# DESIGN OF PARALLEL COUPLED MICROSTRIP BAND PASS FILTER FOR WIRELESS APPLICATION

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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**2K12 / MOC / 05**


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# CERTIFICATE

This is to certify that the thesis report entitled, Design of Parallel Coupled Microstrip Band Pass Filter for Wireless Application" being submitted by **Arun Prakash Anthwal** to the *Department of Electronics and Communication Engineering and Applied Physics*, *Delhi Technological University*, *Delhi* in partial fulfilment of the requirement for award of Master of Technology degree in ***Microwave and Optical Communication*** is a record of bona fide work carried out by him under the supervision and guidance of **Professor Prem R. Chadha**. The matter embodied in this report has not been submitted for the award of any other degree.

**Prof. Prem R. Chadha**

Supervisor

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# DECLARATION

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

Radio frequency (RF) filters operating in the microwave frequency range are needed for applications including wireless and satellite communications as well as military applications. Most communication system contains an RF front end which performs signal processing with RF filters. Planar or printed circuit board (PCB) based filters are popular and relatively practical to design. Band pass filters play a significant role in wireless communication systems. Transmitted and received signals have to be filtered at a certain center frequency with a specific bandwidth. Parallel coupled line microstrip filter are very common for implementation of bandpass and band-stop filters with required bandwidth up to a 20% of fractional bandwidth. Because of weak coupling these type of filter has narrow fractional bandwidth (FBW) but has certain advantages such as low cost fabrication, easy integration and simple design procedure. This thesis describes the design of low cost L-band Parallel coupled microstrip Band pass filter (BPF) for GPS system by using micro strip layout at center frequency 1.575 GHz with a fractional bandwidth of 20%. The simulation of this filter has been done using ADS (Advanced design system) design software by Agilent Technology. And also for calculation of the physical parameter of the coupled line CAD tool LineCalc built in ADS is used.

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# LIST OF SYMBOLS

| | |
|---|---|
| $V_{out}$ | Output Voltage |
| $V_{in}$ | Input Voltage |
| $\Gamma_S$ | Source Reflection Coefficient |
| $\Gamma_L$ | Load Reflection Coefficient |
| $\Delta f$ | Bandwidth |
| $f_o$ | Centre Frequency |
| $L_A$ | Insertion loss |
| $L_R$ | Return loss |
| K | Impedance inverter |
| J | Admittance Inverter |
| $Z_0$ | Characteristic impedance |
| $\beta$ | Propagation constant |
| TanD | Loss Tangent |
| Q | Quality Factor |
| L | Inductor |
| C | Capacitor |
| $t_g$ | Time delay |
| $\Omega$ | Ohm |

# LIST OF ABBREVIATIONS

| | |
|---|---|
| ADS | Advance Design System |
| GPS | Global Positioning System |
| MMIC | Monolithic Microwave Integrated Circuit |
| LPF | Lowpass filter |
| BPF | Band-Pass Filter |
| MEMS | Microelectromechanic System |
| VSWR | Voltage Standing Wave Ratio |
| TEM | Transverse Electromagnetic Field |
| FBW | Fractional bandwidth |
| CAD | Computer Aided Design |

# Chapter 1

# Introduction

The term microwave refers to the frequency range from 300 MHz to 300 GHz, with corresponding electrical wavelengths from 1 m to 1 mm. Microwave components are often distributed elements, where the phase of a voltage or current changes significantly over the physical extent of the device. Even though microwave engineering had its beginning in the early 20th century, significant developments in high frequency solid state devices, microwave integrated circuits, and the ever-increasing applications of modern Microsystems have kept the field active and vibrant. The majority of today's microwave technology applications are in communication systems, radar systems, medical systems and environmental remote sensing. The most ubiquitous use of microwave technology is in cell phone systems. Satellite systems, such as Global Positioning Satellite (GPS) system and Direct Broadcast Satellite (DBS) system, have been extremely successful in providing cellular, video, and data connections worldwide The advantages offered by microwave systems, including wide bandwidths and line-of-sight propagation, have proved to be critical for both terrestrial and satellite communications systems and have thus provided an impetus for the continued development of low-cost miniaturized microwave components. In the whole microwave spectrum, the super high frequency (SUF) band that is from 3 GHz to 30 GHz is widely utilized in military satellite system [1-2].

