DIPLEXER DESIGN USING MICROSTRIP LINES

A Thesis submitted towards the partial fulfilment of the requirement for the award of the degree of

Master of Technology

in

Microwave and Optical Communication

Submitted by

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2K11 / MOC / 14

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JUNE 2013

CERTIFICATE

This is to certify that the dissertation title ***"Diplexer Design using Microstrip lines"*** is the authentic work of **Ms. Varsha Mishra** under my guidance and supervision in the partial fulfillment of requirement towards the degree of Master of Technology in ***Microwave and Optical Communication Engineering***, jointly run by the Department of Electronics and Communication Engineering and Department of Applied Physics in ***Delhi Technological University.***

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I hereby declare that all the information in this document has been obtained and presented in accordance with academic rules and ethical conduct. This report is my own, unaided work. I have fully cited and referenced all material and results that are not original to this work. It is being submitted for the degree of Master of Technology in Engineering at the Delhi Technological University. It has not been submitted before for any degree or examination in any other university.

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ABSTRACT

In this Project, a new microstrip structure is proposed and studied to design microstrip diplexer. Modern wireless communication systems demand RF devices operating in multiple frequency bands. A diplexer, is an essential component in multi-service and multi-band communication systems, is a three terminal device that separates the input signals to two output ports. A well designed diplexer should have low cost and high performance.

To demonstrate the design ideas, the equivalent circuits of the proposed Diplexer are built and studied and then Simulated in ADS software tool. It is found that the different loads on different positions of the proposed half-wavelength resonator make the resonator have different features, which will easily control the characteristic of the diplexers. And here, resistor, open stub, and shorted stub are used as loaded elements. It is found the resistor loaded on the centre of the microstrip line resonator can extremely reduce the unloaded quality factor of even-mode resonant frequency, which can be used to suppress the harmonics of the diplexer. The loaded open stub not only can reduce the size of the diplexer, but also control the frequency ratio between the fundamental frequency and second harmonic of a resonator, which can increases the frequency ratio between the two pass bands of the diplexer. As for the loaded shorted stub, it can enlarge the size of the diplexer.

Here in our project diplexer is designed with only four stubs and two transmission lines. The transmission line used has the same characteristic impedances. While it has a good transmission performances and good isolation between the output ports. Here microstrip diplexer is designed and simulated in ADS software tool. The Diplexer is designed to operate with 2.4 GHz and 2.85GHz frequencies.

Different results for the diplexer have been found in terms of S parameter. Which shows different characteristics of the device. Here a good agreement with the calculated results and the simulated results is achieved.

To my parents….

ACKNOWLEDGEMENT

I am indebted to my thesis supervisor **Dr. Priyanka Jain** (ECE department) and **Prof. R.K. Sinha** (physics Department), for their gracious encouragement and very valued constructive criticism that has driven me to carry the project successfully.

I am deeply grateful to **Prof. Rajiv Kapoor**, Head of Department Electronics and Communication Engineering Delhi Technological University for his support and encouragement in carrying out this project.

I wish to express my heart full thanks to **Prof. S.C. Sharma**, Head of Department of Applied Physics, Delhi Technological University for his support that helped me a lot in successful completion of this project.

I wish to express my heart full thanks to my branch coordinator **Dr. Ajeet Kumar (**physics Department) and my all friends for their goodwill and support that helped me a lot in successful completion of this project.

I express my deep sense of gratitude to my mother **Mrs. Nirmala Mishra** and My Father **Mr. Dinesh Mishra** for their time to time supports and encouragements.

I am also grateful to **Prof. P. B. Sharma,** Vice-Chancellor, Delhi Technological University for providing the research environment in the institute.

Finally I would like to thank **God** for his blessings without which nothing is possible in this world.

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Introduction

1.1 Overview

Diplexers are three port devices which are commonly used behind the wide-band or multi-frequency antennas in transceiver applications, take normally two frequencies into their input port and separate them to two output ports. Diplexing is used to prevent inter modulation and keep reflected power (VSWR) to a minimum for each input transmitter and frequency. While diplexers can combine a relatively wide bandwidth, the major limitation comes with the antenna itself, which must be sufficiently wideband to accept all of the signals being passed through it, and transfer them to the air efficiently. Many other large UHF-/VHF-transmitters use diplexers. The number of transmitters which can share an antenna is restricted by the spacing of their frequency bands. Transmitters whose frequencies are too close together cannot be combined successfully by a diplexer.

Diplexers are also used at medium wave broadcasting stations. However their use is not that common in this frequency range because the corresponding wavelength varies much more across the medium wave band than across the FM band and so it is more practicable to use a separate antenna for each frequency: medium wave transmission sites usually broadcast only on one to four frequencies, while FM-broadcasting sites often uses four and more frequencies.

Diplexers may be used as a back-up device. For example maintenance work at one antenna of a medium wave transmission site that has two antennas transmitting on two frequencies. Then the other antenna can be used for broadcasting both channels. If it is not possible to build a second antenna for the second transmitter due to space constraints, then the diplexer is used permanently.

Several structures have been proposed for diplexers such as microstrip diplexers used bandpass and bandstop filters, diplexer used two band-pass filters, diplexers based on the slow-wave open-loop resonator with high impedance meander line , coupled folded-line resonators, and cross-coupled stepped impedance resonators (SIRs) and microstrip diplexer combined with only three simple coupled resonators. In this project a new structure is proposed for diplexers.

The proposed structure contains only two transmission lines and four stubs with the same characteristic impedances. Despite the simplicity of the proposed diplexer, a good transmission performances and high isolation between the output ports are achieved. The proposed diplexer is introduced, analysed and simulated in ADS software. Finally, the usefulness of the proposed diplexer is verified by fabricating and taking measurement at frequencies 2.4 and 2.85 GHz.

1.2 Diplexer Design process

To design microstrip diplexer a schematic process has been described here, the process can be used as a guide to the designer to obtain their strategy and getting a good result as required.

Initially define two frequencies at which Diplexer has to operate.

By defining frequency, design two microstrip band pass filter which provide maximum gain for one individual frequency.

Calculate input impedance and S parameter for the two BPF.

Connect these two filters to the main transmission line.

Then define three ports as one input and two outputs.

Simulate design in ADS tool. Then go for fabrication.

Analyse both simulated and fabricated results.

1.3 Motivation and Problem Statement

Designing of receiver section always need a device which can separate Channels of different frequency, and also can provide a good isolation between the ports, Diplexers are one of the important components in microwave circuits for channel separation in communication systems. They are commonly using two Band-Pass Filters (BPFs) with a common input.
Some of the microstrip BPFs reported in the literature are based on using parallel-coupled transmission lines, symmetric EBG structure, dual behaviour resonators, periodic stubs, and coupled open-loop resonators [2]–[9].

In some applications, two frequencies of diplexer are very near to each other, necessitating the use of two high-order BPFs. A novel idea in these applications may be the use of two low order BPFs that have a null at centre frequency of one another. In this project, a new structure is proposed for these types of BPFs. The proposed structure contains some microstrip double stubs composed of two parallel non-equal length open ended stubs. A diplexer can be created by simply connecting two Double-Stub BPFs (DS-BPFs) at their inputs through two quarter-wavelength transmission lines.

1.4 Goals/Scope of present work

The main goals of this project are to get acquainted with microwave circuit designing using CAD tool like Agilent Design System. . The present work consists of designing of two low order microwave band pass filter using microstrip transmission line technology. The scope of this project lies on designing and implementation of other microwave device components. Like Diplexer, Triplexes and other channelization devices.

The next section describes the organization of chapters in the thesis.

1.5 Report Organization

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

Chapter 1: presents introduction to project, design process, discusses the motivation and problem statement, goal and scope of present work.

Chapter 2: literature review and the theory involved in the research work of this project have been presented in this chapter

Chapter 3: describes the component used in diplexer design. It contains detail information of component used and their mathematical formulation.

Chapter 4: presents the simulation analysis and summarizes detailed results of simulation.

The Final chapter of the thesis (Chapter 5) presents the conclusions and future aspects of this project. The significance and contribution of present work is summarized.

Chapter 2

Literature Review

2.1 Introduction

This chapter reviews several basic but important concepts that are necessary to comprehend the contents of this report. Here general discussion on filter and diplexer previous work has been presented. And at last introduction to the proposed diplexer is described.

2.2 Literature review

IN MODERN communication systems, filters and Diplexer are always playing important and essential roles. A filter is a hart of the diplexer design. Planar filters are particularly popular structures because they can be fabricated using printed circuit technology and are suitable for commercial applications due to their compact size and low-cost integration [1]. Moreover, planar filters using the structures of parallel-coupled and cross-coupled resonators are preferable and extensively used in communication systems because of their high practicality and high performance [2]–[9]. For a planar filter design, it is necessary to select proper resonator types since resonators are the basic components of a filter. To reduce the resonator size, several types of resonators such as the U-shaped hairpin resonators [4], [5], the open-loop resonators [6], [7], and the folded open-line resonators [9], [10] have been proposed to design different kinds of bandpass filters. However, these structures are too large in size. Among these popular resonators, the most frequently used is the stepped impedance resonator (SIR) because it was originally presented not only to reduce the resonator size, but also to control all of the resonant frequencies by properly adjusting its structural parameters [11], [12].

In multiservice and multiband communications, diplexers are often needed to have some capabilities of high compactness, light weight, and high isolation. Several layouts have been Proposed to satisfy these conditions. To reduce the circuit size, diplexers based on the slow-wave open-loop resonators with high-impedance meander lines [13], the folded coupled-line

Resonators [14], the miniaturized open-loop resonators [15], SIRs [16], and spiral inductor resonators [17] have been proposed.

Diplexer based on the stepped-impedance coupled-line resonator (SICR) [18] can offer a compact size and better isolation performance by equalizing velocities of the odd and even-mode waves in an inhomogeneous medium. In addition, some diplexers are designed to resolve the spurious response problem. The diplexer presented in [19] uses balanced open-circuited periodic stubs as a bandstop network to provide both low loss and high isolation between the channels, but increases the circuit size. In [20], the double-loop resonator has been presented; however, it is not practical for standard manufacturing. A photonic-band gap structure [21] consumes much development time and increases the losses.

In this project, high isolation and compact size microstrip diplexers have been proposed. The designed diplexers are based on combination of two filters, which are constructed with some common resonators sections. By properly locating the fundamental and the first spurious resonant frequencies of SIRs, they can be shared by both filter channels. Based on this guideline, the circuit size can be dramatically reduced. This concept has been verified by simulation results. Since the total number of resonators is reduced, these diplexers are extremely compact in size and are much smaller than those mentioned above.

The next chapter 3 introduces detail description about the basic components used in designing of microstrip Diplexer.

Chapter 3

Component description

3.1 Introduction

The focus of this chapter is to study the basic designing component for diplexer. Diplexer is designed with microstrip lines, so here in this chapter first basic information about transmission lines is introduced in section 3.2, and then section 3.3, 3.4, 3.5 introduces about stub lines, microstrip line and bandpass filter. At last section 3.6 description about Diplexer design is introduced. Here, the basic equations governing scattering parameter are obtained.

3.2 .1 Transmission line Model

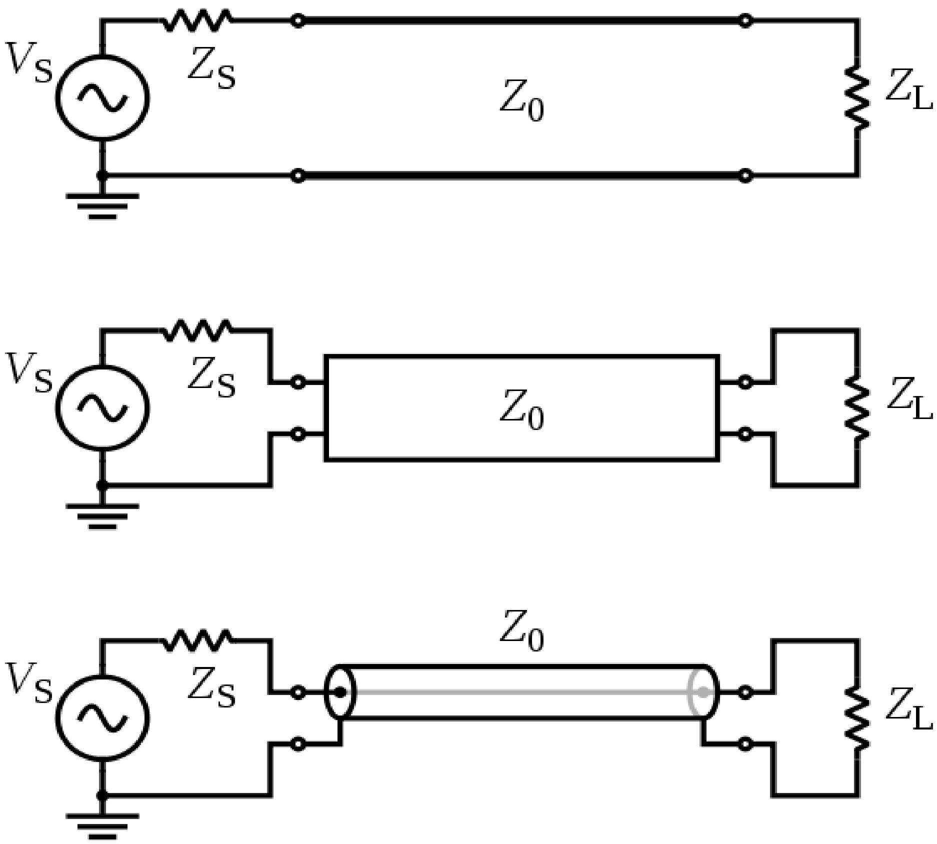

Figure 3.1 Schematic representations of the elementary components of a transmission line

Transmission line is a closed system in which power is transmitted from a source to a destination as shown in figure 3.1. Coaxial cable, waveguide, or other system of conductors that transfers electrical signals from one location to another, are examples of transmission line.

