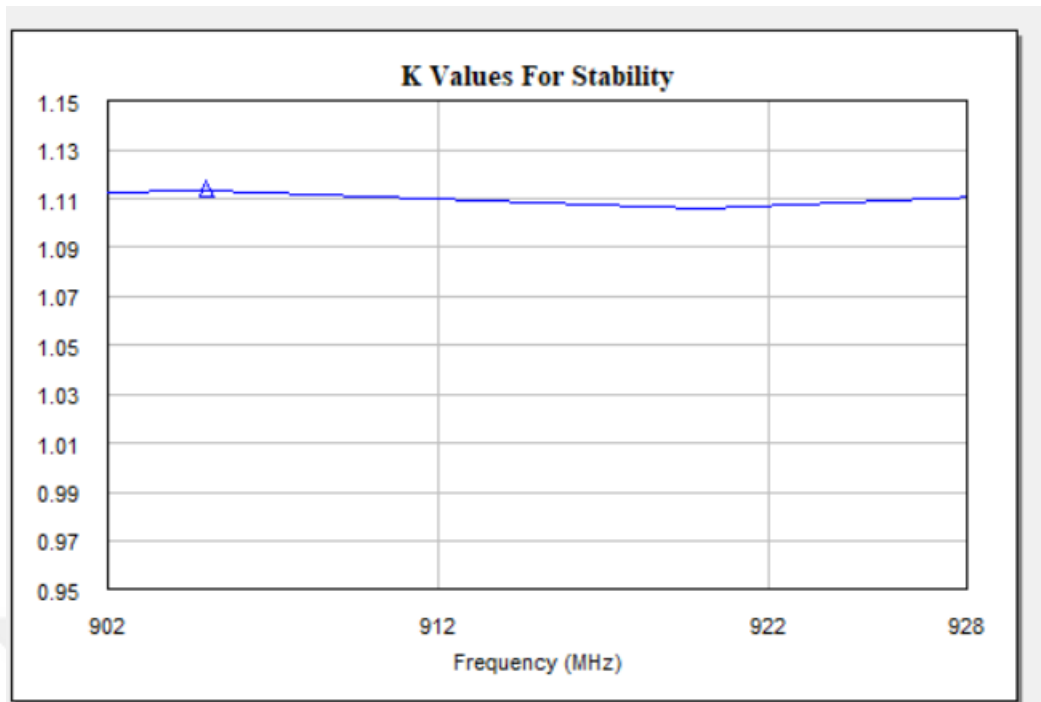


Figure 4.35 LNA stability response from 902 MHz to 928 MHz

Figure 4.35 shows that the K value is greater than 1 in the frequency range. This proves that the LNA is unconditionally stable.

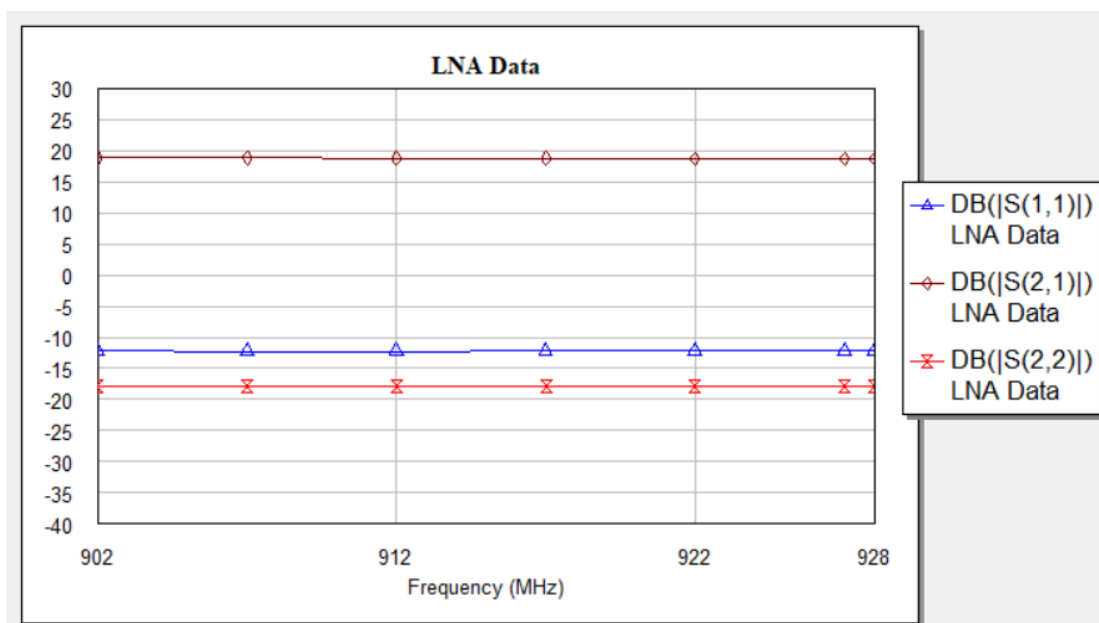


Figure 4.36 LNA S parameters from 902 MHz to 928 MHz

Since the frequency range is suitable for matching with SMD components, SMD components are utilized. Only the LNA's input is matched due to the S12 value being less than -30 dB.

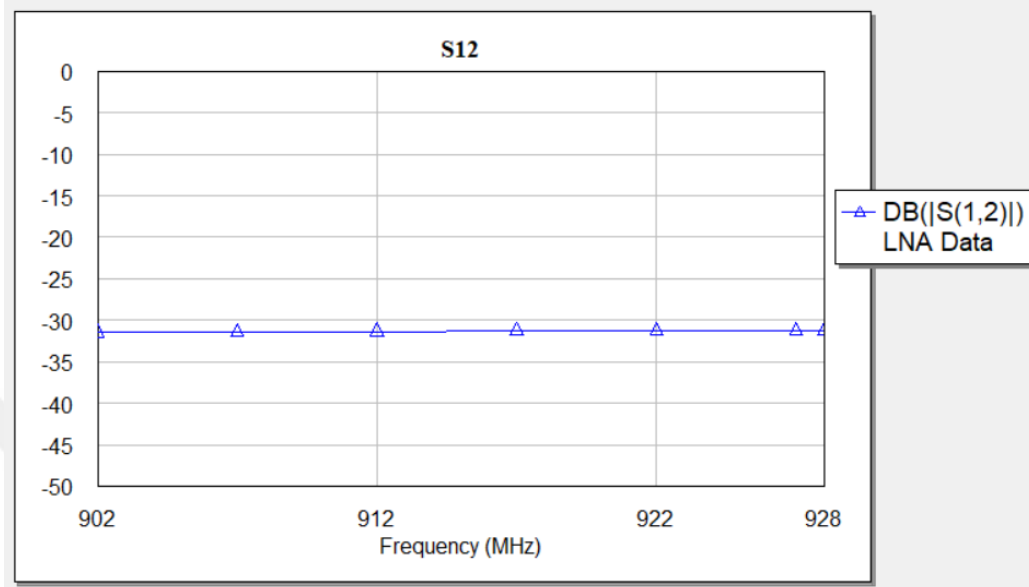


Figure 4.37 LNA S12 data between 902 MHz - 928 MHz

The matching process is conjugate matched with the bandpass filter operating in this frequency range and simulated along with the RF switch and broadband amplifier circuit.

4.7 DC – 8000 MHz LNA Circuit Design

An unconditionally stable LNA is chosen as the amplifier in the broadband LNA circuit. It has been determined that this amplifier does not need to match frequencies below 3000 MHz as a result of the spread simulation.

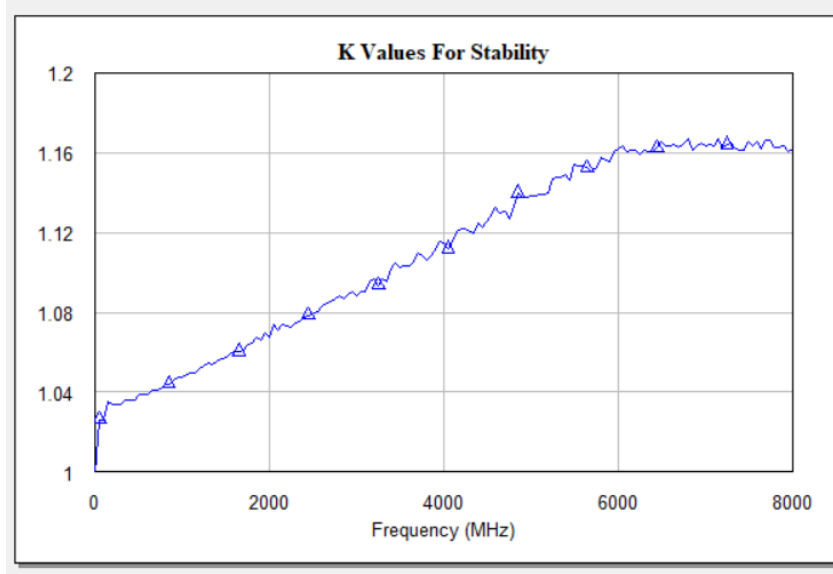


Figure 4.38 LNA stability response from DC to 8000 MHz

Figure 4.38 illustrates that the K value is greater than 1 in the frequency range. This proves that the LNA is unconditionally stable like all other LNAs.

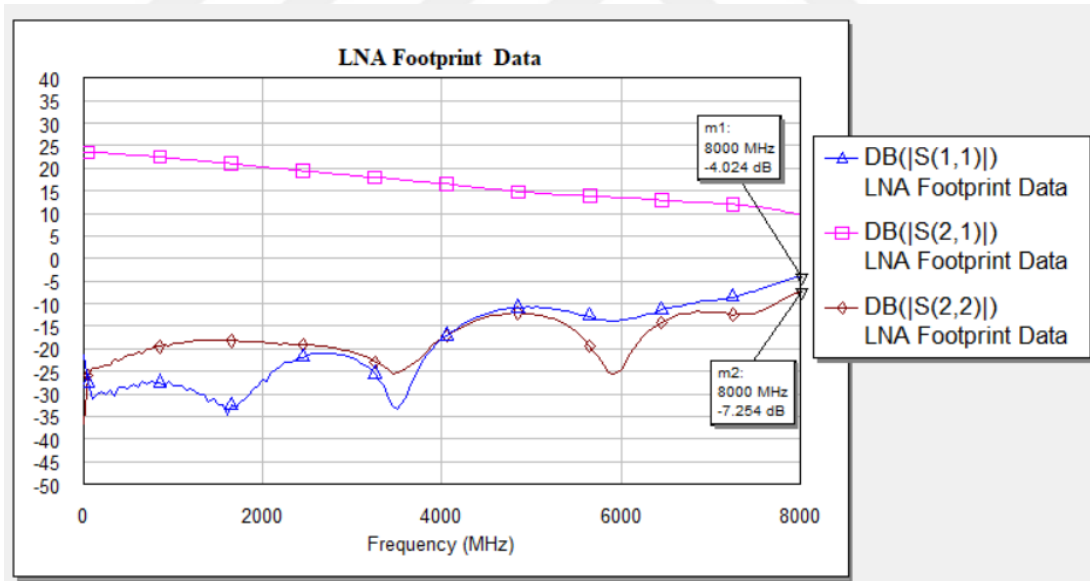


Figure 4.39 LNA footprint S parameters from DC to 8000 MHz

Input and output matching networks for LNA were designed after simulating the LNA footprint S parameters due to the decrease in return loss values as the frequency increases. The view and result of the amplifier circuit between DC and 8000 MHz is as follows:

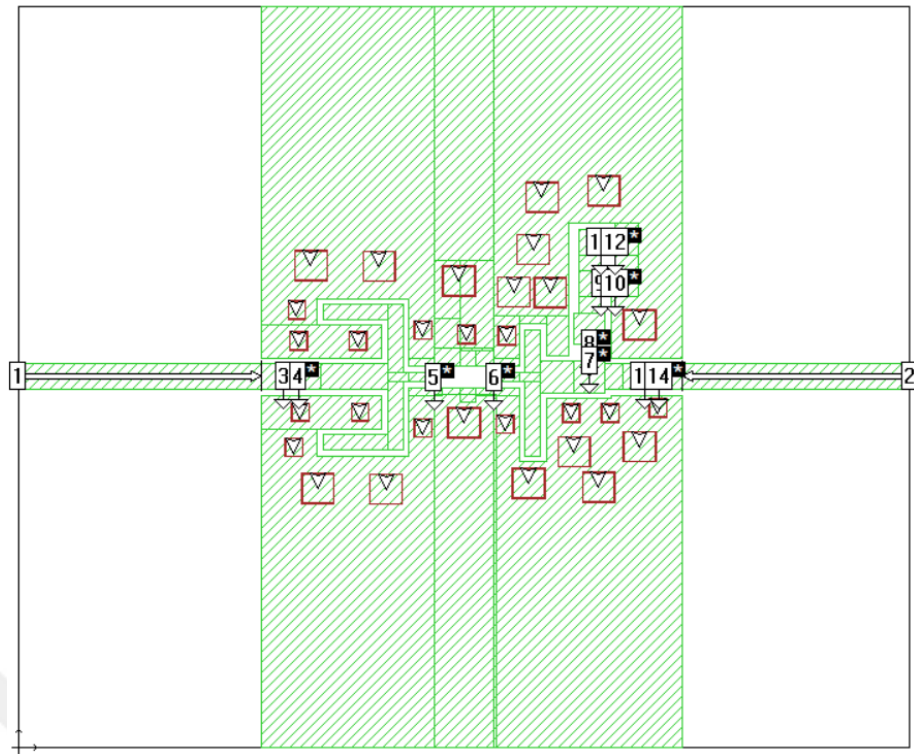


Figure 4.40 LNA circuit between DC - 8000 MHz

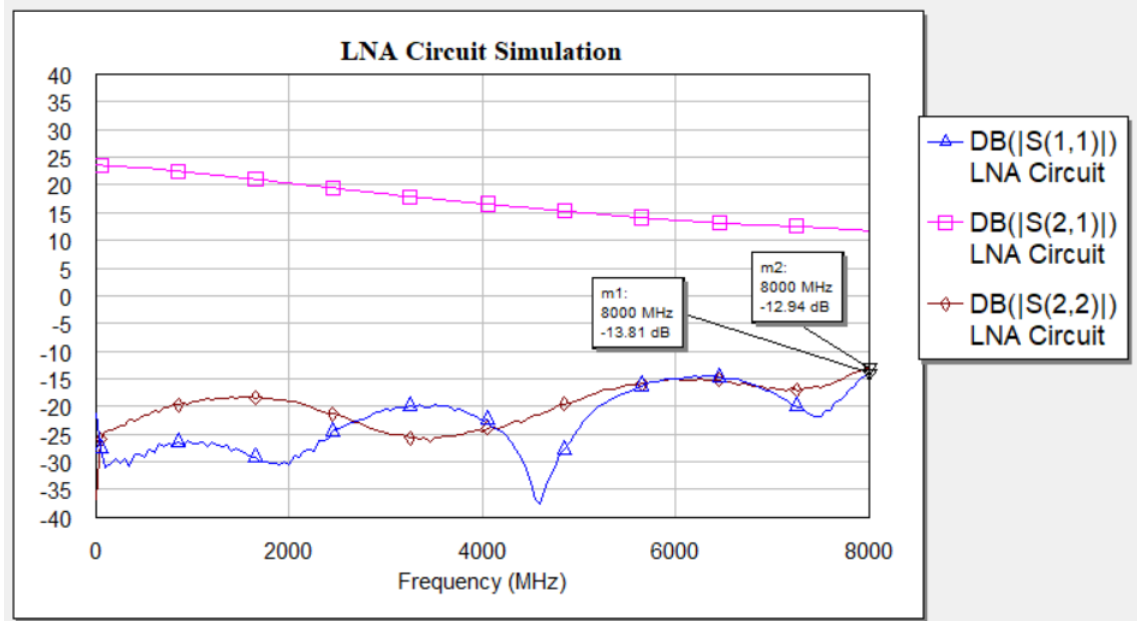


Figure 4.41 LNA circuit simulation result between DC - 8000 MHz

Figure 4.41 shows that the return loss values after the matching procedure have achieved a satisfactory level.

4.8 RF Switch Circuit Design

The RF switch utilized in this thesis receives signals from six ports and delivers them out of a single port. Control pins are used for electronic operation. The other input ports are terminated when a signal is received from one of the input ports. This technique allows the RF switch to be operated only using the input from the designated port and without the use of any additional signals. Which RF switch port will be used in which frequency band of the RF switch depends on the requirements of the project. The following describes how RF switch ports perform in the frequency range to which they are related:

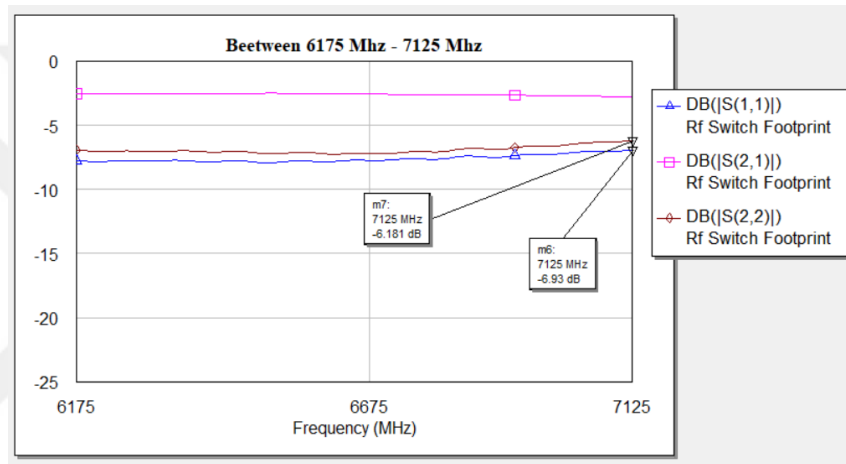


Figure 4.42 RF switch S parameter between 6175 MHz - 7125 MHz

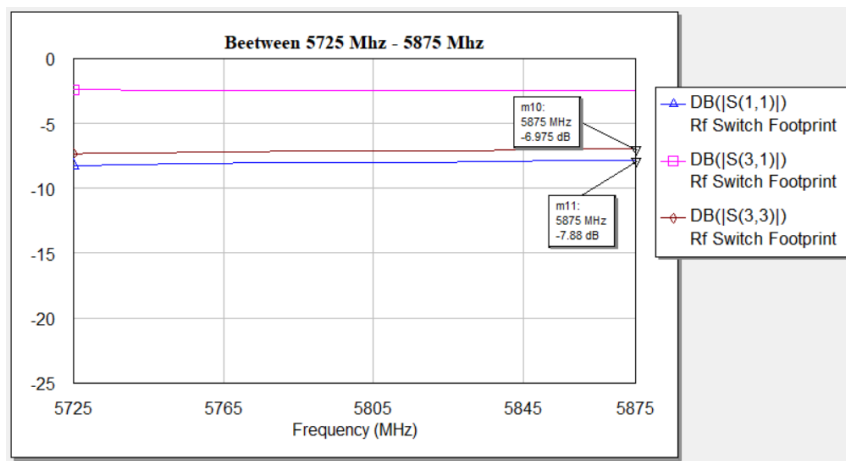


Figure 4.43 RF switch S parameter between 5725 MHz - 5875 MHz

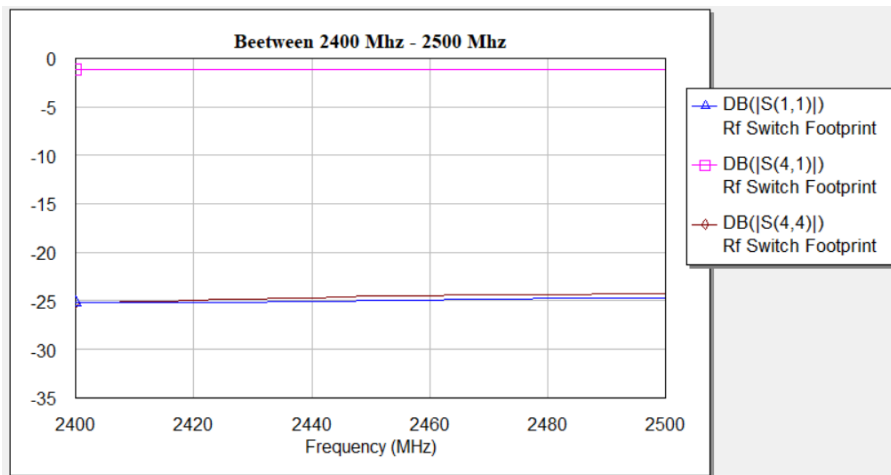


Figure 4.44 RF switch S parameter between 2400 MHz - 2500 MHz

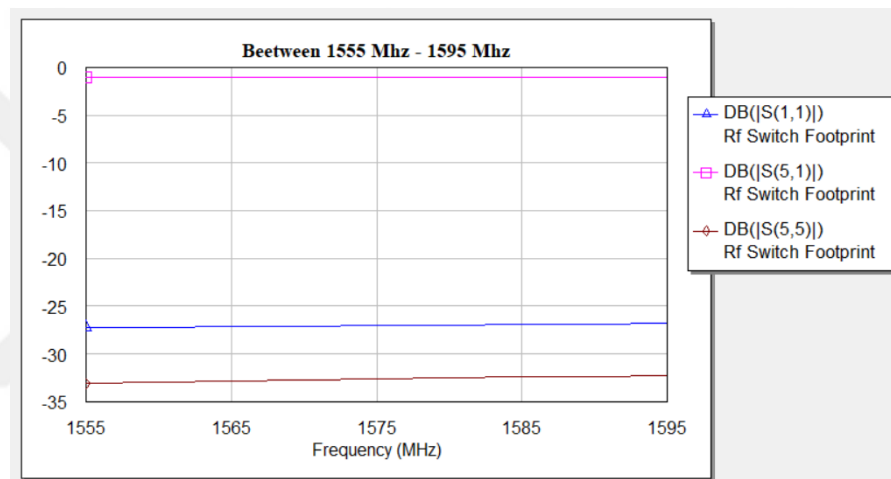


Figure 4.45 RF switch S parameter between 1555 MHz - 1595 MHz

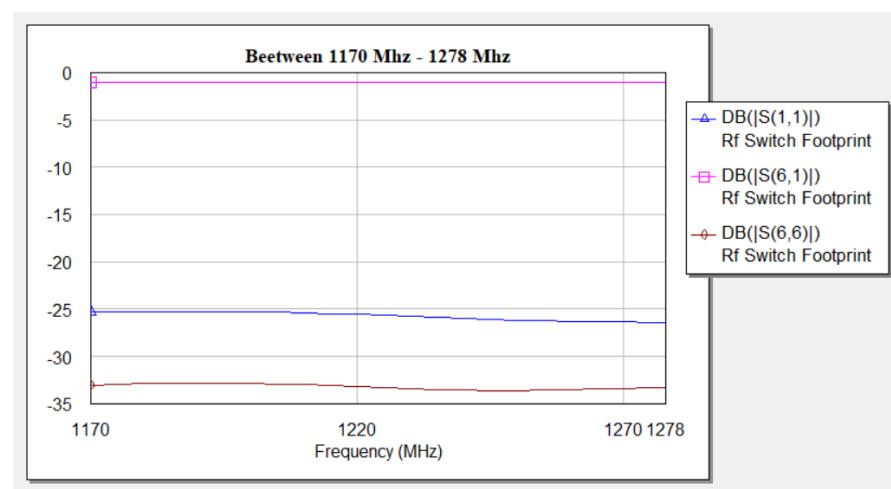


Figure 4.46 RF switch S parameter between 1170 MHz - 1278 MHz

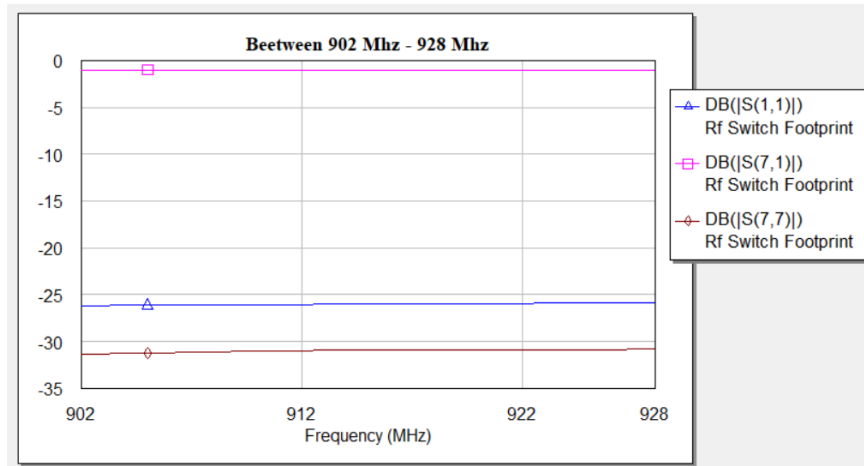


Figure 4.47 RF switch S parameter between 902 MHz - 928 MHz

As illustrated by the graphs, the RF switch input and output ports need to be matched between 5725 MHz and 5875 MHz and between 6175 MHz and 7125 MHz. Since the wavelength on the PCB is quite small in the frequency, the matching process is designed with transmission lines. On the other hand, return loss values indicate that matching procedure is not necessary in other frequency ranges.

The output port of the RF switch is matched before any other port. This matching method is designed to be broadband in scope. Then, the design of matching the RF switch input port layouts between 6175 MHz and 7125 MHz and 5725 MHz and 5875 MHz follows.

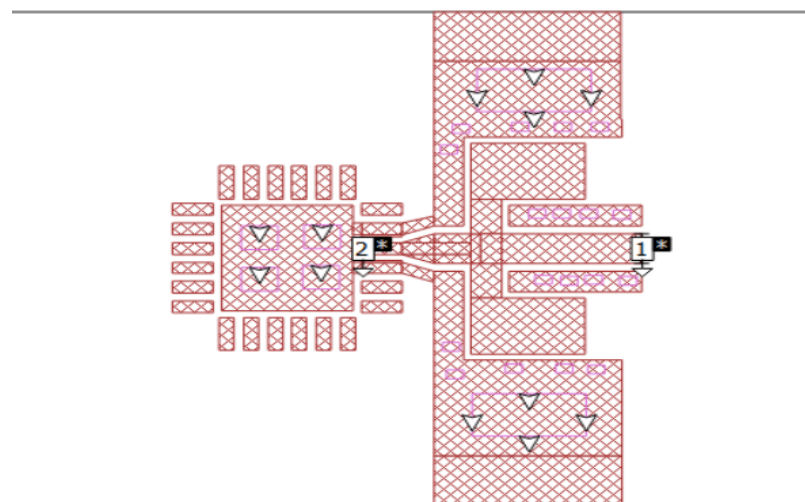


Figure 4.48 RF switch output matching circuit using transmission lines

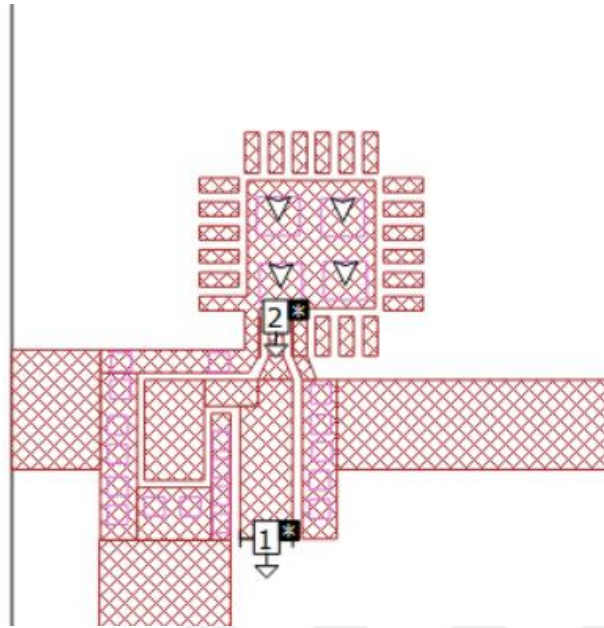


Figure 4.49 RF switch input matching circuit using transmission lines between 5725 MHz - 5875 MHz

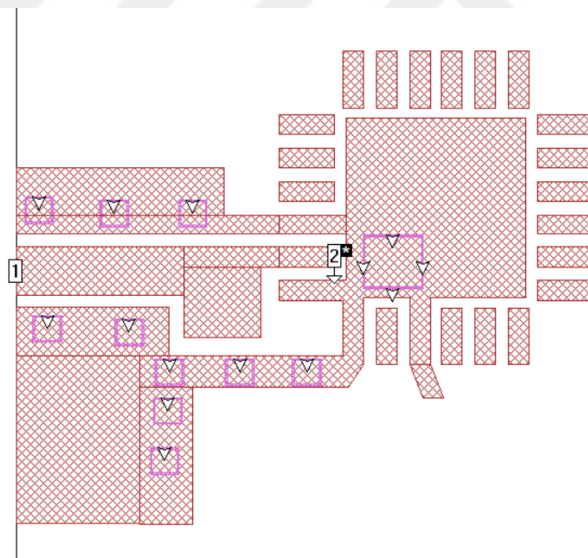


Figure 4.50 RF switch input matching circuit using transmission lines between 6175 MHz - 7125 MHz

The relevant ports of the RF switch between 5725 MHz - 5875 MHz and 6175 MHz - 7125 MHz result in the following simulation results after transmission line matching circuits are designed:

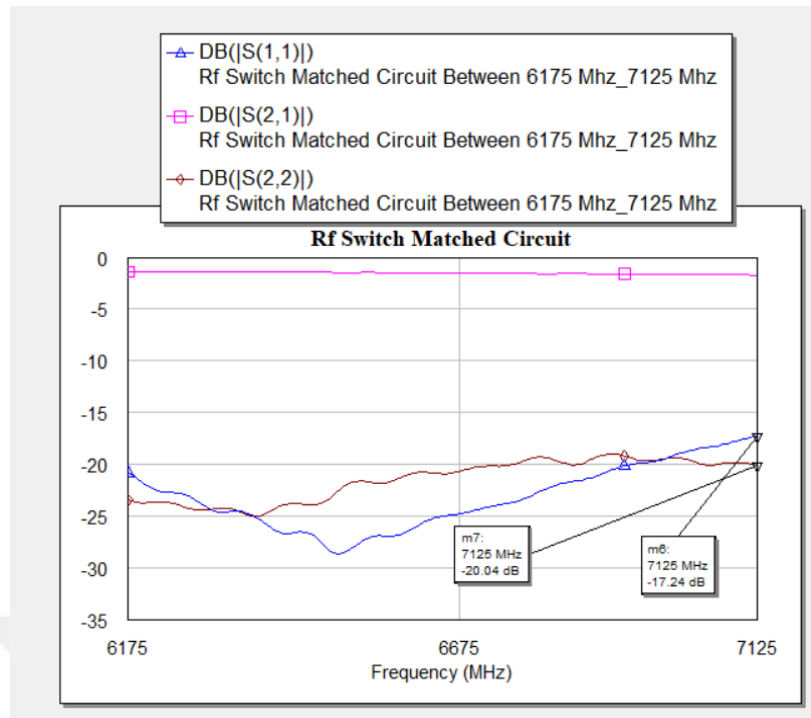


Figure 4.51 Matched RF switch simulation result between 6175 MHz - 7125 MHz

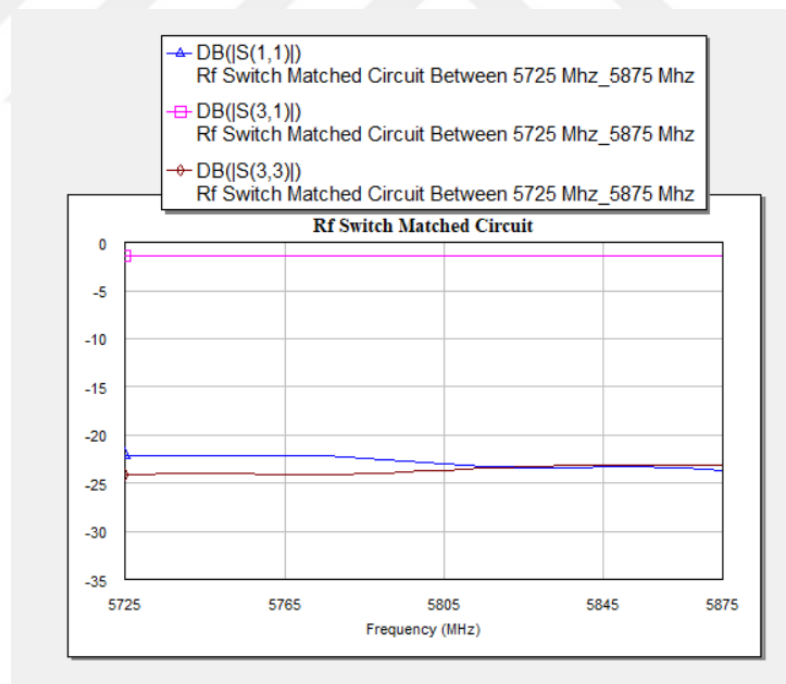


Figure 4.52 Matched RF switch simulation result between 5725 MHz - 5875 MHz

Satisfactory return loss values were also obtained in the relevant frequency ranges after matching procedures with the transmission lines.