In most microwave communication systems, filters play extremely important role. A microwave filters separate or combines different frequencies and is used to select or confine the microwave signals within assigned spectral limits. Emerging applications such as wireless communications continue to challenge microwave filter designers with ever more stringent requirements: higher performance, smaller size, etc. In the following sub-sections, basic characteristics and existing implement method of microwave filters will be introduced, and motivation, objective of the thesis will be presented.

## 1.1 Microwave Filter

A microwave filter is a two-port network used to control the frequency response at a certain point in a microwave system by providing transmission at frequencies within the passband and attenuation in the stopband. Typical frequency responses include low pass, high pass, band pass, and band stop characteristics. Microwave filter applications can be found in virtually any type of microwave communication, radar, or test and measurement system. The fundamental use of filters in electronics engineering is to shape signal spectrum, which is especially crucial in reducing input signal noise in receivers and spurious emissions in transmitters. In electronics engineering, filtering can be intentional, as in the input stage of a receiver, or unintentional, as in the transmission path of a microwave signal. It is important to understand that almost every physical system has some sort of filtering action built in whenever a signal, an input, an output, and a transmission path can be defined in the system.

Microwave passive filters have traditionally been built using waveguides and coaxial lines. Recent microwave printed filters have unique advantages over waveguide and coaxial filters in terms of low cost, repeatability, high accuracy, and compact size. Another advantage of printed filters is their easy integration with active circuits (i.e., filters can be fabricated on the same substrate with transistor amplifiers, oscillators, and other active circuits). Today, microstrip lines are perhaps the most common materials for microwave printed circuits, because of their simplicity, ease of manufacturing and high suitability for incorporation with active devices. Recent advances in materials and fabrication technologies, including metamaterials, monolithic microwave integrated circuit (MMIC), microelectromechanic system (MEMS) have stimulated research on novel microstrip filters.

In the past few decades, CAD of microwave components has greatly progressed due to an increase in the overall design complexity of modern satellite communication systems. Currently, a plethora of CAD tools, e.g. Ansoft's HFSS, Zeland's IE3D, Agilent's ADS, etc., are available for microwave designers. These tools not only help in the design of a microwave circuit, but also facilitate optimization of the circuit for improved performance in terms of power, size, timing, etc. Many novel microwave filters with advanced filtering characteristics have been demonstrated in different CAD tools. [3-7]

## 1.2 Motivation and Objectives

The recent and continuing evolution in telecommunications has implied stringent constraints on microwave systems and, especially on filters. Many applications and notably those involved in mobile and satellite communications, require miniaturizing the system dimensions without sacrificing electrical performances. The work presented in this thesis focuses on the desire to effectively reduce the volume and complexity involved in the design of printed microstrip bandpass filters, while keeping the same frequency response. In this work my aim is to design a bandpass filter for GPS application that will fulfill the entire requirement for the GPS receiver.

## 1.3 Thesis Structure

These outlines briefly describe the main part of this thesis:

- Theory: the basic knowledge of network theory, microstrip line, and microstrip filter will be demonstrated. Overview the method of design of parallel-coupled line filter.
- Implementation and analysis: simulate the filters by using ADS software.
- Conclusion: thesis conclusion

# Chapter 2

# Network Theory

Most of the RF and Microwave systems and devices can be modeled as a two port network. The two port representation basically helps in isolating either a complete circuit or a part of it and finding its characteristic parameters. Once this is done, the isolated part of the circuit, with a set of distinctive properties, enables us to abstract away its specific physical buildup, thus simplifying analysis. Any circuit can be transformed into a two-port network provided that it does not contain an independent source.