Applying the transmission line model based on the telegrapher's equations, the general expression for the characteristic impedance of a transmission line is:

$$Z_0 = \sqrt{\frac{(\mathrm{R}+\mathrm{j}\omega \mathrm{L})}{(\mathrm{G}+\mathrm{j}\omega C)}} \qquad \text{.......... (1)}$$

Where,

R , is the resistance per unit length,

L ,is the inductance per unit length,

G , is the conductance of the dielectric per unit length,

C , is the capacitance per unit length,

j ,is the imaginary unit, and

ω , is the angular frequency.

The voltage and current phasors on the line are related by the characteristic impedance as:

$$\frac{V^+}{I^+} = Z_0 = -\frac{V^-}{I^-} \qquad \text{............ (2).}$$

Where, the superscripts $+$ and $-$ represent forward and backward-travelling waves, respectively.

3.2.2 Telegrapher's Equations

The **Telegrapher's Equations** (or just **Telegraph Equations**) are a pair of linear differential equations which describe the voltage and current on an electrical transmission line with distance and time. They were developed by Oliver Heaviside who created the *transmission line model* based on Maxwell's Equations.

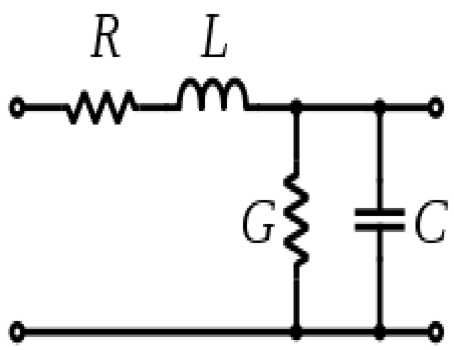

Figure 3.2 Schematic representation of the elementary component of a transmission line.

The transmission line model represents the transmission line as an infinite series of two-port elementary components, each representing an infinitesimally short segment of the transmission line:

The distributed resistance R of the conductors is represented by a series resistor (expressed in ohms per unit length).

The distributed inductance L (due to the magnetic field around the wires, self-inductance, etc.) is represented by a series inductor (henries per unit length).

The capacitance C between the two conductors is represented by a shunt capacitor C (farads per unit length).

The conductance G of the dielectric material separating the two conductors is represented by a shunt resistor between the signal wire and the return wire (Siemens per unit length).

The model consists of an *infinite series* of the elements shown in the figure 3.2, and that the values of the components are specified *per unit length* so the picture of the component can be misleading. R, L, C, and G may also be functions of frequency. An alternative notation is to use R’, L’, C’ and G,’ to emphasize that the values are derivatives with respect to length. These quantities can also be known as the primary line constants to distinguish from the secondary line constants derived from them, these being the propagation constant, attenuation constant and phase constant.

The line voltage $V(\mathrm{x})$ and the current $\mathrm{I(x)}$ can be expressed in the frequency domain as

$$\frac{\partial V(\mathrm{x})}{\partial\,\mathrm{x}} = -(\mathrm{R} + \mathrm{j}\omega\,\mathrm{L})\,\mathrm{I(x)} \qquad \text{.............(3)}$$

$$\frac{\partial I(\mathrm{x})}{\partial\,\mathrm{x}} = -(\mathrm{G} + \mathrm{j}\omega C)\,\mathrm{V(x)} \qquad \text{.............(4)}$$

When the elements R and G are negligibly small the transmission line is considered as a lossless structure. In this hypothetical case, the model depends only on the L and C element which greatly simplifies the analysis. For a lossless transmission line, the second order steady-state Telegrapher's equations are:

$$\frac{\partial^2\,\mathrm{V(x)}}{\partial\,\mathrm{x}^2} + \omega^2 LC \cdot V(\mathrm{x}) = 0 \qquad \text{.............(5)}$$

$$\frac{\partial^2 I(\mathrm{x})}{\partial\,\mathrm{x}^2} + \omega^2 LC \cdot I(\mathrm{x}) = 0 \qquad \text{..............6)}$$

Equation 5 and 6 are wave equations which have plane waves with equal propagation speed in the forward and reverse directions as solutions. The physical significance of this is that electromagnetic waves propagate down transmission lines and in general, there is a reflected component that interferes with the original signal. These equations are fundamental to transmission line theory.

If R and G are not neglected, the Telegrapher's equations become:

$$\frac{\partial^2\,\mathrm{V(x)}}{\partial\,\mathrm{x}^2} = \gamma^2 V(\mathrm{x}) \qquad \text{.............(7)}$$

$$\frac{\partial^2\,\mathrm{I(x)}}{\partial\,\mathrm{x}^2} = \gamma^2 \cdot \mathrm{I(x)} \qquad \text{..............(8)}$$

Where,

$$\gamma = \sqrt{(\mathrm{R} + \mathrm{j}\omega \mathrm{L})(\mathrm{G} + \mathrm{j}\omega C)} \qquad \text{.............(9)}$$

And the characteristic impedance is:

$$Z_0 = \sqrt{\frac{(\mathrm{R} + \mathrm{j}\omega \mathrm{L})}{(\mathrm{G} + \mathrm{j}\omega C)}} \qquad \text{..........(10)}$$

The solutions for $V(\mathrm{x})$ and $\mathrm{I(x)}$ are:

$$\mathrm{V(x)} = \mathrm{V}^{+}\mathrm{e}^{-\gamma x} + \mathrm{V}^{-}\mathrm{e}^{\gamma x} \qquad \text{..........(11)}$$

$$\mathrm{I(x)} = \frac{1}{Z_0}(\mathrm{V}^{+}\mathrm{e}^{-\gamma x} - \mathrm{V}^{-}\mathrm{e}^{\gamma x}) \qquad \text{...........(12)}$$

3.2.3 Characteristic Impedance

The characteristic impedance or surge impedance of a uniform transmission line, usually written Z_0, is the ratio of the amplitudes of a single pair of voltage and current waves propagating along the line in the absence of reflections. SI unit of characteristic impedance is ohm.

The characteristic impedance of a lossless transmission line is purely real, that is, there is no imaginary component ($(Z_0 = |Z_0| + j0)$). Characteristic impedance appears like a resistance in this case, such that power generated by a source on one end of an infinitely long lossless transmission line is transmitted through the line but is not dissipated in the line itself.

A transmission line of finite length (lossless or lossy) that is terminated at one end with a resistor equal to the characteristic impedance ($Z_L = Z_0$) appears to the source like an infinitely long transmission line.

3.2.4 Input Impedance of Lossless Transmission Line

For a lossless transmission line, it can be shown that the impedance measured at a given position l from the load impedance Z_L is

$$Z_{in}(\mathrm{l}) = Z_0 \frac{Z_L + jZ_0(\beta \mathrm{l})}{Z_0 + jZ_L(\beta \mathrm{l})} \qquad \text{...........(13)}$$

Where, $\beta = \dfrac{2\pi}{\lambda}$ Phase constant.

β, is generally different inside the transmission line to what it would be in free-space.

3.2.5 Half Wave Length Transmission Line

For the special case where $\beta l = n\pi$ where n is an integer (meaning that the length of the line is a multiple of half a wavelength), the expression reduces to the load impedance so that

$$Z_{in} = Z_L \qquad \text{.........(14)}$$

For all n. This includes the case when $n = 0$, meaning that the length of the transmission line is negligibly small compared to the wavelength. The physical significance of this is that the transmission line can be ignored (i.e. treated as a wire) in either case.

3.2.6 Quarter Wave Length Transmission Line

For the case where the length of the line is one quarter wavelength long, or an odd multiple of a quarter wavelengths long, the input impedance becomes

$$Z_{in} = \frac{Z_0^2}{Z_L} \qquad \text{..........(15)}$$

3.2.7 Matched Load Line

Another special case is when the load impedance is equal to the characteristic impedance of the line (i.e. the line is *matched*), in which case the impedance reduces to the characteristic impedance of the line so that

$$Z_{in} = Z_L = Z_0 \qquad \text{............(16)}$$

For all l and λ.

3.2.8 Scattering Parameters:

The scattering or *S* parameters of a two-port network are defined in terms of the wave variables

$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2=0} \qquad S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1=0}$$

$$S_{21} = \left.\frac{b_2}{a_1}\right|_{a_2=0} \qquad S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1=0} \qquad \text{.......... (17.a)}$$

These definitions may be written in the matrix form as

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \cdot \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \qquad \text{..........(17.b)}$$

Where, the matrix containing the S parameters is referred to as the scattering matrix or S matrix, which may simply be denoted by [*S*].

The parameters S_{11} and S_{22} are also called the reflection coefficients, whereas S_{12} and S_{21} the transmission coefficients. These are the parameters directly measurable at microwave frequencies. The *S* parameters are in general complex, and it is convenient to express them in terms of amplitudes and phases, i.e.,

$$20\log|S_{mn}|\,dB \quad \text{m,n=1,2} \qquad \text{...........(18)}$$

Where, the logarithm operation is base 10.. For filter characterization, we may define two parameters

$$L_A = -20\log|S_{mn}|\,dB \quad \text{m,n =1,2(m} \neq \text{n)} \qquad \text{...........(19)}$$

$$L_R = 20\log|S_{mn}|\,dB \quad \text{n=1,2} \qquad \text{.........(20)}$$

Where, L_A denotes the insertion loss between ports *n* and *m* and L_R represents the return loss at nth port.

3.3 Stub description

In microwave and radio frequency engineering a **stub** is the transmission line or waveguide that is connected at one end only.

The free end of the stub is either left open-circuit or short-circuited (especially in the case of waveguides). Neglecting transmission line losses, the input impedance of the stub is purely reactive; either capacitive or inductive, depending on the electrical length of the stub, and on whether it is open or short circuit.

Stubs may thus be considered to be frequency-dependent capacitors and frequency-dependent inductors, because stubs take on reactive properties as a function of their electrical length, stubs are most common in UHF or microwave circuits where the line lengths are more manageable. Stubs are commonly used in antenna impedance matching circuits and frequency selective filters. Smith charts can also be used to determine length of the stub to obtain a desired reactance.

3.3.1 Short Circuited Stub

The input impedance of a lossless short circuited line is,

$$Z_{SC} = jZ_0 \tan(\beta l) \qquad \text{...........(21)}$$

Where, j is the imaginary unit, Z_0 is the characteristic impedance of the line, β is the phase constant of the line, and l is the physical length of the line.

Thus, depending on whether $\tan(\beta l)$ is positive or negative, the stub will be inductive or capacitive, respectively.

Length of a stub to act as a capacitor *C* at an angular frequency of ω is then given by:

$$l = \frac{1}{\beta}\left[(\mathrm{n}+1)\pi - \arctan\left(\frac{1}{\omega C Z_0} \right) \right] \qquad \text{...........(22)}$$

The length of a stub to act as an inductor *L* at the same frequency is given by:

$$l = \frac{1}{\beta}\left[\mathrm{n}\,\pi + \arctan\left(\frac{\omega l}{Z_0} \right) \right] \qquad \text{...........(23)}$$

3.3.2 Open Circuited stub

The input impedance of a lossless open circuit stub is given by

$$Z_{OC} = -jZ_0 \cot(\beta l) \qquad \text{...........(24)}$$

It follows that whether, $\cot(\beta l)$ is positive or negative, the stub will be capacitive or inductive, respectively.