4.9 All System Simulation Results

The entire system simulation was performed after that the RF switch was matched. The simulation results of bandpass filters, LNA circuits, RF switch, and broadband LNA circuit in each frequency range are integrated during this simulation. The following are the RF design requirements for these system simulations:

Frequency Ranges	Input Return Loss	Output Return Loss	Gain
Simulation Between 6175 MHz - 7125 MHz	>10 dB	>10 dB	>20 dB
Simulation Between 5725 MHz - 5875 MHz	>10 dB	>10 dB	>20 dB
Simulation Between 2400 MHz - 2500 MHz	>10 dB	>10 dB	>25 dB
Simulation Between 1555 MHz - 1595 MHz	>10 dB	>10 dB	>30 dB
Simulation Between 1170 MHz - 1278 MHz	>10 dB	>10 dB	>33 dB
Simulation Between 902 MHz - 928 MHz	>10 dB	>10 dB	>35 dB

Table 4.1 RF simulation requirements

As the pseudonyms of the LNAs are listed in Table 1.1, the simulations of the frequency ranges specified in that table will be designated (FR1,FR2,FR3,FR4,FR5,FR6), respectively. All system simulations are illustrated in the following graphs:

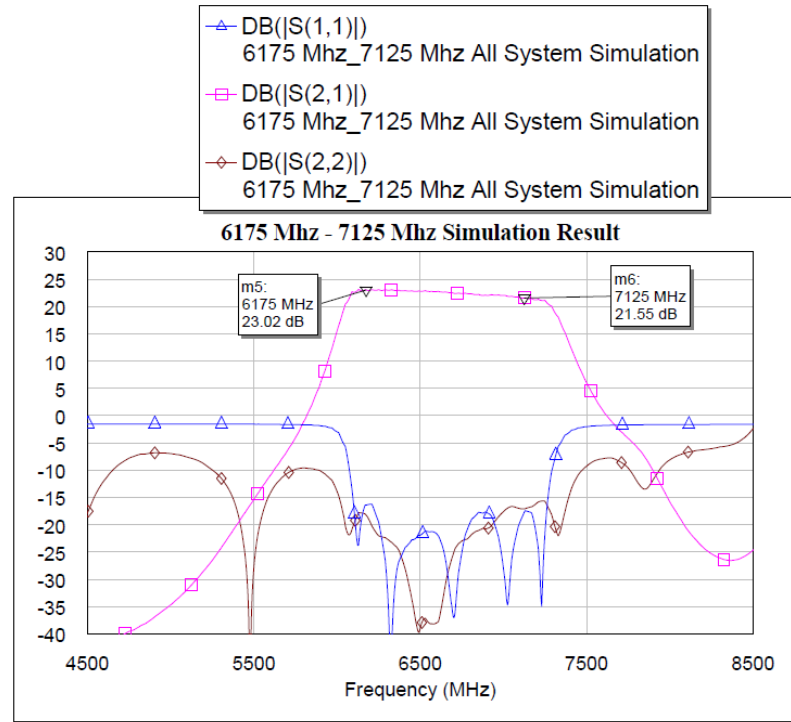


Figure 4.53 All system simulation results for FR1

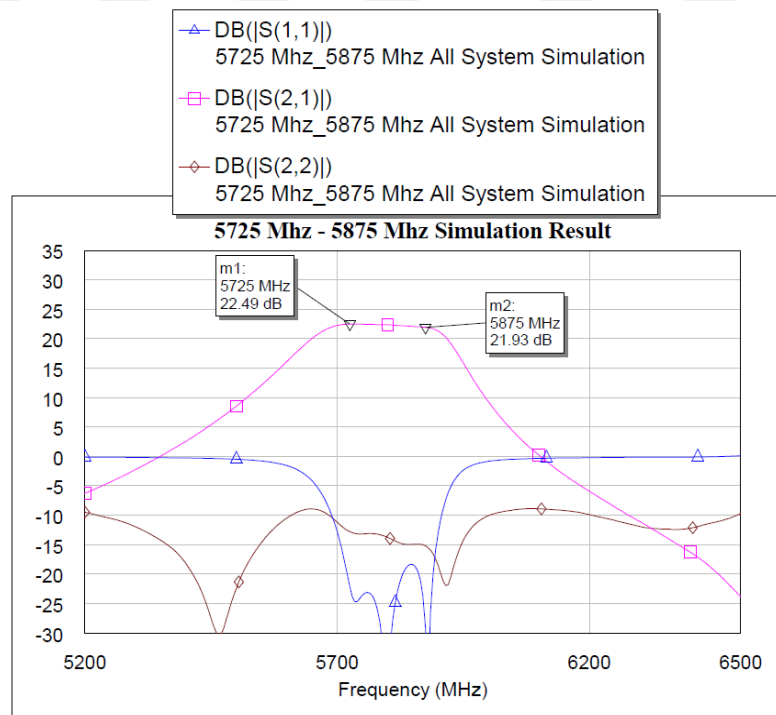


Figure 4.54 All system simulation results for FR2

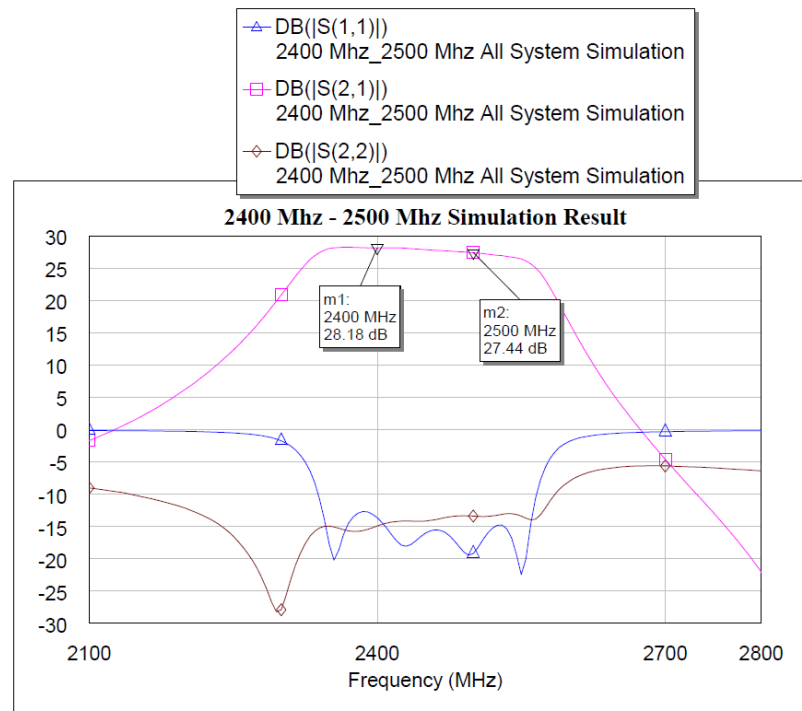


Figure 4.55 All system simulation results for FR3

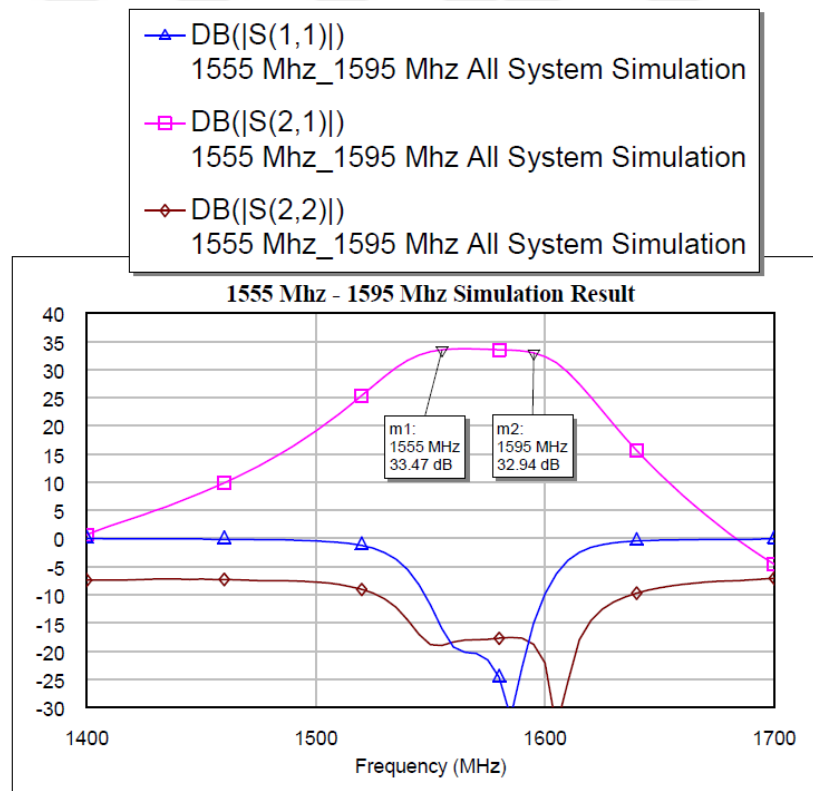


Figure 4.56 All system simulation results for FR4

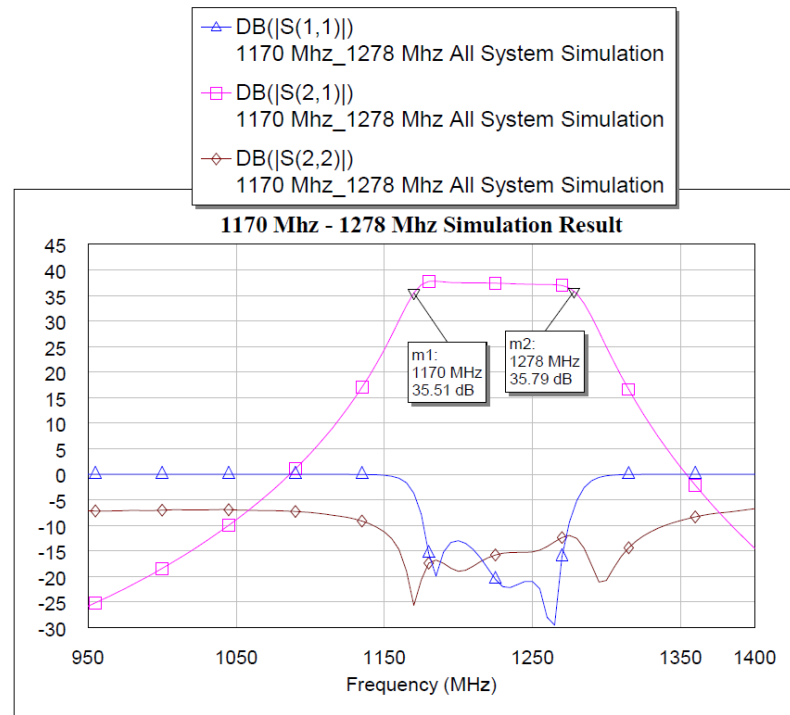


Figure 4.57 All system simulation result for FR5

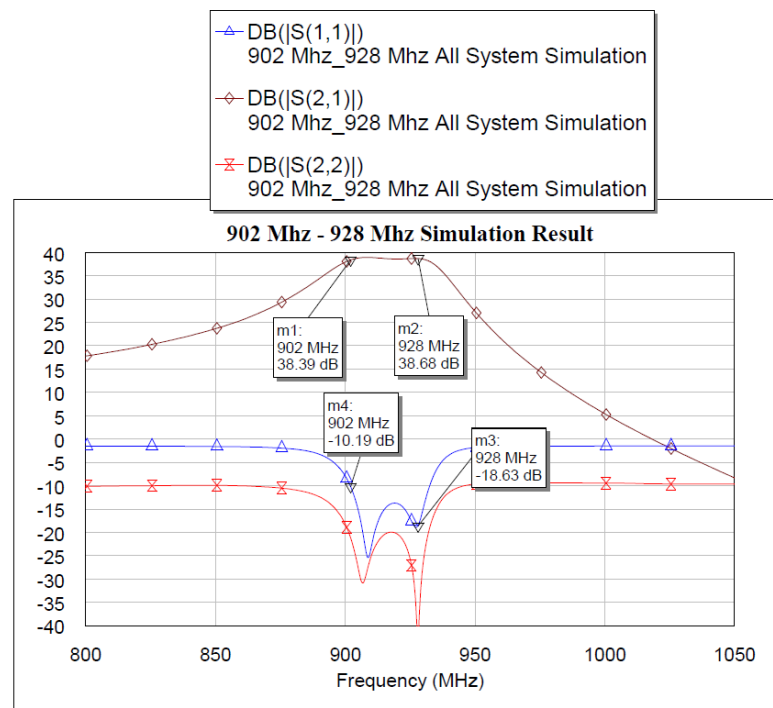


Figure 4.58 All system simulation result for FR6

The simulation results have been achieved at a satisfactory level as a consequence of cautious matching processes for each frequency range for all designs.

5. IMPLEMENTATION AND MEASUREMENT RESULTS

All designs were assembled on a single printed circuit board (PCB) after the designs and simulations were completed. After that the PCB was fabricated and the components were placed and VNA measurements were performed (Vector Network Analyzer).

5.1 Measurement Technique