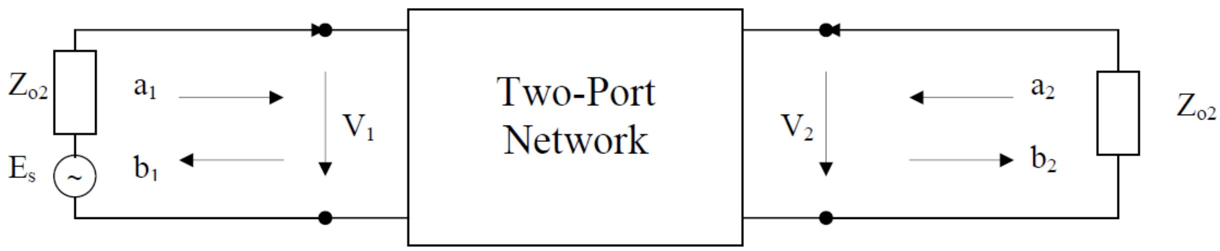


**Figure 2.1:** Two port network with its wave variables [1]

Where $V_1$, $V_2$ and $I_1$, $I_2$ are the voltages and currents at respective ports and $Z_{01}$ $Z_{02}$ are the terminal impedances. At RF and Microwave frequencies it is difficult to measure the voltages, thus new wave variables $a_1$ ,$b_1$ and $a_2$ ,$b_2$ are introduced with *a* signifying the incident wave and *b* implying the reflected wave.

The wave variables in terms of voltage and current are defined as follows:

$$V_n = \sqrt{Z_{0n}}(a_n + b_n) \tag{2.1}$$

$$I_n = \frac{1}{\sqrt{Z_{0n}}}(a_n - b_n) \text{ for n=1 and n=2} \tag{2.2}$$

or

$$a_n = \frac{1}{2}\left(\frac{V_n}{\sqrt{Z_{on}}} + \sqrt{Z_{on}} I_n\right) \tag{2.3}$$

$$b_n = \frac{1}{2}\left(\frac{V_n}{\sqrt{Z_{on}}} - \sqrt{Z_{on}} I_n\right) \text{ For n=1 and n=2} \tag{2.4}$$

This gives power at each port $P_n$

$$P_n = \frac{1}{2}\mathrm{Re}\left(V_n \cdot I_n^*\right) = \frac{1}{2}\left(a_n a_n^* - b_n b_n^*\right) \tag{2.5}$$

Now we will see few types of parameters used to define a two port network

## 2.1 Scattering Parameters

Since the main problem with RF was that one has to rely on open and short circuiting the other port if one has to define Y and Z parameters for RF circuits therefore scattering parameter has the advantage that they can be measured by matching source and load impedance to the reference impedance. To represent a two-port network at microwave frequencies, scattering parameters(S - parameters) can be used. S-parameters themselves ($S_{11}$, $S_{12}$, $S_{21}$, and $S_{22}$) represent reflection and transmission coefficients of the two-port under certain “matched” conditions. $S_{11}$ is the reflection coefficient and $S_{21}$ is the transmission coefficient at port 1 when port 2 is terminated in a load whose impedance is equal to that of the transmission line characteristic impedance. Likewise $S_{22}$ is the reflection coefficient and $S_{12}$ is the transmission coefficient at port 2 when port 1 is terminated in a matched load. The S parameter representation for two port network is as shown in Figure (2.1).

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

And the $a_1$, $b_1$, the terminal current and terminal voltage relation is expressed in Equations (2.6) and (2.7),

$$a_i = \frac{v_i + z_0 i_i}{2\sqrt{z_0}} \tag{2.6.}$$

$$b_i = \frac{v_i - z_0 i_i}{2\sqrt{z_0}} \tag{2.7}$$

Here $Z_0$ is the reference impedance and it is taken as 50 Ω for all practical application. The overall scattering parameter in the matrix form can expressed as shown in Equation (2.8),

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} s_{11} & s_{12} \\ s_{21} & s_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \tag{2.8}$$

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

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

$$s_{11} = \frac{b_1}{a_1} \tag{2.9}$$

$$s_{12} = \frac{b_1}{a_2} \tag{2.10}$$

$$s_{21} = \frac{b_2}{a_1} \tag{2.11}$$

$$s_{22} = \frac{b_2}{a_2} \tag{2.12}$$

Where

$S_{11}$ is input reflection coefficient,

$S_{12}$ is reverse transmission coefficient,

$S_{21}$ is forward transmission coefficient,

$S_{22}$ output reflection coefficient.