Length of an open circuit stub to act as an Inductor *L* at an angular frequency of ω is:

$$l = \frac{1}{\beta}\left[(\mathrm{n}+1)\pi - \mathrm{arccot}\left(\frac{\omega l}{Z_0} \right) \right] \qquad \text{.........(25)}$$

The length of an open circuit stub to act as a capacitor *C* at the same frequency is:

$$l = \frac{1}{\beta}\left[n\pi + \text{arccot}\left(\frac{1}{\omega C Z_0} \right) \right] \qquad \text{...........(26)}$$

3.3.3 Resonant Stub:

Stubs are often used as resonant circuits in distributed element filters. An open circuit stub of length l will have a capacitive impedance at low frequency when $\beta l \leq \pi/2$. Above this frequency the impedance is inductive. At precisely $\beta l = 3\pi/2$ the stub presents a short circuit. This is qualitatively the same behaviour as a series resonant circuit. For a lossless line the phase change constant is proportional to frequency,

$$\beta = \frac{\omega}{v} \qquad \text{..........(27)}$$

Where, *v* is the velocity of propagation and is constant with frequency for a lossless line. For such a case the resonant frequency is given by,

$$\omega_0 = \frac{\pi v}{2l} \qquad \text{...........(28)}$$

ω_0 is qualitatively like a resonant frequency of lumped element resonant circuit. But the impedance function is not precisely the same quantitatively. In particular, the impedance will not continue to rise monotonically with frequency after resonance.

It will rise until the point where $\beta l = \pi$ where it will be open circuit. After this point (which is actually an anti-resonance point) the impedance will again become capacitive and start to fall. It will continue to fall until at $\beta l = 3\pi/2$ where it presents a short circuit.

At this point filtering action of the stub has totally failed. This response of the stub continues to repeat with increasing frequency alternating between resonance and anti-resonance. It is not only a characteristic of stubs, but of all distributed element filters, that there is some frequency beyond which the filter fails and multiple unwanted pass bands are produced.

Similarly, a short circuit stub is an anti-resonator at $\pi/2$, that is, it behaves as a parallel resonant circuit, but again fails as $3\pi/2$ is approached.

3.3.4 Stub Matching:

Stubs can be used to match load impedance to the transmission line characteristic impedance. The stub is positioned a distance from the load. This distance is selected in such a manner that the resistive part of load impedance is made equal to the resistive part of characteristic impedance at that point by impedance transformer.

The length of the stub is so chosen that it exactly cancels the reactive part of the presented impedance. That is, the stub is made capacitive or inductive according to whether the main line is presenting inductive or capacitive impedance respectively.

This is not the same as the actual impedance of the load since the reactive part of the load impedance will be subject to impedance transformer action as well as the resistive part. Matching stubs can be made adjustable so that matching can be corrected on test.

A single stub will only achieve a perfect match at one specific frequency. For wideband matching several stubs may be used spaced along the main transmission line. The resulting structure is filter-like and filter design techniques are applied.

For instance, the matching network may be designed as a Chebyshev filter but is optimized for impedance matching instead of passband transmission. The resulting transmission function of the network has a passband ripple like the Chebyshev filter, but the ripples never reach 0dB insertion loss at any point in the passband, as they would do for the standard filter.

3.3.5 Richard's Transformation:

Richards' transformation, a close correspondence exists between lumped inductors and capacitors in the *p*-plane and short- and open-circuited transmission lines in the *t*-plane.

As a one-port inductive element with an impedance $Z = pL$, a lumped inductor corresponds to a short-circuited line element (stub) with an input impedance $Z = tZc = jZc \tan \theta$, where Zc is the characteristic impedance of the line.

A lumped capacitor with an admittance $Y = pC$ corresponds to an open circuited stub of input admittance $Y = tYc = jYc \tan \theta$ and characteristic admittance Yc.

This transformation is illustrated in figure 3.3 and as a consequence the short-circuited and open-circuited line elements are sometimes referred to as the t plane inductor and capacitor, respectively, and use the corresponding lumped-element symbols as well.

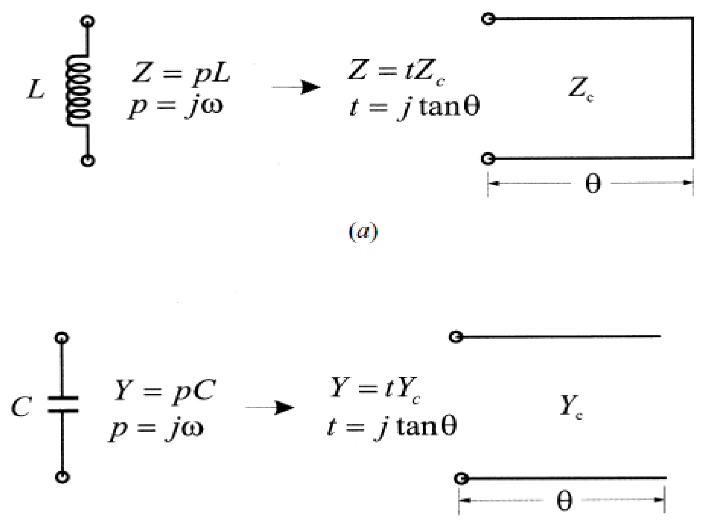

Figure 3.3 Lumped and distributed element correspondence under Richards' transformation [1]

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3.3.6 Kuroda Identities

The Kuroda identities shown in Figure 3.4 and 3.5, form a basis to achieve such transformations, where the commensurate line elements with the same electrical length are assumed for each identity.

The first two Kuroda identities interchange a unit element with a shunt open-circuited stub or a series short-circuited stub, and a unit element with a series short-circuited stub or a shunt open-circuited stub. The other two Kuroda identities, involving the ideal transformers, interchange stubs of the same kind.

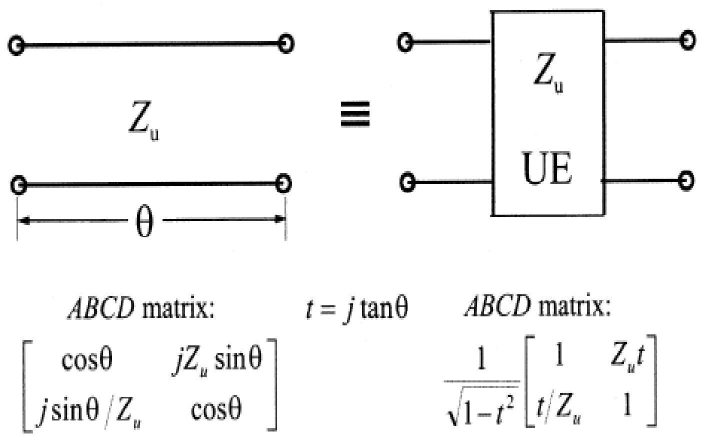

Figure 3.4 Unit elements (UE).

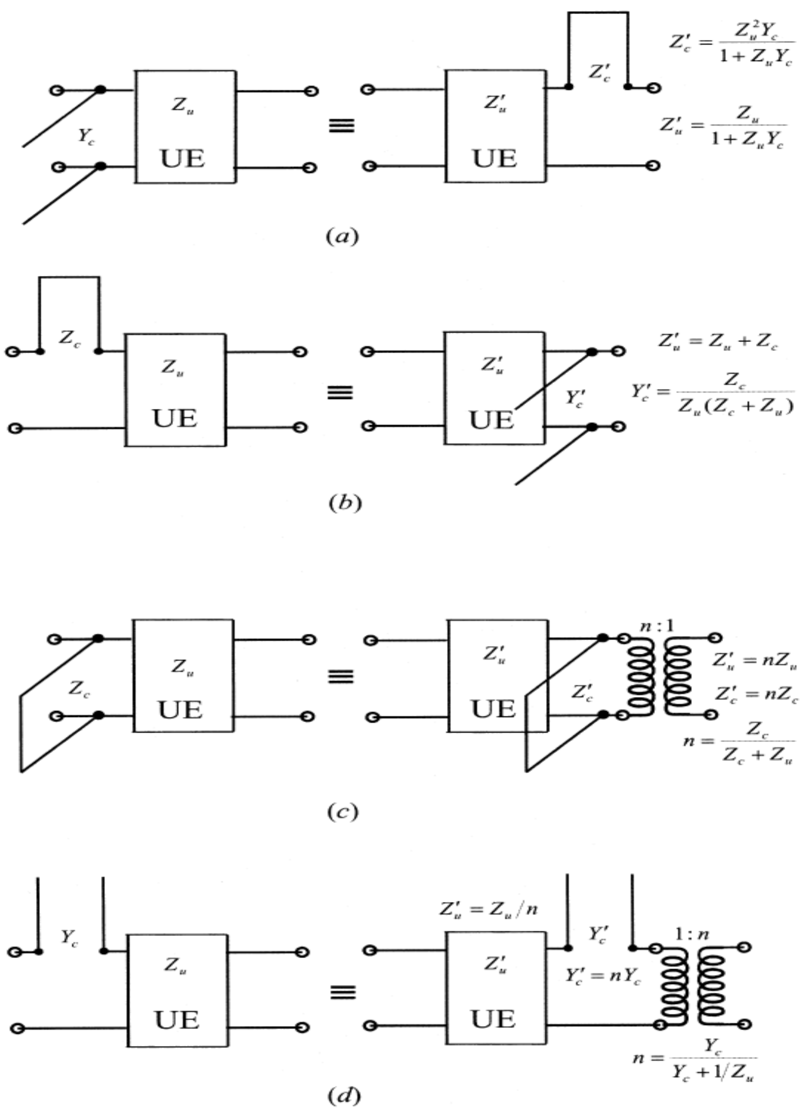

Figure 3.5 Kuroda identities [1]

3.4 Microstrip Line:

Microstrip is a type of electrical transmission line which can be fabricated using printed circuit board technology, and is used to convey microwave-frequency signals. It consists of a conducting strip separated from a ground plane by a dielectric layer known as the substrate as shown in figure 3.6. Microwave components such as antennas, couplers, filters, power dividers etc. can be formed from microstrip, the entire device existing as the pattern of metallization on the substrate. Microstrip is less expensive than traditional waveguide technology, as well as being far lighter and more compact.

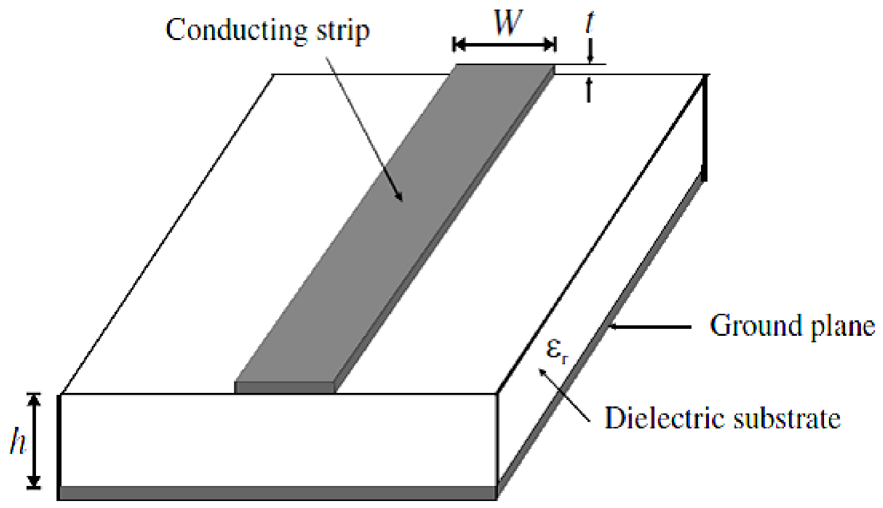

Figure 3.6 Schematic diagram for microstrip line

The **disadvantages of microstrip** compared with waveguide are the generally lower power handling capacity, and higher losses. Also, unlike waveguide, microstrip is not enclosed, and is therefore susceptible to cross-talk and unintentional radiation. For lowest cost, microstrip devices may be built on an ordinary FR-4 (standard PCB) substrate. However it is often found that the dielectric losses in FR4 are too high at microwave frequencies, and that the dielectric constant is not sufficiently tightly controlled. For these reasons, an alumina substrate is commonly used.

On a smaller scale, microstrip transmission lines are also built into monolithic microwave integrated circuits. Microstrip lines are also used in high-speed digital PCB designs, where signals need to be routed from one part of the assembly to another with minimal distortion, and avoiding high cross-talk and radiation.

Microstrip is very similar to stripline and coplanar waveguide, and it is possible to integrate all three on the same substrate.