The input power of the VNA was adjusted considering the saturation point of the LNAs before measurement. The signal enters via one of the six input ports, whereas the remaining five input ports are terminated. Signal transmission from the desired pin is enabled via the RF switch by sending the proper commands to the RF switch. The port from which the signal is output without modification is connected to the VNA. The Ethernet link that controls the RF switch supplies DC power to the LNAs. All LNAs have been selected to function at 5 Volts. The VNA utilized is a Copper Mountain Technologies model with the following features:

Frequency Range	9 kHz to 8.5 GHz
Measured Parameters	S11, S22, S21, S12
Sweep Types	Linear Frequency, Log Frequency, Segment, Power Sweep
Dynamic Range	130 dB typ. (10 Hz IF BW)
Measurement Speed	70 μ s -55 dBm to +5 dBm
Output power Adjustment Range	-55 dBm to +5 dBm
Measurement Points Per Sweep	Up to 200,001

Table 5.1 VNA Parameters

5.2 Comparison Of Simulation Results With Measurement Results

The measurement results were obtained by separately measuring each frequency range of the circuit systems. The results of the simulation were compared to the obtained measurements.

5.2.1 Comparison Of Simulation Results With Measurement Results Between 6175 MHz - 7125 MHz

Tuning procedure is not required in this frequency range. The following measurements are as follows:

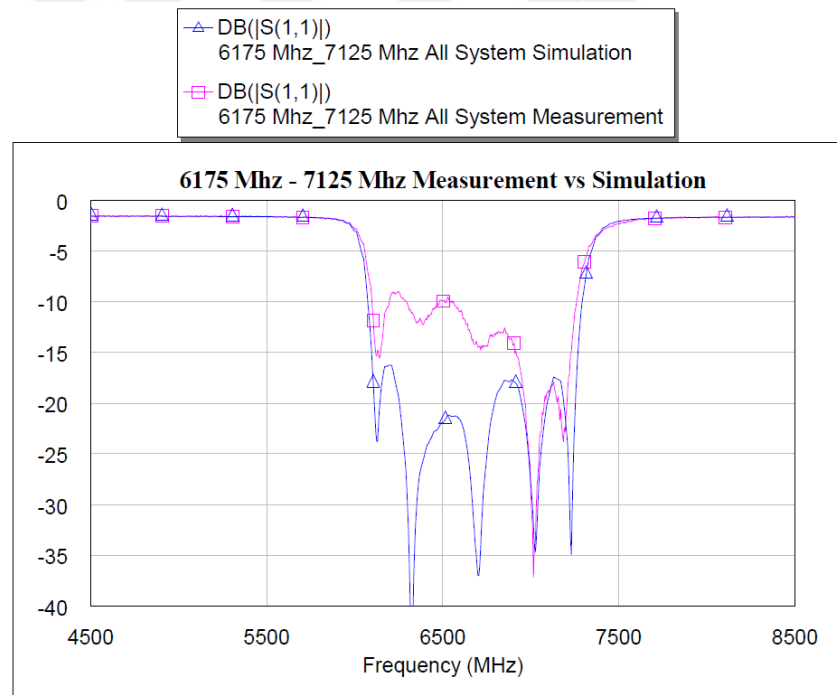


Figure 5.1 Comparison of measurement result and simulation result between 6175 MHz - 7125 MHz (S11)

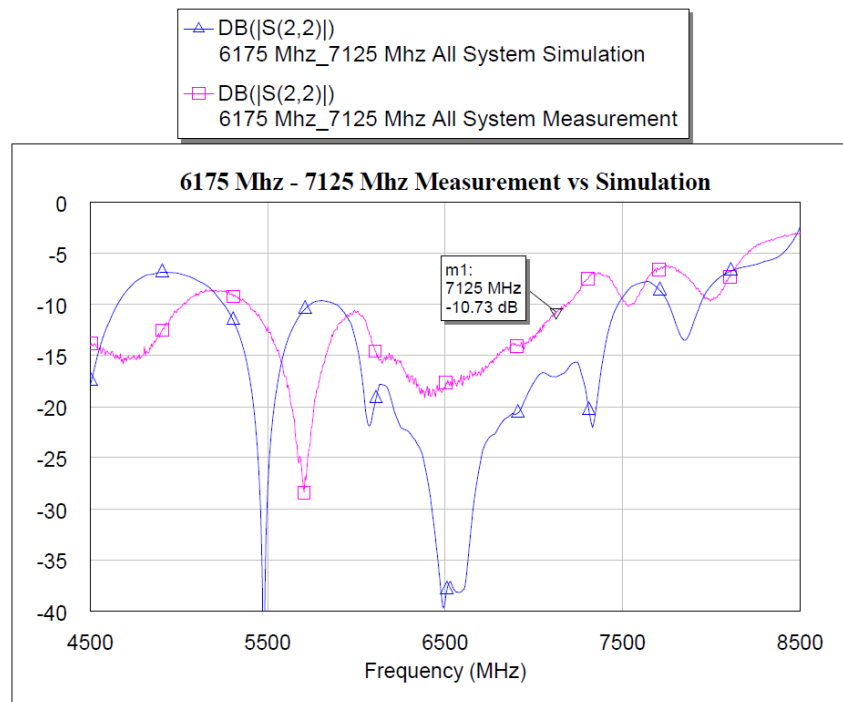


Figure 5.2 Comparison of measurement result and simulation result between 6175 MHz - 7125 MHz (S22)

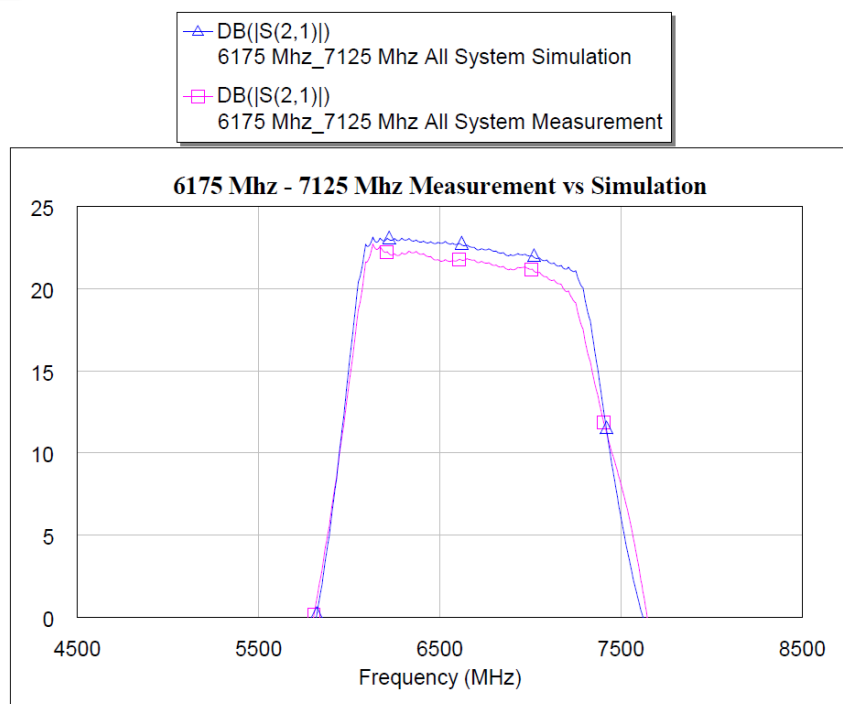


Figure 5.3 Comparison of measurement result and simulation result between 6175 MHz - 7125 MHz (S21)

Figure 5.2 and Figure 5.3 show that the measurement results and simulation results are compatible with each other. Figure 5.1 illustrates that while the level of agreement between the simulation and measurement results is not perfect, it is tolerable.