3.4.1 Characteristic Impedance of Micro strip Line:

The characteristic impedance Z_0 of microstrip is a function of the ratio of the height to the width W/H (and ratio of width to height H/W) of the transmission line, and also has separate solutions depending on the value of W/H. the characteristic impedance Z_0 of microstrip is calculated by:

When, $\left(\frac{W}{H}\right) \leq 1$

$$Z_0 = \frac{60}{\varepsilon_{eff}} \ln\left(8\frac{W}{H} + 0.25\frac{W}{H}\right) \qquad \text{........ (29)}$$

and when $\left(\frac{W}{H}\right) \geq 1$

$$Z_0 = \frac{120\pi}{\sqrt{\varepsilon_{eff}} \times \left[\frac{W}{H} + 1.393 + \frac{2}{3}\ln\left(\frac{W}{H} + 1.444\right)\right]} \qquad \text{......... (30)}$$

3.4.2 Effective Dielectric Constant:

Because part of the fields from the microstrip conductor exist in air, the effective dielectric constant is somewhat less than the substrate's dielectric constant , the effective dielectric constant ε_{eff} of microstrip is calculated by:

When $\left(\frac{W}{H}\right) \leq 1$

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[\left(1 + 12\left(\frac{W}{H}\right)\right)^{-1/2} + 0.04\left(1 - \left(\frac{W}{H}\right)\right)^2\right] \qquad \text{.......... (31)}$$

also $\left(\frac{W}{H}\right) \geq 1$

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + 12\left(\frac{W}{H}\right)\right)^{-1/2} \qquad \text{........... (32)}$$

The effective dielectric constant is a function of width to height ratio of the microstrip line (W/H), as well as the dielectric constant of the substrate material.

3.4.3 Guided Wavelength, Propagation Constant, Phase Velocity, and Electrical Length:

Once the effective dielectric constant of a microstrip is determined, the guided wavelength of the quasi-TEM mode of microstrip is given by

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_{eff}}} \qquad \text{.............(33)}$$

Where, λ_0 is the free space wavelength at operation frequency *f*. More conveniently, where the frequency is given in gigahertz (GHz), the guided wavelengths can be evaluated directly in millimetres as follows:

$$\lambda_g = \frac{300}{f\,(\mathrm{GHz})\sqrt{\varepsilon_{eff}}} \qquad \text{......... (34)}$$

The associated propagation constant and phase velocity v_p can be determined by

$$\beta = \frac{2\pi}{\lambda_g} \quad \text{............. (35)}$$

$$v_p = \frac{\omega}{\beta} = \frac{c}{\sqrt{\varepsilon_{eff}}} \quad \text{............(36)}$$

Where, c is the velocity of light ($c = 3.0 \times 10^8$ m/s) in free space The electrical length for a given physical length l of the microstrip is defined by

$$\theta = \beta l \quad \text{.............(37)}$$

Therefore $\theta = \pi/2$ when $l = \lambda_g/4$, and $\theta = \pi$ when $l = \lambda_g/2$. These so-called quarter wavelength and half-wavelength microstrip lines are important for design of microstrip filters.

3.4.4 Microstrip Losses:

The losses of microstrip line include conductor loss, dielectric loss and radiation loss. The propagation constant on a lossy transmission line is complex; namely, $\gamma = \alpha + j\beta$, where the real part α in nepers per unit length is the attenuation constant, which is the sum of the attenuation constants arising from each effect.

In practice, α may prefer to express in decibels (dB) per unit length, which can be related

$$\alpha_c \text{ (dB/unit length)} = (20 \log_{10} e)\alpha \text{ (Nepers/unit length)} \quad \text{........(38)}$$

$$\approx 8.8686\alpha \text{ (Nepers/unit length)}$$

3.4.4.1 Conductor loss:

A simple expression for the estimation of the attenuation produced by the conductor loss is given by

$$\alpha_c = \frac{8.686 R_s}{Z_c W} \quad \text{dB/Unit length} \qquad ....(39)$$

In which Zc is the characteristic impedance of the microstrip of the width *W*, and *Rs* represents the surface resistance in ohms per square for the strip conductor and ground plane. For a conductor

$$R_s = \sqrt{\frac{\omega \mu_0}{2\sigma}} \qquad ....(40)$$

where σ is the conductivity, μ_0 is the permeability of free space, and ω is the angular frequency.

3.4.4.2 Dielectric loss:

The attenuation due to the dielectric loss in microstrip can be determined by

$$\alpha_d = 8.686\pi \left(\frac{\varepsilon_{eff} - 1}{\varepsilon_r - 1} \right) \frac{\varepsilon_r}{\varepsilon_{eff}} \frac{\tan\delta}{\lambda_g} \quad \text{dB/unit length} \qquad ....(41)$$

where tan δ denotes the loss tangent of the dielectric substrate.Since the microstrip is a semiopen structure, any radiation is either free to propagate away or to induce currents on the metallic enclosure, causing the radiation loss or the so-called housing loss.

3.5.1 Band-pass filter and Band-stop filter

A band-pass filter is a device that passes frequencies within a certain range and rejects (attenuates) frequencies outside that range.

A band-stop filter or band-rejection filter is a filter that passes most frequencies unaltered, but attenuates those in a specific range to very low levels. It is the opposite of a band-pass filter. A notch filter is a band-stop filter with a narrow stop band (high Q factor). Frequency response of band stop and band pass filter is shown in figure 3.7.

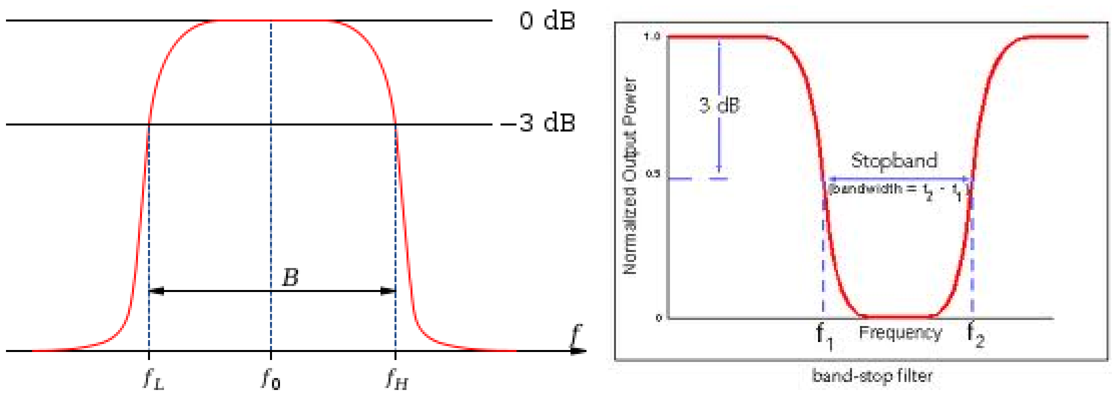

Figure 3.7 (a) Band pass filter (b) Band stop filter

3.5.2 LC Circuits:

The simplest band pass filters are LC filters. They use inductors and capacitors ((through always there is some extra resistance in circuit which affects the operation).these components are connected in series or parallel with each other; the resulting circuits are called series resonant are parallel resonant circuits, respectively.

The word resonant is used because these circuits respond to particular frequencies, much like the strings on a violin or guitar .For the reason; they are also often called as tuned circuits.

3.5.2.1 Bandwidth:

The bandwidth of the filter is simply the difference between the upper and lower cut off frequencies. The shape factor is the ratio of bandwidths measured using two different attenuation values to determine the cut-off frequency, e.g., a shape factor of 2:1 at 30/3 dB means the bandwidth measured between frequencies at 30 dB attenuation is twice of that measured between frequencies at 3 dB attenuation.

3.5.2.2 Quality Factor:

Q is defined in terms of the ratio of the energy stored in the resonator to the energy supplied by a generator, per cycle, to keep signal amplitude constant, at a frequency (the resonant frequency), f_r, where the stored energy is constant with time:

$Q = 2\pi \times$ Energy stored/Energy dissipated per cycle

Or $Q = 2\pi f_r \times$ Energy stored / Power Loss

The factor 2π makes Q expressible in simpler terms, involving only the coefficients of the second-order differential equation describing most resonant systems, electrical or mechanical. In electrical systems, the stored energy is the sum of energies stored in lossless inductors and capacitors; the lost energy is the sum of the energies dissipated in resistors per cycle. In mechanical systems, the stored energy is the maximum possible stored energy, or the total energy, i.e. the sum of the potential and kinetic energies at some point in time; the lost energy is the work done by an external conservative force, per cycle, to maintain amplitude.

For high values of Q, the following definition is also mathematically accurate:

$$Q = \frac{f_r}{\Delta f} = \frac{\omega_r}{\Delta\omega} \qquad \text{............(42)}$$

Where, f_r is the resonant frequency, Δf is the bandwidth, $\omega_r = 2\pi f_r$ is the angular resonant frequency, and $\Delta\omega$ is the angular bandwidth.

3.5.3 Series Resonance: Circuit diagram and frequency response of series resonant are shown in figure 3.8 and 3.9 respectively

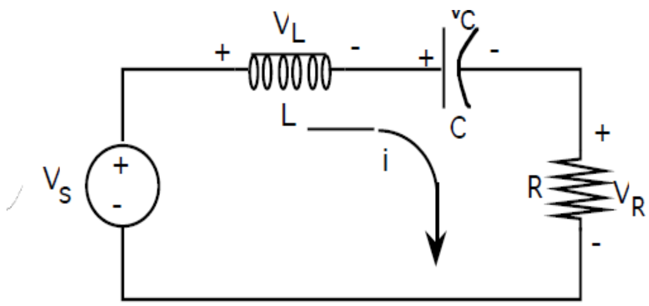

Figure 3.8 the Series Resonant Circuit

Resonance frequency, $$\omega_0 = \frac{1}{\sqrt{LC}}$$ ....(43)

The bandwidth of the series circuit is defined as the range of frequencies in which the amplitude of the current is equal to or greater than (1 /√ 2 = √2 / 2) times its maximum amplitude

Bandwidth, $$B = \omega_2 - \omega_1 = R/L$$ ....(44)

and quality factor $$Q = \frac{\omega_0}{B} = \frac{1}{R}\sqrt{\frac{C}{L}}$$ ....(45)

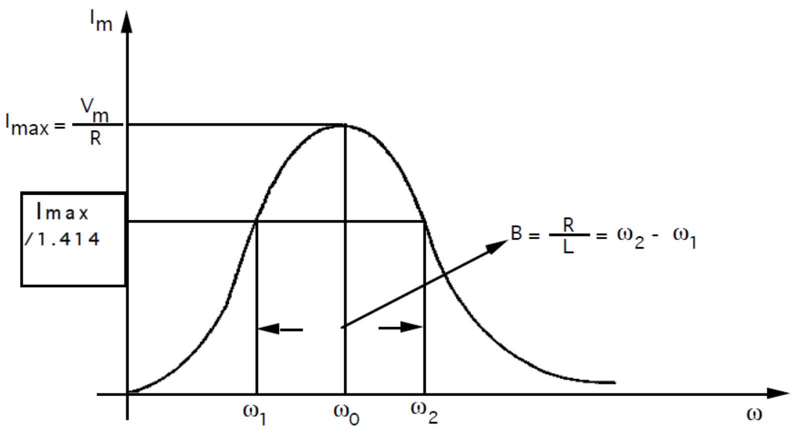

Figure 3.9 Frequency Response of a Series - Resonant Circuit

3.5.4 Parallel Resonance: Circuit diagram and frequency response of parallel resonant are shown in figure 3.10 and 3.11 respectively.

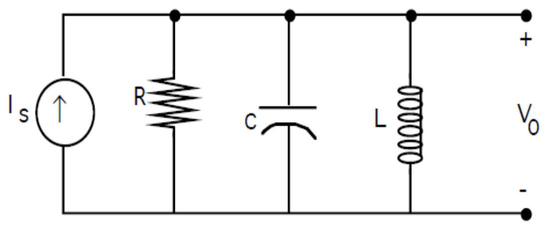

Figure 3.10 The Parallel Resonant circuit

The resonant frequency is,

$$\omega_0 = \frac{1}{\sqrt{LC}} \quad ....(46)$$

The bandwidth, $$B = \omega_2 - \omega_1 = 1/RC \quad ....(47)$$

The quality factor, $$Q = \frac{\omega_0}{B} = R\sqrt{\frac{C}{L}} \quad ....(48)$$

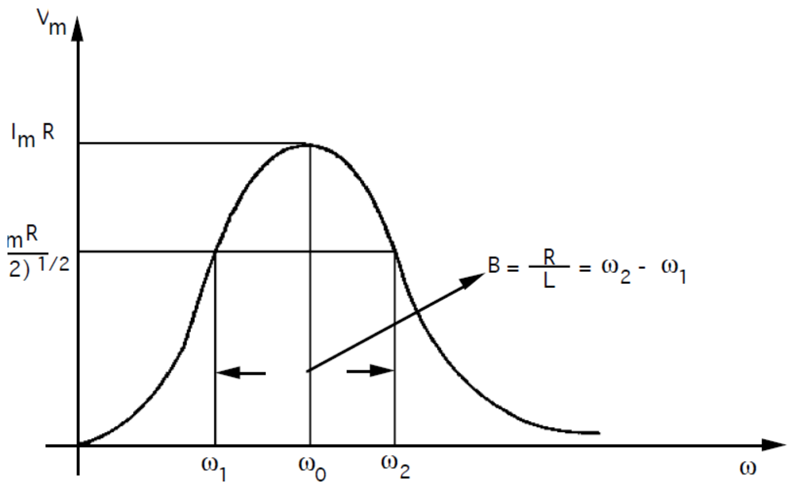

Figure 3.11 Frequency Response of the Parallel - Resonant Circuit

3.6.1 Diplexers

It's a three-port network that splits incoming signals from a common port into two paths (sometimes called "channels"), dependent on frequency. A diplexer is the simplest form of a multiplexer, which can split signals from one common port into many different paths. The incoming signals must be offset in frequency by an appreciable percentage so that filters can do their job sorting them out.