5.2.2 Comparison Of Simulation Results With Measurement Results Between 5725 MHz - 5875 MHz

The tuning process is not required while measuring in this frequency range. The comparison between the measurement and simulation results is represented in the figures:

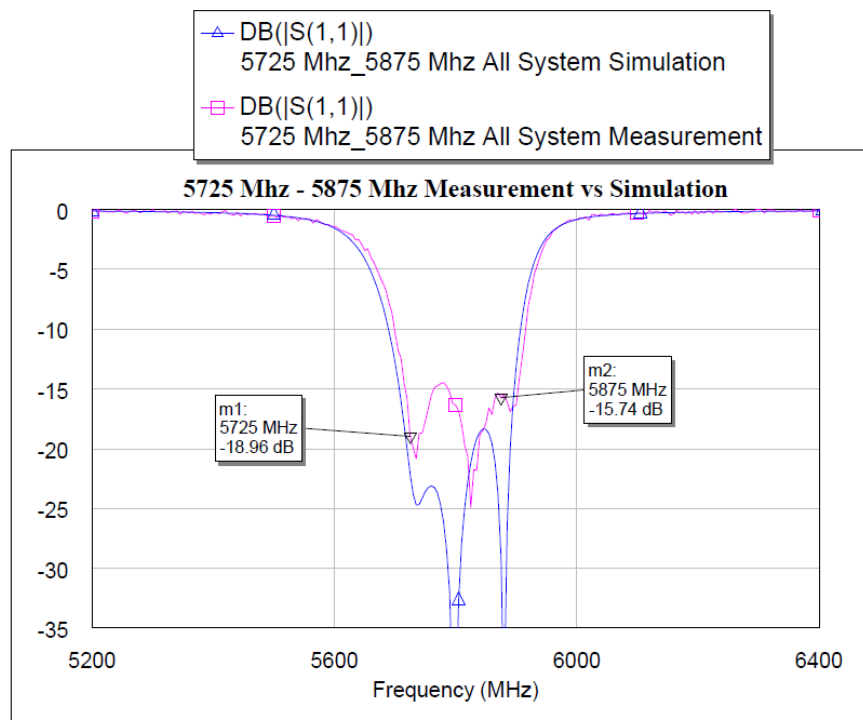


Figure 5.4 Comparison of measurement result and simulation result between 5725 MHz - 5875 MHz (S11)

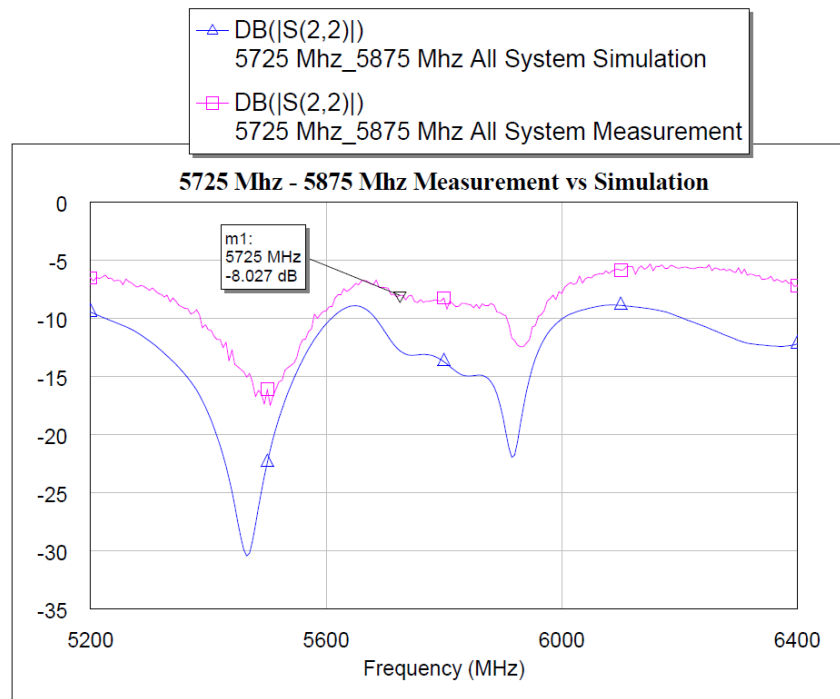


Figure 5.5 Comparison of measurement result and simulation result between 5725 MHz - 5875 MHz (S22)

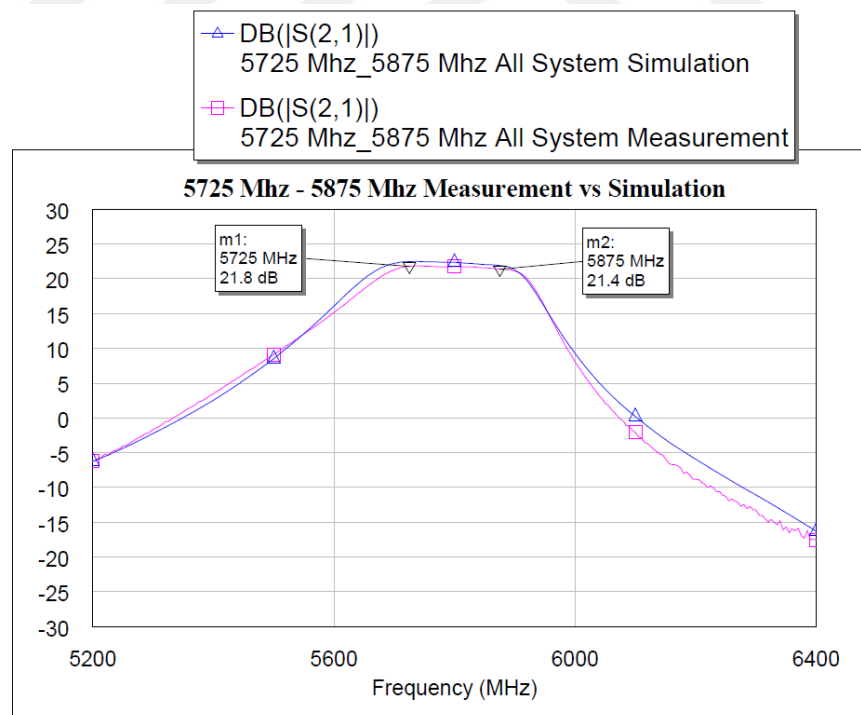


Figure 5.6 Comparison of measurement result and simulation result between 5725 MHz - 5875 MHz (S21)

The measurement and simulation results are consistent with one another, as shown in Figures 5.4 and 5.6. Figure 5.5 shows that while there is a bit of inconsistency between the simulation and measurement results, it is tolerable.

5.2.3 Comparison Of Simulation Results With Measurement Results Between 2400 MHz - 2500 MHz

No tuning procedure was required for the measurement results in this frequency range. The main reason for this is that the parasitic effects of SMD components diminish with decreasing frequency. The resulting measurements are as follows:

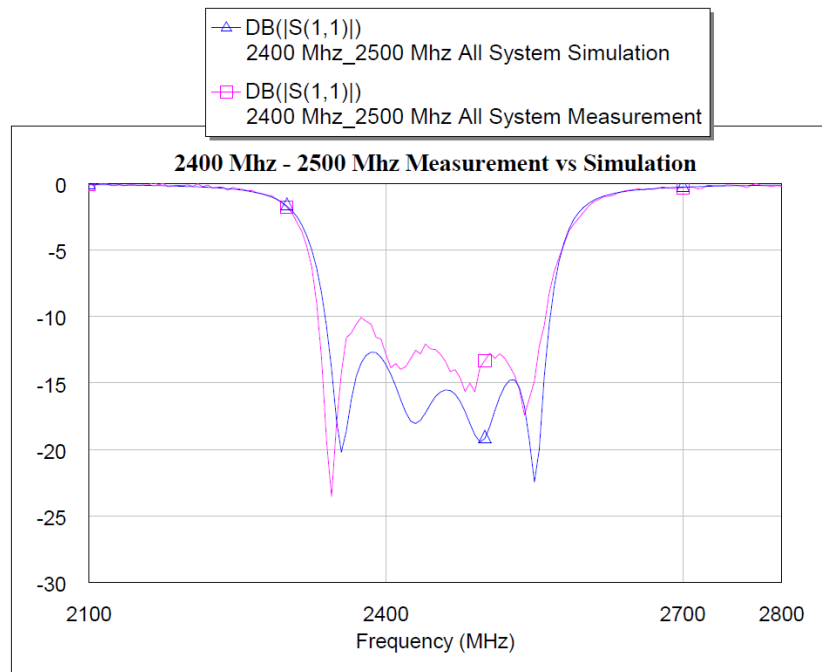


Figure 5.7 Comparison of measurement result and simulation result between 2400 MHz - 2500 MHz (S11)

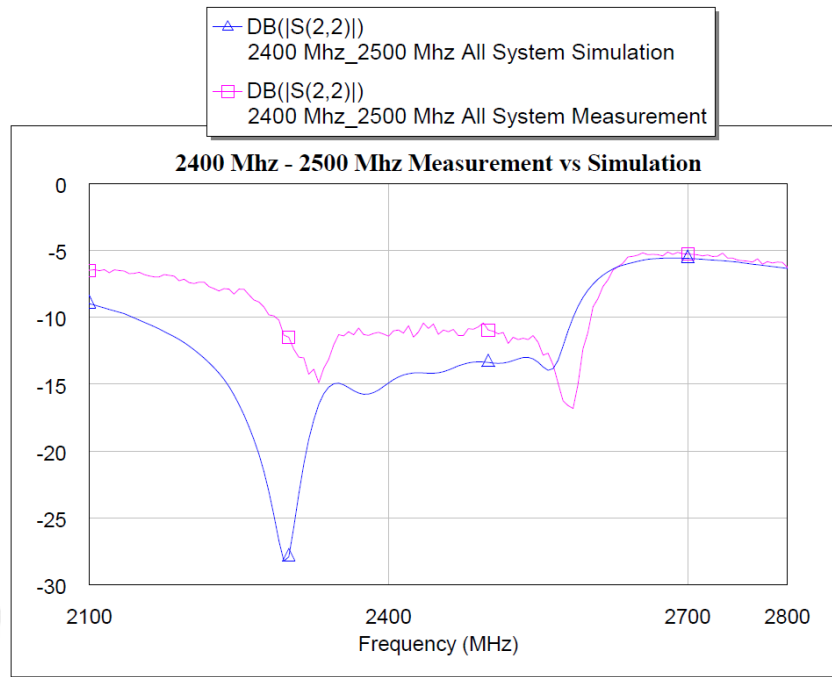


Figure 5.8 Comparison of measurement result and simulation result between 2400 MHz - 2500 MHz (S22)

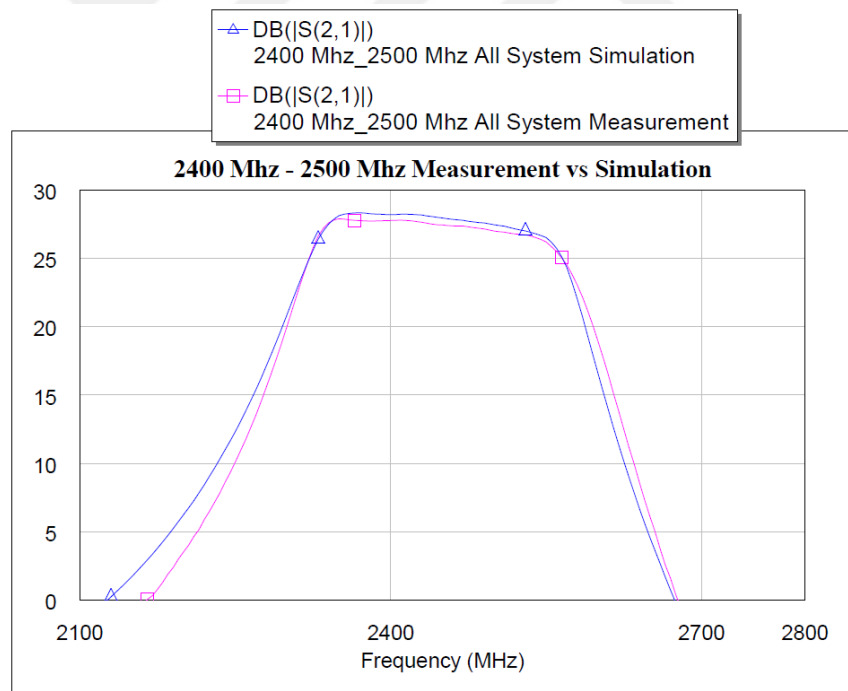


Figure 5.9 Comparison of measurement result and simulation result between 2400 MHz - 2500 MHz (S21)

The measurement and simulation results are consistent with one another, as shown in all figures in this chapter.