A diplexer could be used to route signals to two different receivers, based on frequency. Or it could be used to create a "matched" filter that is non-reflective outside of the intended pass band. It could also be used as a bias tee, to feed active device with DC power.

A **diplexer** is a passive device that implements frequency domain multiplexing. Two ports (e.g., L and H) are multiplexed onto a third port (e.g., S). The signals on ports L and H occupy disjoint frequency bands. Consequently, the signals on L and H can coexist on port S without interfering with each other.

Typically, the signal on port L will occupy a single low frequency band and the signal on port H will occupy a higher frequency band. In that situation, the diplexer consists of a low pass filter connecting ports L and S and high pass filter connecting ports H and S. Ideally, all the signal power on port L is transferred to the S port and vice versa. All the signal power on port H is transferred to port S and vice versa. In this way separation of the signals is complete. None of the low band signal is transferred from the S port to the H port. In the real world, some power will be lost, and some signal power will leak to the wrong port.

Television diplexer consisting of a high-pass filter (left) and a low-pass filter (right). The antenna is connected to the screw terminals to the left of centre.

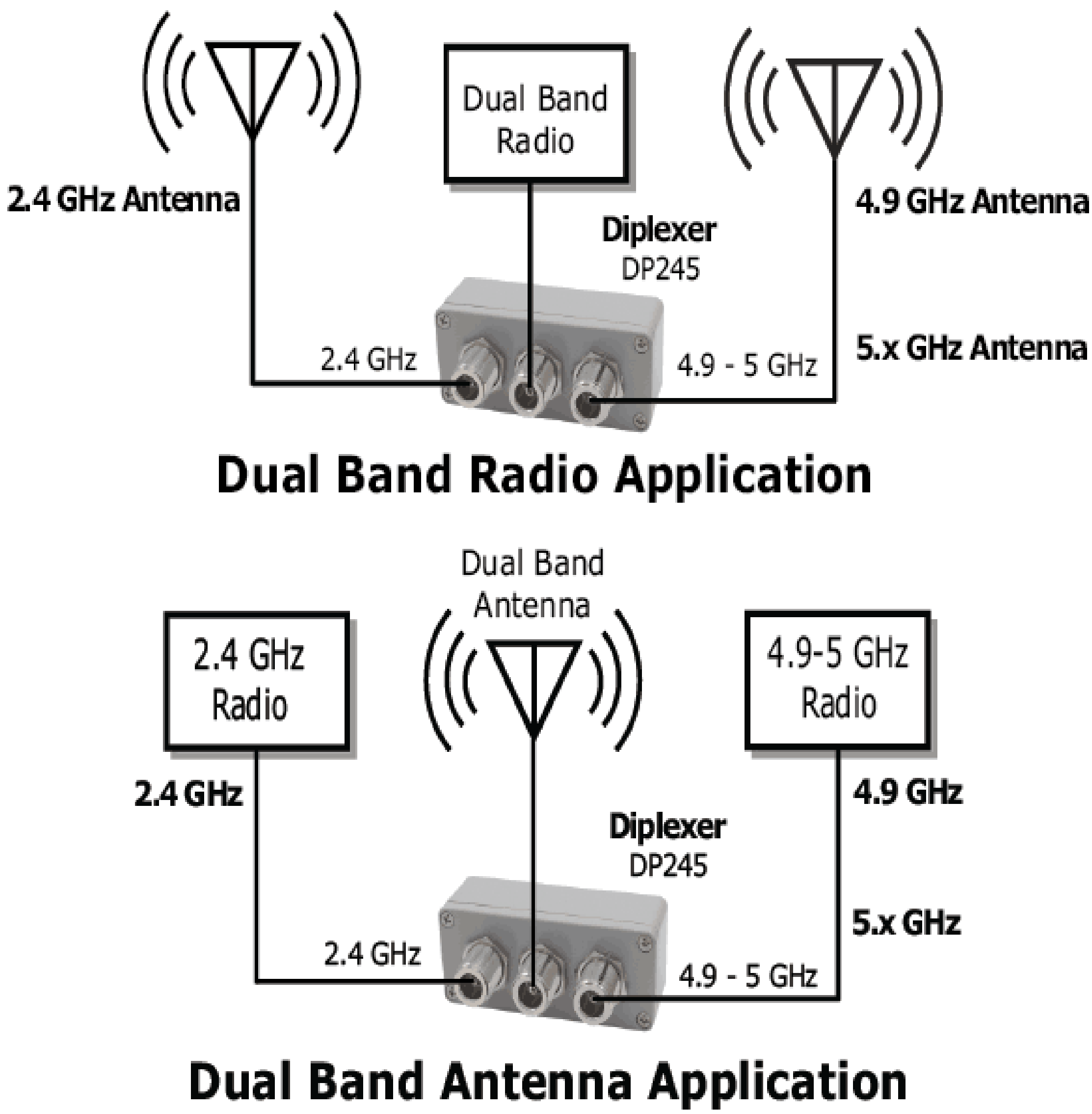

Figure 3.12 diplexers system

Here figure 3.12 shows a diplexer system, first figure shows a transmitter terminal where two signal of frequencies 2.4GHz and 4.9GHz are feeded from a radio device and then transmitted separately using two antennas. While second one show a receiver section where two signals of frequencies 2.4GHz and 4.9GHz are received by a dual band antenna and then separated in two different terminal of diplexer making diplexer to work as a frequency separator

3.6.2 Design and Formulation:

Diplexers are three port devices which are commonly used in wide-band or multi-frequency antennas in transceiver applications. a diplexer take normally two frequencies into their input port and separate them to two output ports.

Several structures have been proposed for diplexers such as microstrip diplexers which uses bandpass and bandstop filters. Diplexer using two band-pass filters are based on the slow-wave open-loop resonator with high impedance meander line, coupled folded-line resonators, and cross-coupled stepped impedance resonators (In this Project, a new structure is proposed for diplexers.

The proposed structure contains only two transmission lines and four stubs with the same characteristic impedances. Despite the simplicity of the proposed diplexer, a good transmission performances and high isolation between the output ports are achieved. The proposed diplexer is introduced and analysed. Finally, the usefulness of the proposed diplexer is verified by designing, fabrication and measurement of a microstrip diplexer at frequencies 2.4 and 2.85 GHz.

3.6.2.1 Design Theory

Figure 3.13 depicts the proposed microstrip diplexer consisting of two transmission lines, two open circuit stubs and two short circuit stubs, whose characteristic impedances are the same as Z_0. In this figure, λ_{g1} and λ_{g2} are the wavelengths in microstrip medium at two working frequencies f_1 and f_2, where f_2 is assumed greater than f_1 and also less than 3 f_1 to assure the length of the lower short circuit stub to be positive.

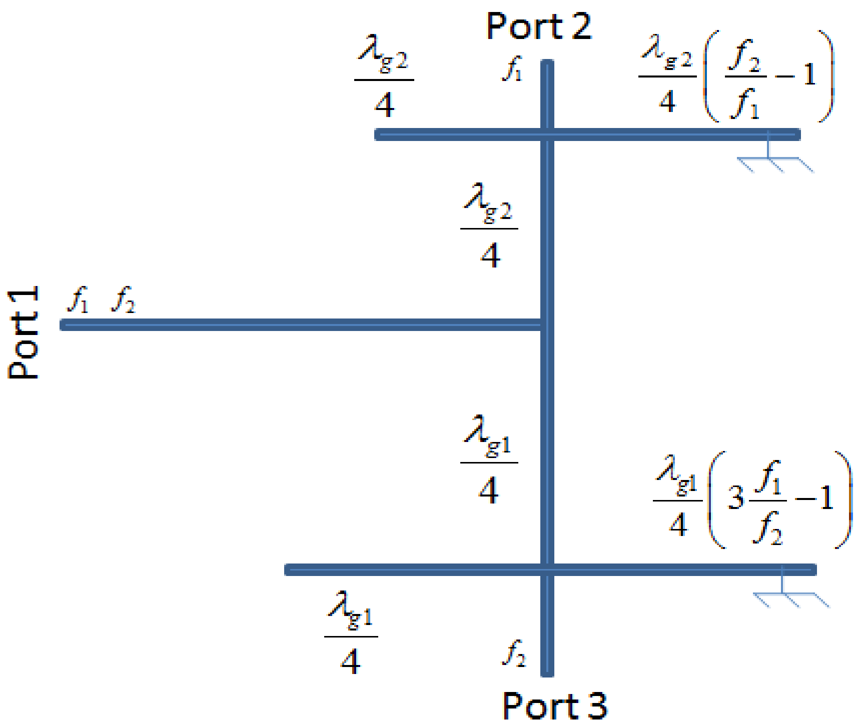

Figure 3.13 Diplexer design

The input signal at Port 1 is composed of two frequencies f_1 and f_2. The output signal at Port 2 has only f_1 frequency component of the input signal as well as the output signal at Port 3 has only the frequency f_2 component. Here, the behaviour of the proposed diplexer is explained qualitatively. The pair of short and open circuit stubs at the upper branch of the proposed diplexer yield infinite and zero impedance, respectively, at frequencies f_1 and f_2.

On the other hand, the pair of short and open circuit stubs at the lower branch of the proposed diplexer yield infinite and zero impedance, respectively, at frequencies f_2 and f_1. Therefore, the proposed diplexer is reduced to the circuits shown in figure 3.14 at frequencies f_1 and f_2 and also their odd multiplications. From figure 3.14, it is simply investigable that the input port is impedance matched as well as the isolation between two output ports is infinite at both frequencies f_1 and f_2. Also, the signal of frequencies f_1 and f_2 are separately delivered to Ports 2 and 3, respectively.

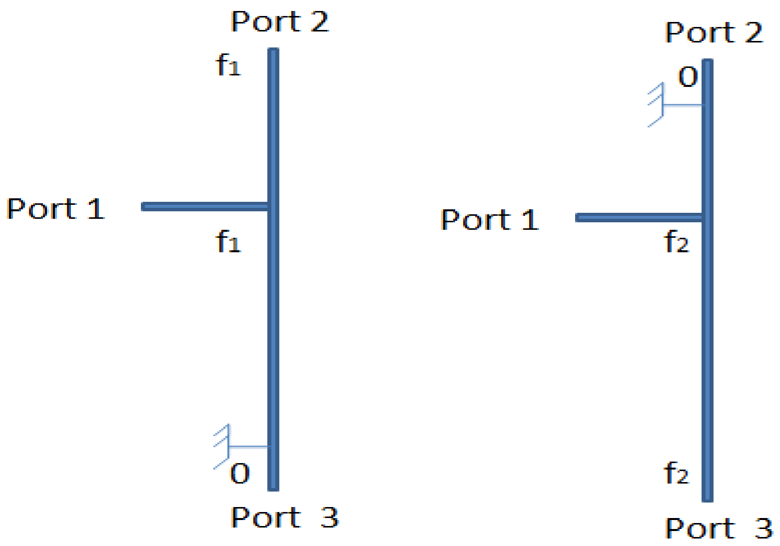

Figure 3.14 The equivalent circuits of diplexer at frequencies f1 (left) and f2 (right)

Therefore, it is understandable that the proposed simple structure shown in figure3.15 can be an ideal diplexer for frequencies f_1 and f_2 and their odd multiplications. Moreover, the proposed structure can be used as a de-diplexer (combiner) because its output ports are matched at their corresponding frequencies f_1 and f_2. One can determine the scattering parameters of the proposed diplexer at arbitrary frequency f.

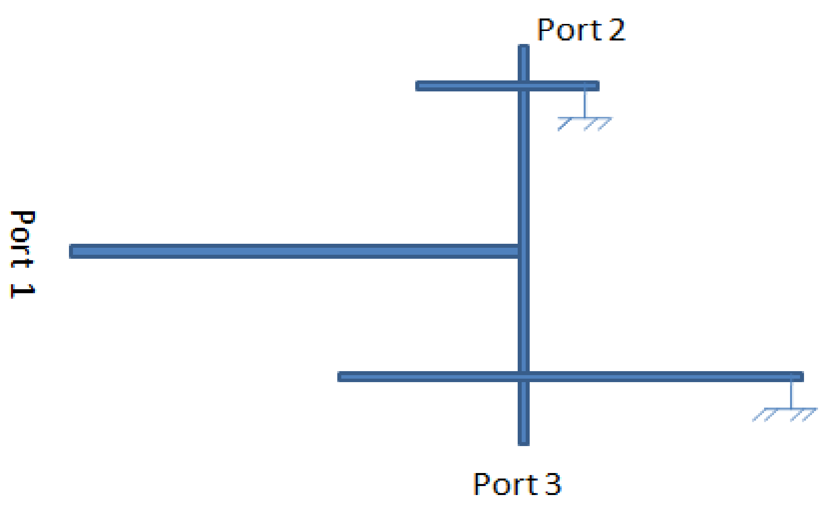

Figure 3.15Proposed Model of Microstrip Diplexer

3.6.2.2 Analysis of Diplexer design

The scattering parameters of the diplexer can be obtained using circuit relations as followings

$$S_{11} = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \qquad \text{.......... (49)}$$

$$S_{21} = \left(\cos(\theta_2) - j\frac{Z_0}{Z_{in1}}\sin(\theta_2) \right)\frac{2Z_{in}}{Z_{in} + Z_0} \qquad \text{........... (50)}$$

$$S_{31} = \left(\cos(\theta_1) - j\frac{Z_0}{Z_{in2}}\sin(\theta_1) \right)\frac{2Z_{in}}{Z_{in} + Z_0} \qquad \text{.......... (51)}$$

$$S_{23} = S_{32} = \left(\cos(\theta_1) - j\frac{Z_0}{Z_{in2}}\sin(\theta_1) \right)\times\left(\cos(\theta_2) - j\frac{Z_0}{Z_2^{'}}\sin(\theta_2) \right)\frac{2Z_2}{Z_2 + Z_0} \qquad \text{............ (52)}$$

Where, Z_{in} is the impedance of the Port 1 and also Z_{in1} andZ_{in2} are the impedances of the upper and lower half circuits, respectively. Furthermore, Z_2 and Z_2 are the impedance of the Port 2 with and without considering two upper stubs, respectively. Moreover, the $\theta 1$ and $\theta 2$ are the following electrical lengths at frequency f .

$$\theta_1 = \frac{\pi}{2}\frac{f}{f_1} \qquad \text{............ (53)}$$

$$\theta_2 = \frac{\pi}{2}\frac{f}{f_2} \qquad \text{............ (54)}$$

Next chapter 4 describes the detail process of ADS simulation based on the design and also result obtained.

Chapter 4

Simulation and Results Analysis

4.1 Introduction

This chapter illustrates the ADS simulation of proposed microstrip diplexer design. Section 4.2 describes the ADS simulation design and result obtained and their description is described in section 4.3 and its sub parts. Section 4.4 shows the layout structure of microstrip diplexer obtained after Fabrication.

4.2 ADS simulation model

Figure 4.1 shows the ADS design pallet for the proposed diplexer design. In this project all ports have 50Ω characteristic impedance, all other designing parameter value are given below

Microstrip Substrate properties:

h : Substrate thickness=0.8 mm

ε_r : Relative dielectric constant=4.3

t: Conductor thickness=20 μm

Transmission line properties:

TL1 length=4.35cm

TL2 length=17.31mm

TL3 length=14.5mm

TL4 length=14.5mm

TL5 length=17.31mm

TL6 length=2.78mm

TL7 length=26.26mm

All Transmission Line have same width =1.5358mm

Here combination transmission line 3, 4 and 6 works as a microstrip bandpass filter, having centre frequency at 2.4GHz while combination of line 2, 5 and 7 works as another bandpass filter cantered at 2.85GHz. Transmission line is common to both of lines. And input is applied through this line port1.

For the proposed design S parameter simulation is find and analysed at the next section.

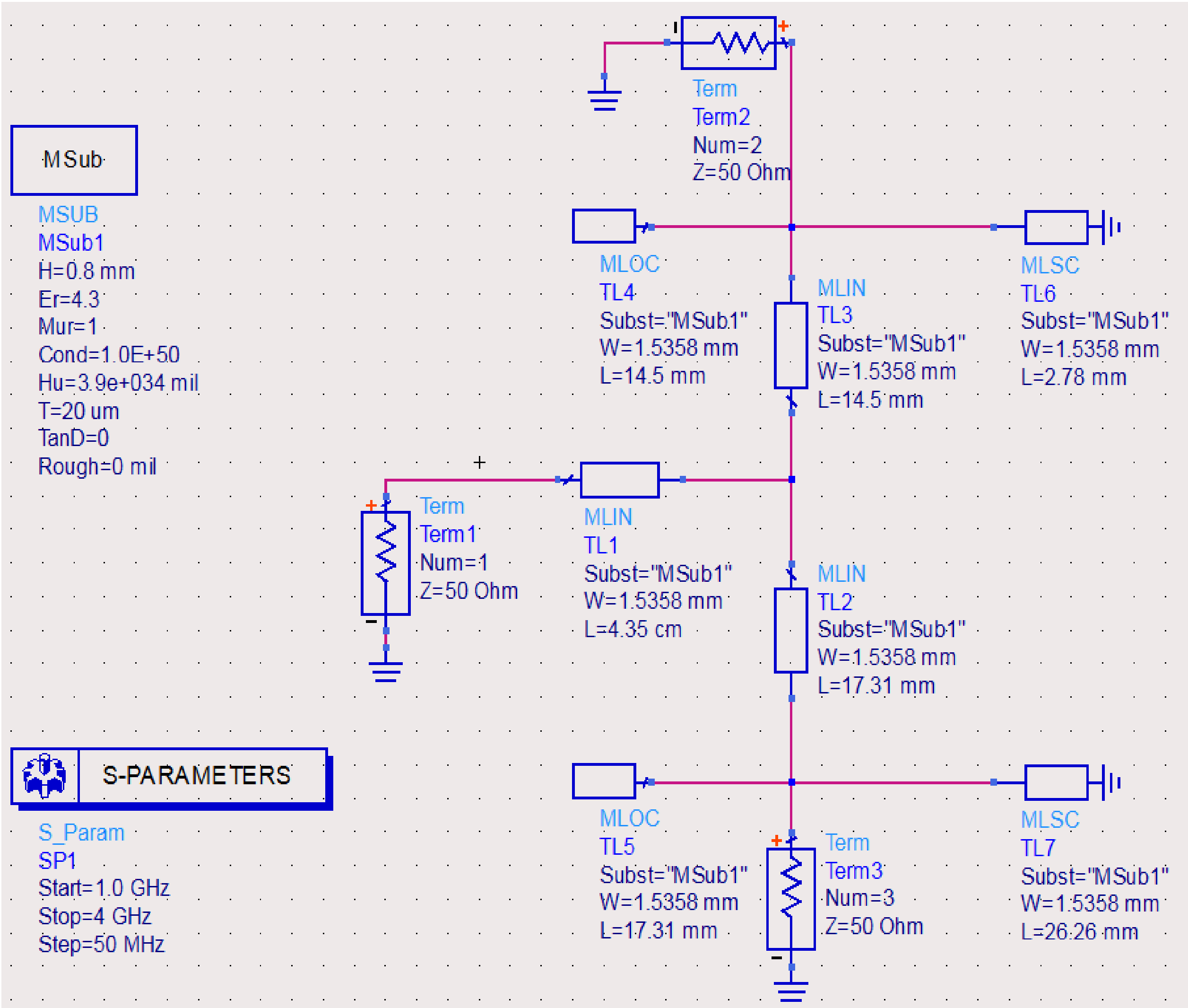

Figure 4.1 Equivalent circuits of microstrip diplexer

4.3 Result Analysis

Proposed model is simulated and all the scattering parameters are illustrated, in figure 4.2. Return loss (i.e. S11 in dB) which is -34.642 dB at frequencies 2.4GHz and -24.3dB at frequency 2.85 GHz. Figure 4.3 shows the transmission loss for port 2 and port 3. At f = 2.4

GHz S_{21}= -0.001 dB and S_{31} = -43.877 dB i.e. port 2 passes the power for frequency f=2.4 GHz while at the same time port 3 curbs the power

4.3.1 Observation of Return loss parameter S11:

Here from the figure 4.2 we can observe that at frequency 2.4GHz and near to frequency 2.85 GHz we have minimum return loss, which indicates that at these two frequencies signal will be transmitted in the system to the output ports 2 and 3.

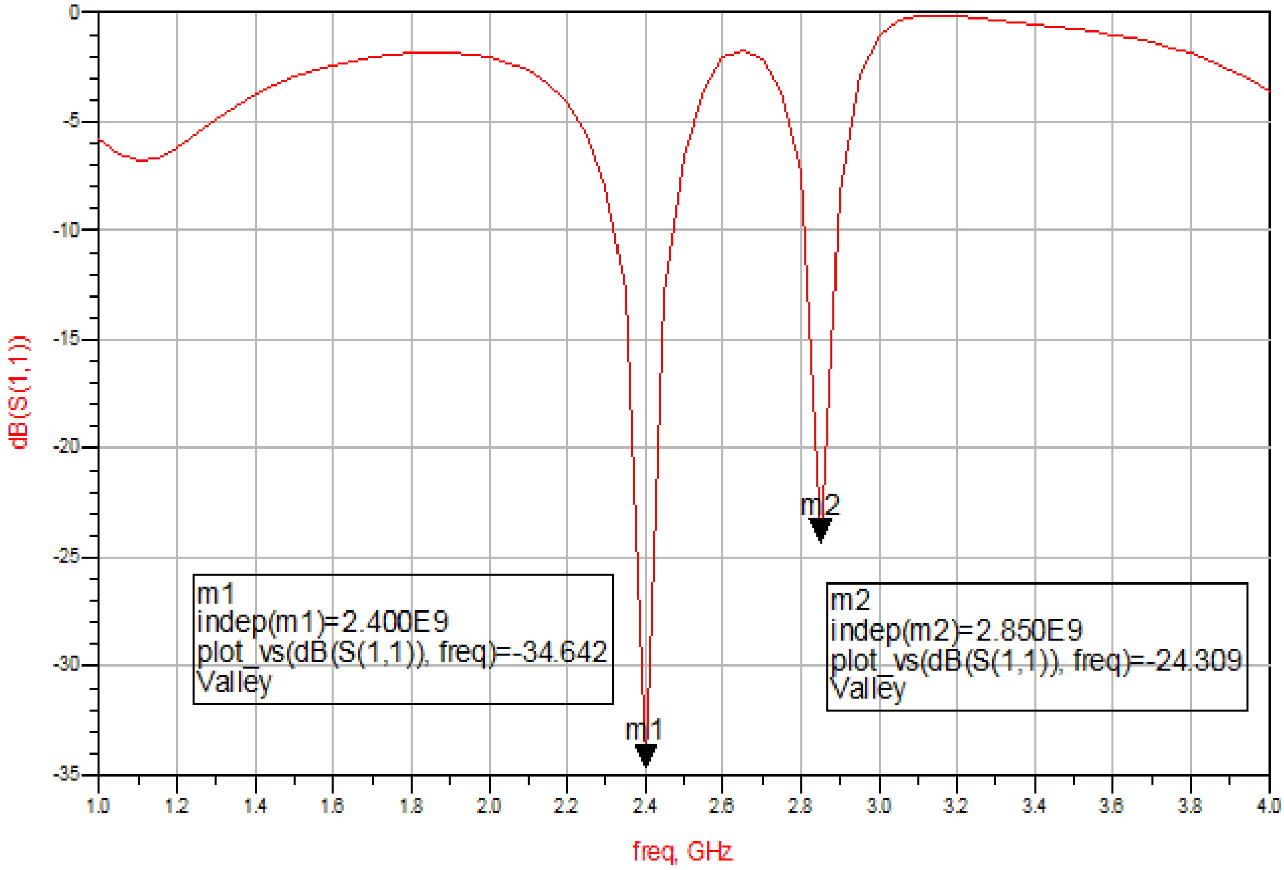

Figure 4.2 Return losses (S11) for Port 1

Variation in S_{11} parameter with frequency is listed below at table 4.1.

freq	S(1,1)
1.000 GHz	0.511 / -61.106
1.050 GHz	0.471 / -66.460
1.100 GHz	0.455 / -70.369
1.150 GHz	0.463 / -74.524
1.200 GHz	0.490 / -80.360
1.250 GHz	0.528 / -88.312
1.300 GHz	0.570 / -98.109
1.350 GHz	0.610 / -109.287
1.400 GHz	0.648 / -121.447
1.450 GHz	0.681 / -134.313
1.500 GHz	0.710 / -147.710
1.550 GHz	0.735 / -161.542
1.600 GHz	0.756 / -175.762
1.650 GHz	0.773 / 169.639
1.700 GHz	0.787 / 154.645
1.750 GHz	0.798 / 139.221
1.800 GHz	0.805 / 123.313
1.850 GHz	0.808 / 106.852
1.900 GHz	0.807 / 89.755
1.950 GHz	0.801 / 71.918
2.000 GHz	0.788 / 53.224
2.050 GHz	0.767 / 33.532
2.100 GHz	0.735 / 12.680
2.150 GHz	0.687 / -9.519
2.200 GHz	0.620 / -33.284
2.250 GHz	0.526 / -58.868
2.300 GHz	0.398 / -86.578
2.350 GHz	0.230 / -116.848
2.400 GHz	0.019 / -157.543
2.450 GHz	0.226 / -3.041
2.500 GHz	0.470 / -40.378
2.550 GHz	0.668 / -78.042
2.600 GHz	0.787 / -114.895
2.650 GHz	0.821 / -150.368
2.700 GHz	0.778 / 175.122
2.750 GHz	0.657 / 140.257
2.800 GHz	0.430 / 102.879
2.850 GHz	0.061 / 59.630
2.900 GHz	0.387 / -163.596
2.950 GHz	0.721 / 154.850
3.000 GHz	0.888 / 120.796

Table 4.1 observation of S_{11} parameter

From table it is clear that signal of frequency 2.4GHz and 2.8GHz has minimum loss.