5.2.4 Comparison Of Simulation Results With Measurement Results Between 1555 MHz - 1595 MHz

There is no need for tuning in this frequency range. The following are the measuring results:

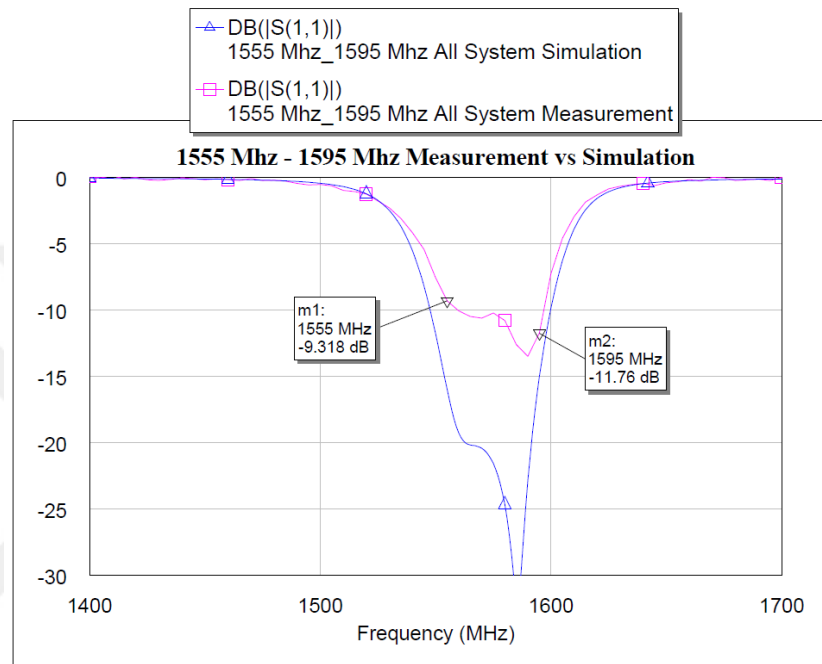


Figure 5.10 Comparison of measurement result and simulation result between 1555 MHz - 1595 MHz (S11)

Figure 5.10 demonstrates that while there is some acceptable inconsistency between the simulation and measurement results.

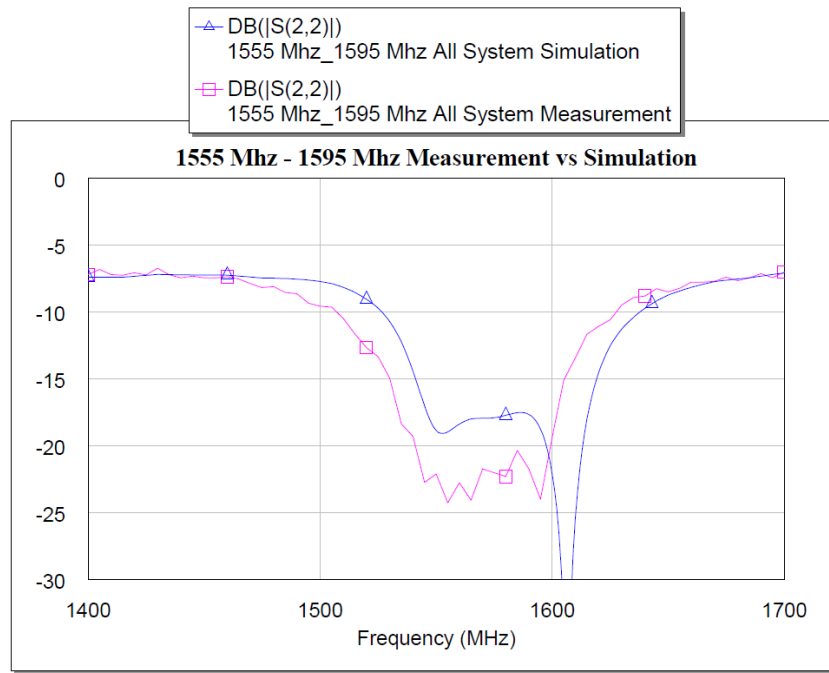


Figure 5.11 Comparison of measurement result and simulation result between 1555 MHz - 1595 MHz (S22)

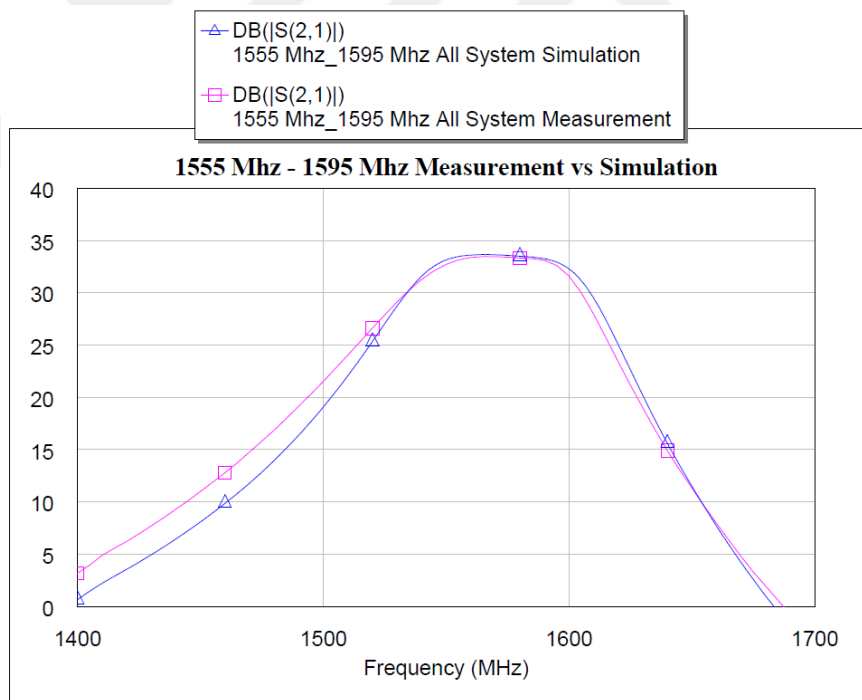


Figure 5.12 Comparison of measurement result and simulation result between 1555 MHz - 1595 MHz (S21)

Compared to Figure 5.10, Figures 5.11 and 5.12 show measurement and simulation results that are more consistent. In fact, the measurement result shown in Figure 5.11 is really superior to the simulated result.

5.2.5 Comparison Of Simulation Results With Measurement Results Between 1170 MHz - 1278 MHz

Tuning procedure is not required in this frequency range. The following measurements are as follows:

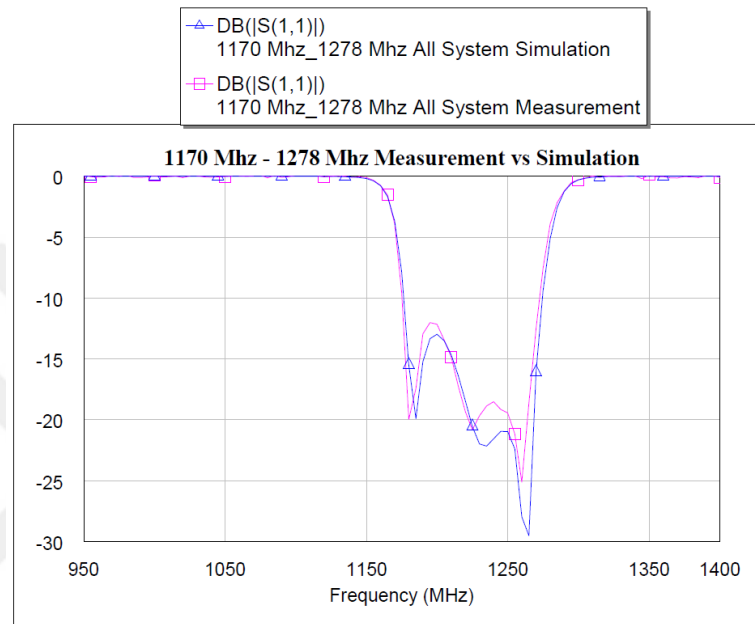


Figure 5.13 Comparison of measurement result and simulation result between 1170 MHz - 1278 MHz (S11)

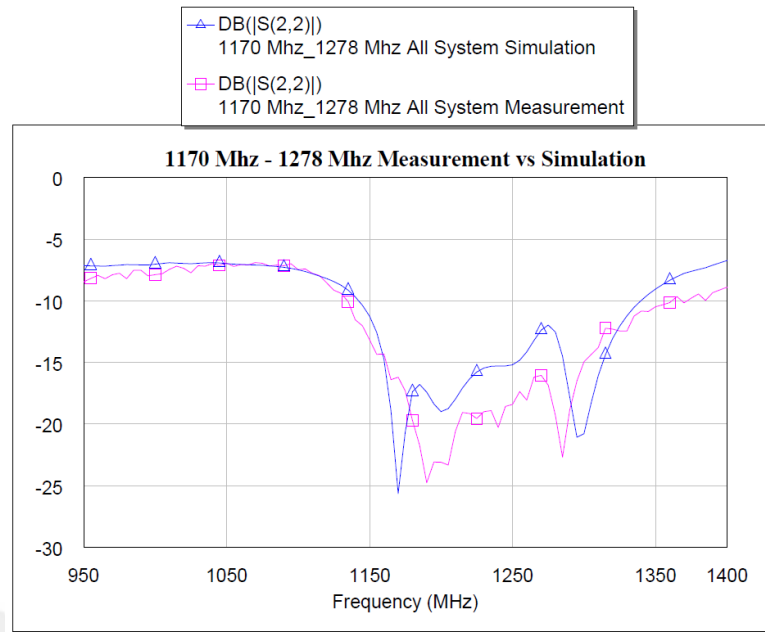


Figure 5.14 Comparison of measurement result and simulation result between 1170 MHz - 1278 MHz (S22)

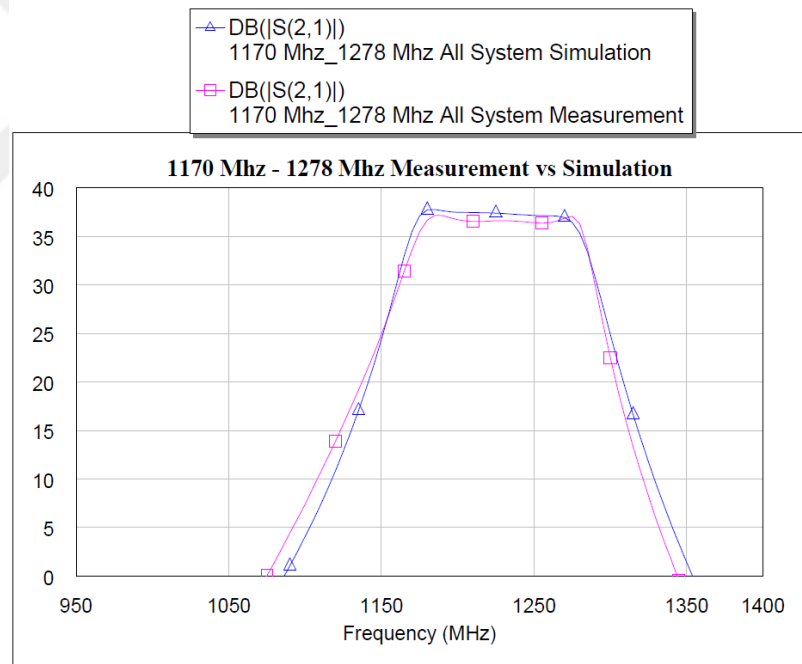


Figure 5.15 Comparison of measurement result and simulation result between 1170 MHz - 1278 MHz (S21)

All the figures demonstrate how the measurement and simulation results match one another in this chapter.