4.3.2 Observation of S21parameter:

Transmission parameter S_{21} helps to find the transmitted power from port1 to port 2, here figure 4.3 shows the observed S_{21} parameter with respect to the frequency.

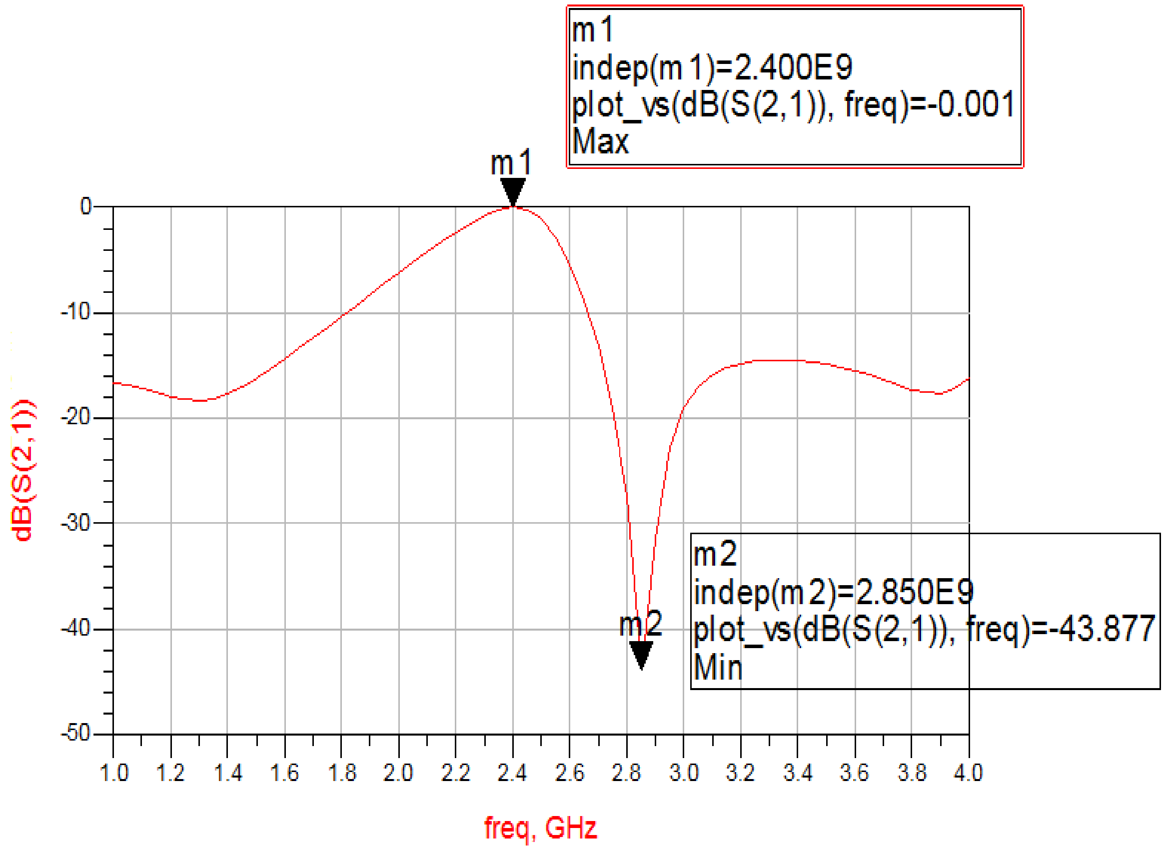

Figure 4.3 Properties of Transmission parameter (S21)

S_{21} shows its maximum value at 2.4GHz, which is -0.001dB, shows that it will allow transmission of signal at 2.4 GHz and provide high loses for the other frequency. A list of S_{21} values at other frequencies is given in table 4.2.

freq	S(2,1)
1.000 GHz	0.146 / -68.986
1.050 GHz	0.143 / -77.314
1.100 GHz	0.138 / -84.573
1.150 GHz	0.132 / -90.446
1.200 GHz	0.127 / -94.739
1.250 GHz	0.123 / -97.493
1.300 GHz	0.122 / -99.045
1.350 GHz	0.125 / -99.996
1.400 GHz	0.131 / -101.026
1.450 GHz	0.142 / -102.674
1.500 GHz	0.155 / -105.229
1.550 GHz	0.172 / -108.769
1.600 GHz	0.193 / -113.247
1.650 GHz	0.216 / -118.571
1.700 GHz	0.242 / -124.653
1.750 GHz	0.272 / -131.425
1.800 GHz	0.305 / -138.846
1.850 GHz	0.342 / -146.905
1.900 GHz	0.385 / -155.615
1.950 GHz	0.432 / -165.016
2.000 GHz	0.485 / -175.174
2.050 GHz	0.545 / 173.823
2.100 GHz	0.611 / 161.855
2.150 GHz	0.683 / 148.777
2.200 GHz	0.760 / 134.407
2.250 GHz	0.839 / 118.530
2.300 GHz	0.913 / 100.890
2.350 GHz	0.972 / 81.211
2.400 GHz	1.000 / 59.279
2.450 GHz	0.974 / 35.156
2.500 GHz	0.878 / 9.499
2.550 GHz	0.721 / -16.348
2.600 GHz	0.538 / -40.895
2.650 GHz	0.364 / -63.179
2.700 GHz	0.220 / -82.730
2.750 GHz	0.112 / -98.465
2.800 GHz	0.042 / -105.033
2.850 GHz	0.006 / -91.541
2.900 GHz	0.027 / 91.777
2.950 GHz	0.073 / 75.681
3.000 GHz	0.113 / 57.850

Table 4.2 Observation of S21 parameter

Table 4.2 shows that S_{21} parameter is -91.777 at 2.85GHz, which means it will provide good isolation between port 2 and 3 for signal at 2.4GHz. While it also provide loss of signal for other frequency components.

4.3.3 Observation of S_{31}parameter:

Transmission parameter S_{31} helps to find the transmitted power from port1 to port 3, here diplexer is designed to transfer maximum power to port3 at frequency 2.85GHz. Figure 4.3 shows the observed S_{31} parameter with respect to the frequency.

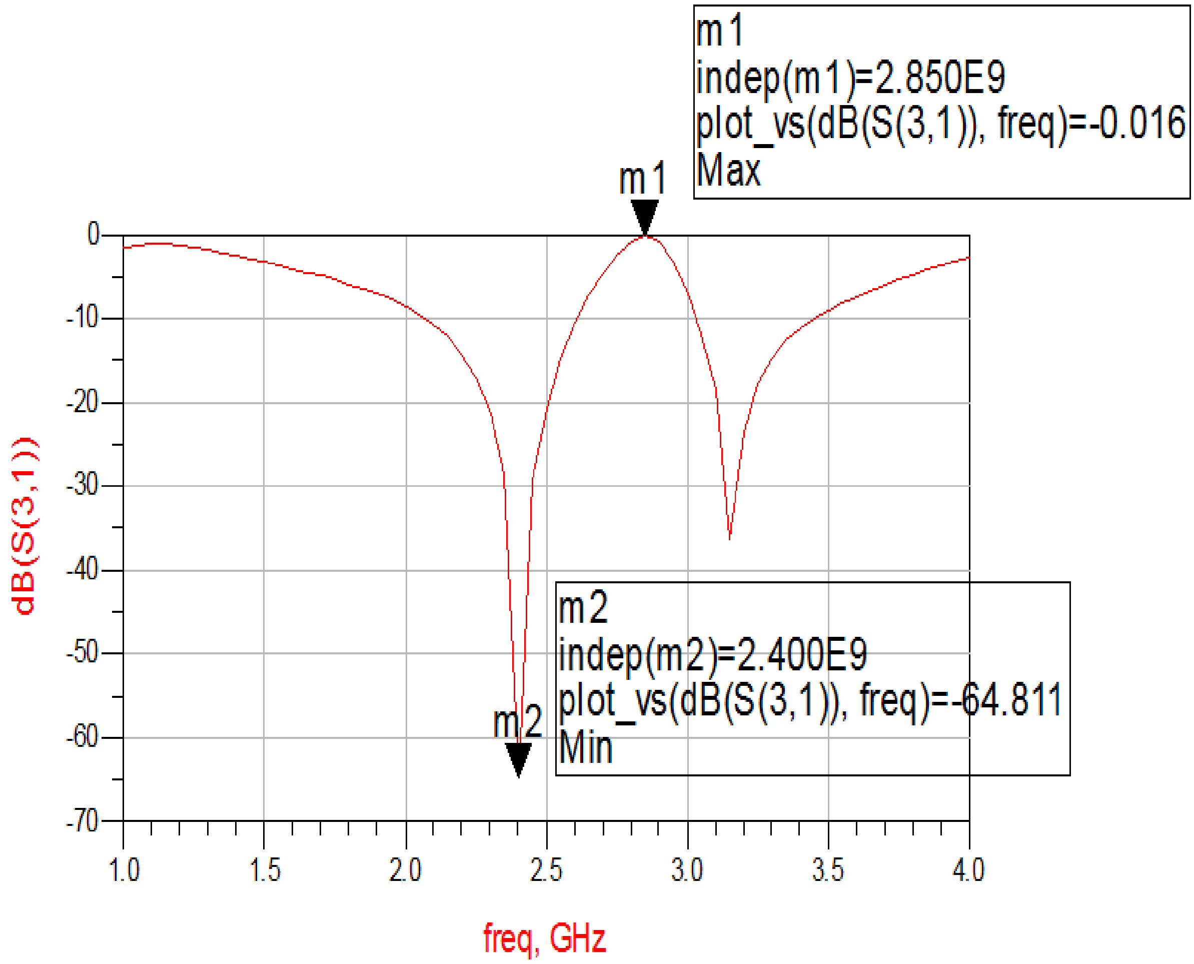

Figure 4.4 transmission parameter S(3,1)

S_{31} shows its maximum value at 2.85GHz, which is -0.016dB, shows that it will allow transmission of signal at 2.85 GHz and provide high loses for the other frequency. A list of S_{31} values at other frequencies is given in table 4.3

freq	S(3,1)
1.000 GHz	0.847 / -103.434
1.050 GHz	0.870 / -116.857
1.100 GHz	0.880 / -130.180
1.150 GHz	0.876 / -143.230
1.200 GHz	0.862 / -155.880
1.250 GHz	0.840 / -168.065
1.300 GHz	0.813 / -179.770
1.350 GHz	0.782 / 168.981
1.400 GHz	0.751 / 158.140
1.450 GHz	0.718 / 147.644
1.500 GHz	0.687 / 137.427
1.550 GHz	0.656 / 127.421
1.600 GHz	0.626 / 117.559
1.650 GHz	0.596 / 107.777
1.700 GHz	0.567 / 98.012
1.750 GHz	0.538 / 88.201
1.800 GHz	0.509 / 78.283
1.850 GHz	0.479 / 68.193
1.900 GHz	0.448 / 57.866
1.950 GHz	0.414 / 47.236
2.000 GHz	0.378 / 36.237
2.050 GHz	0.339 / 24.807
2.100 GHz	0.295 / 12.907
2.150 GHz	0.247 / 0.536
2.200 GHz	0.195 / -12.203
2.250 GHz	0.140 / -24.980
2.300 GHz	0.086 / -36.918
2.350 GHz	0.038 / -45.822
2.400 GHz	5.747E-4 / 132.813
2.450 GHz	0.034 / 140.089
2.500 GHz	0.090 / 141.232
2.550 GHz	0.182 / 129.834
2.600 GHz	0.304 / 111.848
2.650 GHz	0.441 / 91.473
2.700 GHz	0.588 / 69.965
2.750 GHz	0.745 / 46.789
2.800 GHz	0.902 / 20.215
2.850 GHz	0.998 / -11.685
2.900 GHz	0.922 / -47.319
2.950 GHz	0.689 / -79.862
3.000 GHz	0.446 / -105.445

Table 4.3 Observation of S31 parameter

4.3.4 Isolation parameter (S_{32}):

S_{32} parameter investigates the isolation between the ports 3 and port2. It is recommended to calculate S_{32} as find the isolation between two ports at operating frequencies. For proposed diplexer design S_{32} is given below

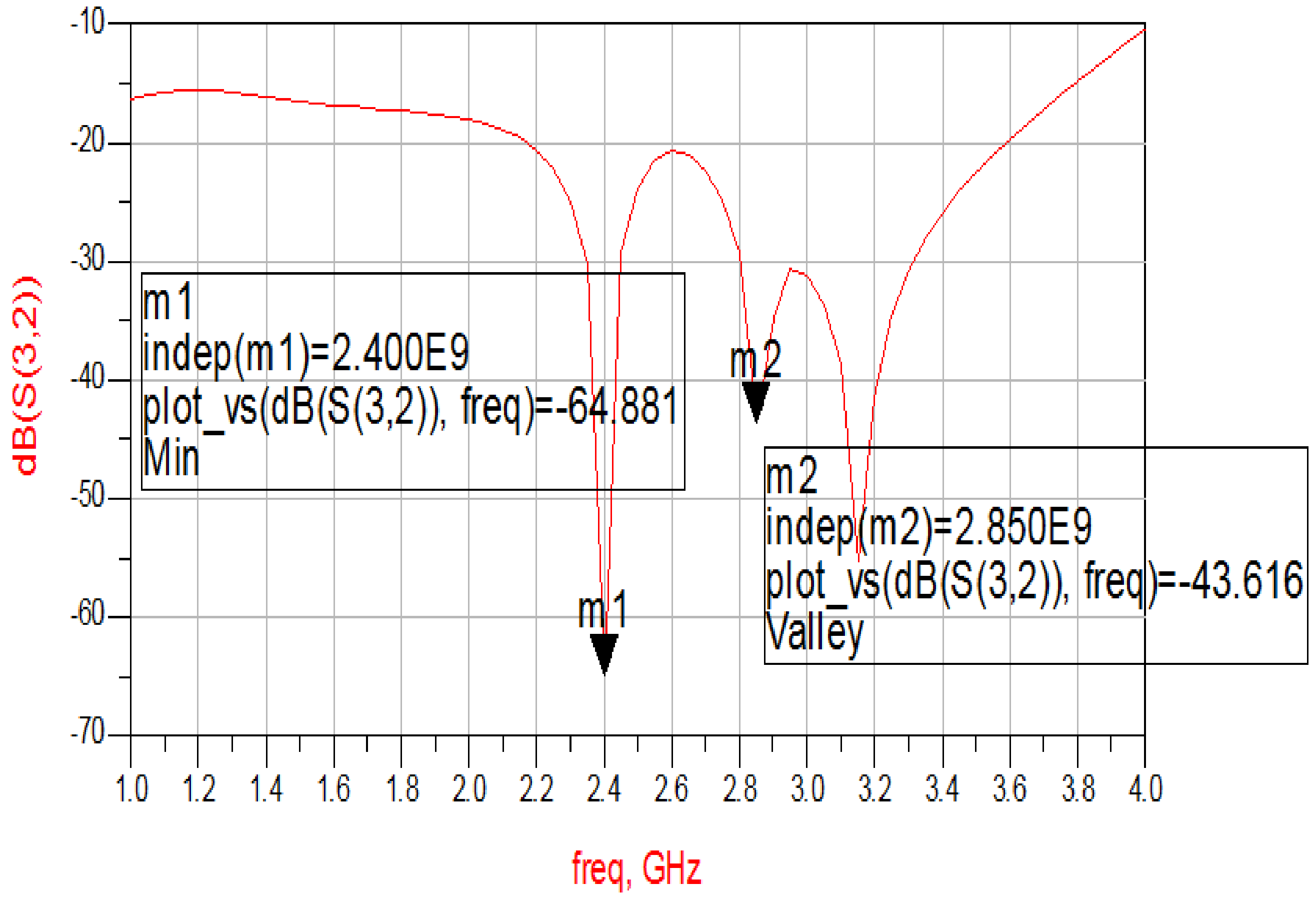

Figure 4.5 observation of S_{32} parameter

S_{32} obtained less value at frequency 2.4GHz and at 2.85 GHz, which shows that there will be no transmission of signal between port 2 and port3 for these frequencies, providing a good isolation between ports for the frequency 2.4 GHz and 2.85GHz. Values of S_{32} are listed in table 4.4

freq	S(3,2)
1.000 GHz	0.154 / -14.772
1.050 GHz	0.160 / -23.798
1.100 GHz	0.164 / -32.733
1.150 GHz	0.166 / -41.404
1.200 GHz	0.166 / -49.688
1.250 GHz	0.165 / -57.520
1.300 GHz	0.163 / -64.886
1.350 GHz	0.160 / -71.814
1.400 GHz	0.157 / -78.354
1.450 GHz	0.153 / -84.571
1.500 GHz	0.150 / -90.535
1.550 GHz	0.148 / -96.319
1.600 GHz	0.145 / -101.994
1.650 GHz	0.142 / -107.630
1.700 GHz	0.140 / -113.299
1.750 GHz	0.138 / -119.072
1.800 GHz	0.136 / -125.026
1.850 GHz	0.134 / -131.242
1.900 GHz	0.131 / -137.808
1.950 GHz	0.129 / -144.825
2.000 GHz	0.125 / -152.407
2.050 GHz	0.120 / -160.684
2.100 GHz	0.114 / -169.809
2.150 GHz	0.105 / -179.960
2.200 GHz	0.093 / 168.656
2.250 GHz	0.077 / 155.794
2.300 GHz	0.057 / 141.169
2.350 GHz	0.030 / 124.471
2.400 GHz	5.701E-4 / -74.555
2.450 GHz	0.034 / -95.904
2.500 GHz	0.064 / -119.000
2.550 GHz	0.084 / -142.629
2.600 GHz	0.092 / -165.524
2.650 GHz	0.088 / 172.849
2.700 GHz	0.076 / 152.024
2.750 GHz	0.058 / 130.640
2.800 GHz	0.035 / 106.547
2.850 GHz	0.007 / 77.578
2.900 GHz	0.018 / -134.816
2.950 GHz	0.029 / -163.899
3.000 GHz	0.028 / 174.138

Table 4.4 Observation of S32 parameter

From table 4.4 S32 value at frequency 2.4GHz and 2.85GHz are 5.701E-4 and 0.007 which indicate no transmission of signal at these two frequencies between ports 2and3. All other frequency component will be controlled through bandpass filters.

4.3.5 Layout of Diplexer design:

Figure 4.6 shows the observed Layout design for the proposed diplexer design. This layout is obtained using ADS software.

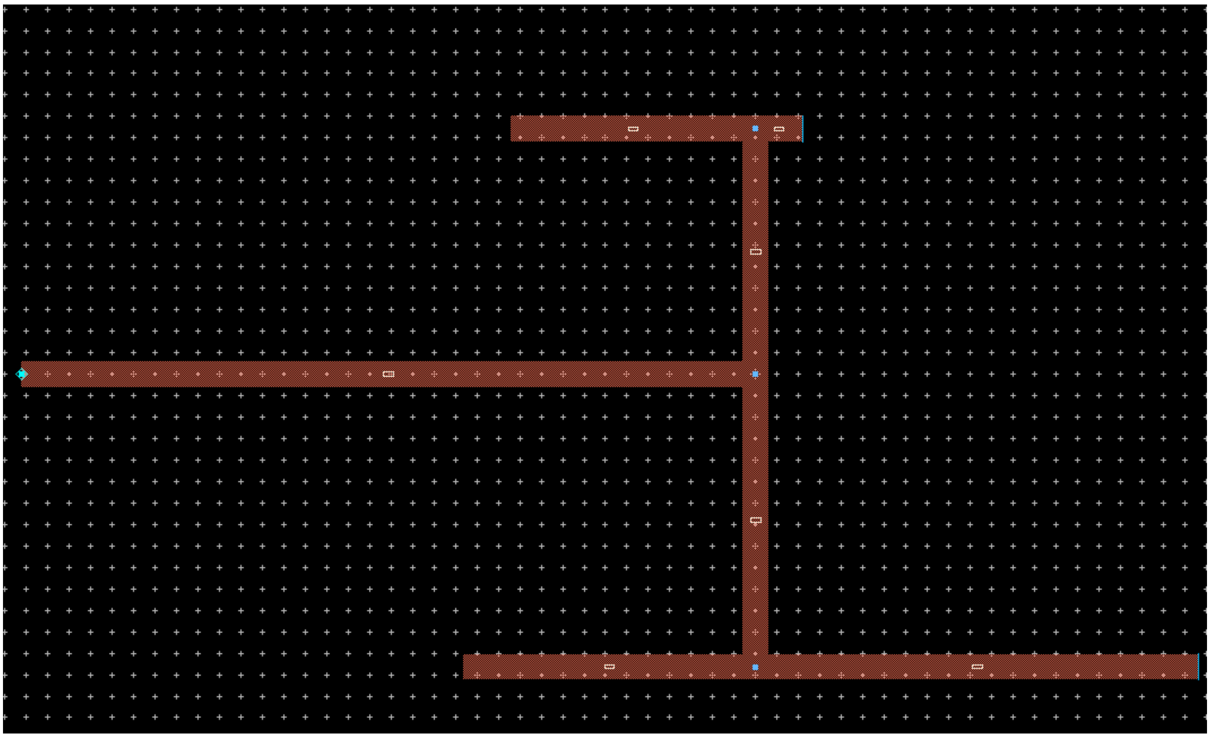

Figure 4.6 diplexer layout

4.4 Fabricated Diplexer design

Figure 4.7 Fabricated Diplexer design

Figure 4.7 shows fabricated diplexer design based on proposed structure. Results can be verified by operating the device with appropriate system.

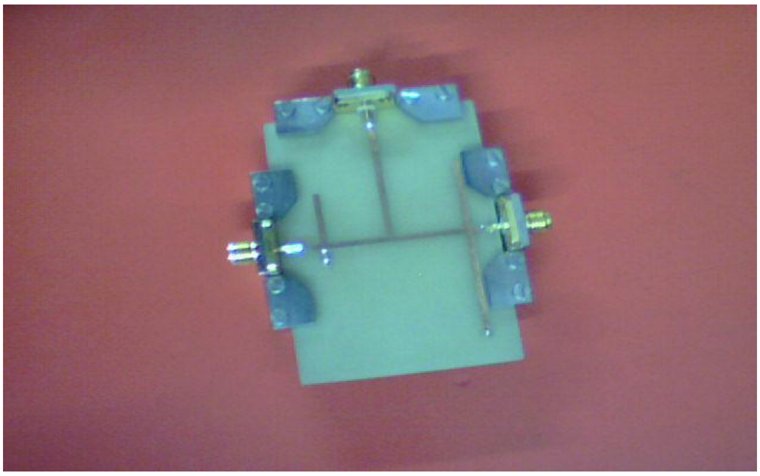

Figure 4.8 Fabricated Diplexer design with three ports

Testing has been done for the given Diplexer design.

Next chapter contains conclusion obtained for this project and future scope for the present work.

Chapter 5

Conclusion and scope for Future work

5.1 Conclusions

The objective of this work has been to

- Microstrip diplexer has been design using Simple and accurate designing formula.
- Return loss (i.e. S11 in dB) which is -34.642 dB at frequencies 2.4GHz and -24.3dB at frequency 2.85 GHz, indicates minimum return loss (maximum transmission) at these two frequencies at port1.
- Transmission parameter S_{21} is obtained is -0.001dB at 2.4GHz while it is -43.877dB at 2.85 GHz. which indicates that transmission of 2.4GHz signal from port 1 to port 2 is possible with minimum attenuation and signal corresponds to frequency 2.85GHz will be highly attenuated.
- S_{31} is calculated as -0.016dB at 2.85GHz and -64.811dB at 2.4GHz which shows that signal with frequency 2.85GHz will transfer from port1 to port2 with minimum attenuation. And signal with 2.4GHz frequency will be highly attenuated.
- Observed S_{32} parameter is -64.881dB and -43.616dB at 2.4GHz and 2.85GHz respectively, shows a good isolation between port2 and port3.

The proposed design is simulated in ADS software tool, and then fabricated. And Fabricated results are verified with the theoretical obtained values.

5.2 Future Scope of present work

By using different configuration of serial, short, and open stub Transmission line we can implement many other circuits like triplexes. The future extension of this project shall be:

- Different types of filters such as Low pass, high pass, bandpass and bandstop filter for high frequency application [22].
- Implementation of higher order BPF using microstrip transmission line. And design of triplexes and other frequency selective device for microwave application.

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