5.2.6 Comparison Of Simulation Results With Measurement Results Between 902 MHz - 928 MHz

Tuning procedure is not required in this frequency range. The following measurements are as follows:

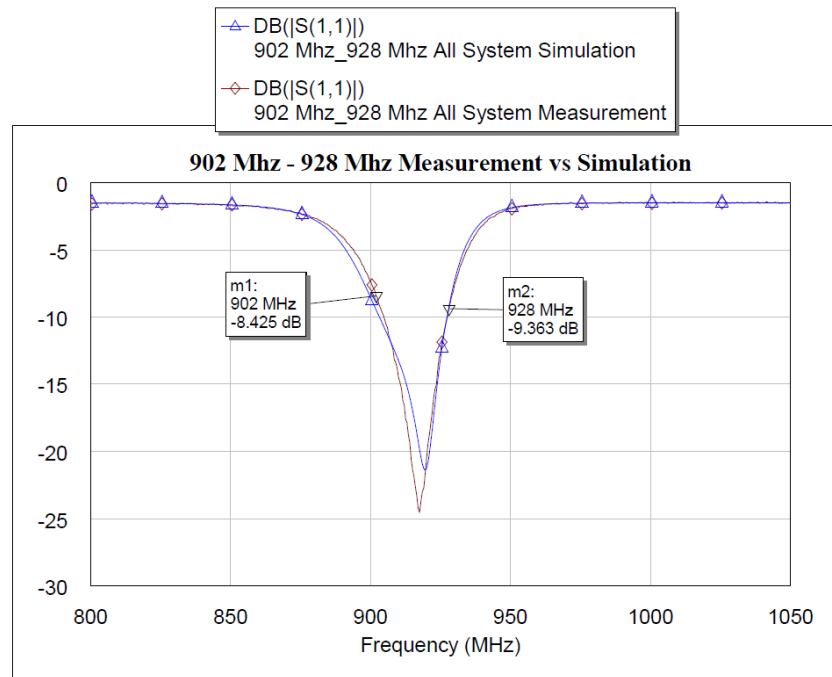


Figure 5.16 Comparison of measurement result and simulation result between 902 MHz - 928 MHz (S11)

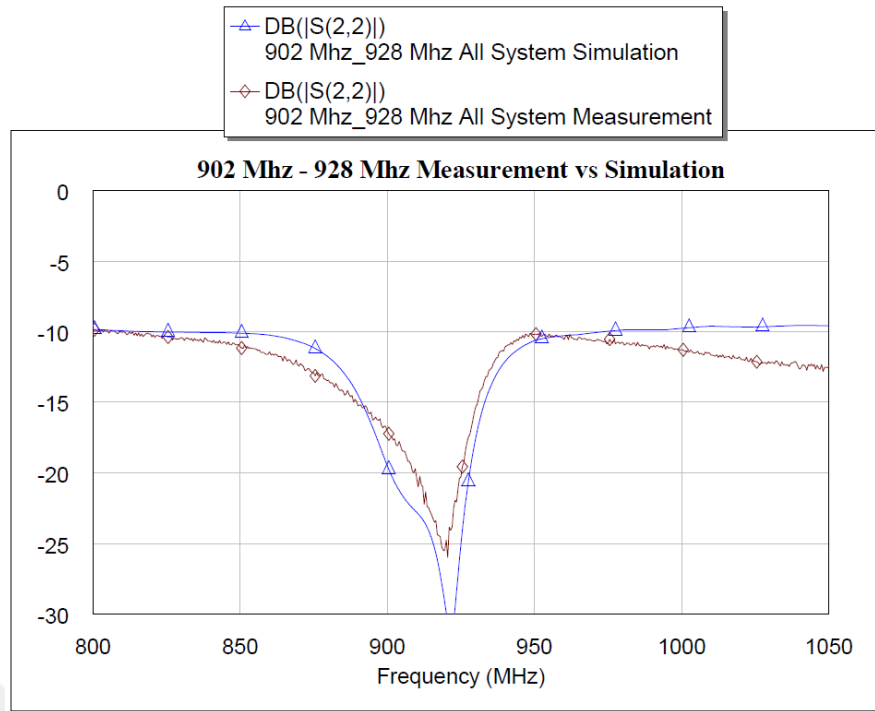


Figure 5.17 Comparison of measurement result and simulation result between 902 MHz - 928 MHz (S22)

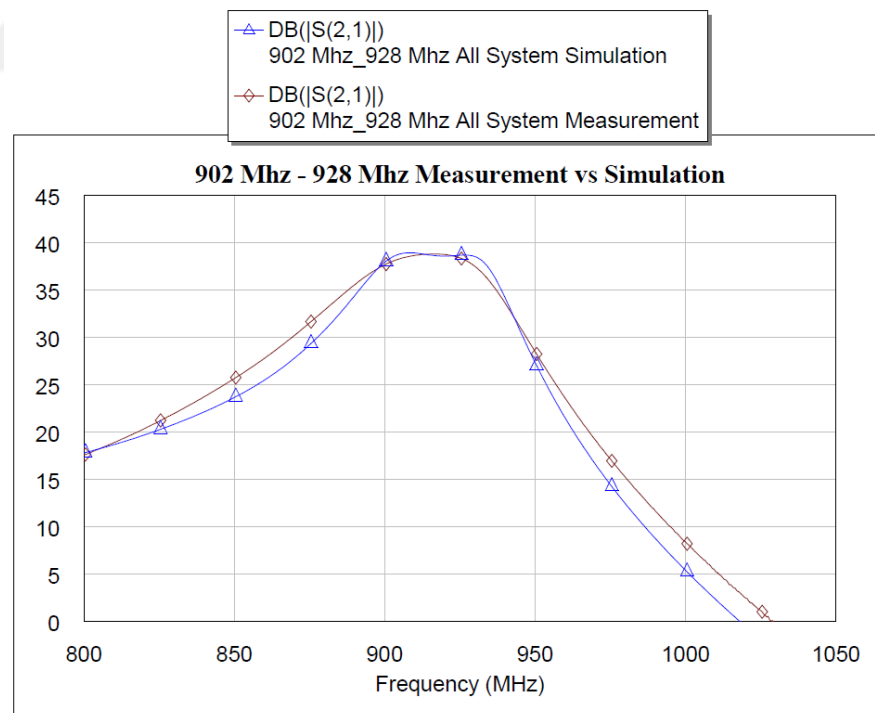


Figure 5.18 Comparison of measurement result and simulation result between 902 MHz - 928 MHz (S21)

All figures show how the measurement and simulation results match with each other, also in this section.

5.3 Evaluation Of Measurement Results

Results from the comparison of measurement and simulation results were satisfactory. Simulation of all transmission lines down to the components added to the PCB has been considered to ensure that the measurement results are quite close to the simulation results. The PCB placed in the PCB case has been used in a way that will have the least impact on the measurement results. Therefore, absorbers are placed on specific sections of the case's cover. Consequently, the desired measurement results are still achieved even with the cover attached. A second reason why absorbers are put in the PCB case is because the PCB cannot be simulated as a whole due to the size of the PCB circuit. It was impossible to simulate an entire PCB with even the most advanced computer. One method of overcoming this difficulty is to fit the PCB and PCB case through each other as a cavity. However, since this design is a prototype, this more expensive option was not preferred and the desired results were achieved with absorbers.

The absorber used is as shown in the picture:



Figure 5.19 Absorber used in PCB case

6. CONCLUSION AND FUTURE WORK

The implementations performed within the scope of this thesis have resulted in the development of a complex and specialized product made up of multiple microwave components. This product operates by switching incoming signals between 902 MHz and 7125 MHz with an RF switch and amplifying these signals in matched condition. As a result, a single product has been designed on a single PCB, as opposed to using microwave circuits that are purchased individually, due to the procedures that were carried out. This product can also be sold for commercial market as a result of the amplification of signals received within the given frequency range.

The measurement results of this product show that this prototype is of commercially feasible quality. This product's specifications are detailed in Table 6.1:

Frequency Ranges	Input Return Loss	Output Return Loss	Gain
Between 6175 MHz - 7125 MHz	>8 dB	>8 dB	>20 dB
Between 5725 MHz - 5875 MHz	>10 dB	>7 dB	>20 dB
Between 2400 MHz - 2500 MHz	>9 dB	>9 dB	>25 dB
Between 1555 MHz - 1595 MHz	>8 dB	>10 dB	>30 dB
Between 1170 MHz - 1278 MHz	>10 dB	>10 dB	>33 dB
Between 902 MHz - 928 MHz	>8 dB	>10 dB	>35 dB

Table 6.1 Fabricated prototype product measurements

Low noise amplifiers were used in the development of this device, as is well known. Since noise figure data for these low noise amplifiers could not be obtained from the manufacturers, noise analysis could not be performed in the simulation.

The fundamental contribution of this thesis is that it is not only a circuit constructed with RF switches, but also a high gain RF switch circuit filtered with bandpass filters within its own circuit and has commercial applications. Numerous varieties of RF switches and amplifier circuits were encountered within the scope of the thesis on the international RF electronics market. However, no products have been

found in which the RF switch is integrated with amplifier circuits and bandpass filters in these frequency ranges.

The following stages of the product developed within the scope of the thesis can be designed in various frequency ranges.



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APPENDIX

APPENDIX A: Publications Within The Scope Of Thesis

Budak A.M., Güneş A., Arvas E., (2022), “Yüksek Kazançlı Geniş Bantlı SP6T Uyumlanmış Anahtar Matrisi”, Gebze Technical University Institute of Natural and Applied Sciences, 6th Graduate Studies Symposium & Presentation Days, 01-02 June.