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

## 2.2 Open-Circuit Impedance Parameter

Impedance parameters are very useful in designing impedance matching and power distribution systems. Two port networks can either be voltage or current driven. The open-circuit impedance or *Z* parameters of a two-port network are defined as

$$Z_{11} = \frac{\mathrm{V_1}}{\mathrm{I_1}} \text{ (When } I_2 = 0) Z_{12} = \frac{\mathrm{V_1}}{\mathrm{I_2}} \text{ (When } I_1 = 0)$$

$$Z_{21} = \frac{\mathrm{V_2}}{\mathrm{I_1}} \text{ (When } I_2 = 0) Z_{22} = \frac{\mathrm{V_2}}{\mathrm{I_2}} \text{ (When } I_1 = 0)$$

Where $I_n$ = 0 implies a perfect open-circuit at port *n*. These definitions can be written as

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} \mathrm{Z}_{11} & \mathrm{Z}_{12} \\ \mathrm{Z}_{21} & \mathrm{Z}_{22} \end{bmatrix} . \begin{bmatrix} \mathrm{I}_1 \\ \mathrm{I}_2 \end{bmatrix} \tag{2.13}$$

The matrix, which contains the Z parameters, is known as the open-circuit impedance or *Z* matrix and is denoted by $[Z]$.

For reciprocal networks $\mathrm{Z}_{12} = \mathrm{Z}_{21}$. If networks are symmetrical, $\mathrm{Z}_{12} = \mathrm{Z}_{21}$ and $\mathrm{Z}_{11} = \mathrm{Z}_{22}$. For a lossless network, the Z parameters are all purely imaginary.

## 2.3 Short-Circuit Admittance Parameter

Admittance parameters are very useful for describing the network when impedance parameters may not exist. This is solved by finding the second set of parameters by expressing the terminal current in terms of the voltage.

The short-circuit admittance or *Y* parameters of a two-port network are defined as

$$\mathrm{Y}_{11} = \frac{\mathrm{I}_1}{\mathrm{V}_1} \text{ (When } V_2 = 0) \mathrm{Y}_{12} = \frac{\mathrm{I}_1}{\mathrm{V}_2} \text{(When } V_1 = 0)$$

$$\mathrm{Y}_{21} = \frac{\mathrm{I}_2}{\mathrm{V}_1} \text{ (When } V_2 = 0) \mathrm{Y}_{22} = \frac{\mathrm{I}_2}{\mathrm{V}_2} \text{ (When } V_1 = 0)$$

In which $v_n$= 0 implies a perfect short-circuit at port *n*. The definitions of the *Y* parameters may also be written as

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y_{11} & Y_{12} \\ Y_{21} & Y_{22} \end{bmatrix} . \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \tag{2.14}$$

Where the matrix containing the *Y* parameters is called the short-circuit admittance or simply *Y* matrix, and may be denoted by [*Y*]. For reciprocal networks $Y_{12} = Y_{21}$. In addition to this, if networks are symmetrical $Y_{11} = Y_{22}$ for a lossless network, the *Y* parameters are all purely imaginary.

## 2.4 ABCD Parameter

In *ABCD* parameter the input port voltage and current are considered variable and equation is formed in terms of the output voltage and current.

The *ABCD* parameters of a two-port network are given by:

$$A = \frac{V_1}{V_2} \text{ (When } I_2 = 0), \quad B = \frac{V_1}{-I_2} \text{ (When } V_2 = 0)$$

$$C = \frac{I_1}{V_2} \text{ (When } I_2 = 0), \quad D = \frac{I_1}{-I_2} \text{ (When } V_2 = 0)$$

These parameters are actually defined in a set of linear equations in matrix notation

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} . \begin{bmatrix} V_2 \\ -I_2 \end{bmatrix} \tag{2.15}$$

Where the matrix comprised of the *ABCD* parameters is called the *ABCD* matrix. Sometimes, it may also be referred to as the transfer or chain matrix. The *ABCD* parameters have the following properties:

AD – BC= 1 For a reciprocal network

A = D For a symmetrical network

*ABCD* parameters are useful in analysis when the network can be broken into cascaded sub networks

## 2.5 Important Definition

### 2.5.1 Insertion Loss

The loss resulting from the insertion of a network in a transmission line, expressed as the reciprocal of the ratio of the signal power delivered to that part of the line following the network to the signal power delivered to that same part before insertion. It is usually expressed in dB.

$$L_A = -20\log\left|S_{mn}\right| \text{ dB} \qquad m,n=1,2 \qquad (m \neq n) \qquad (2.16)$$

Where $L_A$ denotes the insertion loss between the ports n and m.

### 2.5.2 Return Loss

The Return Loss of a line is the ratio of the power reflected back from the line to the power transmitted into the line. It is usually expressed in dB

$$L_R = 20\log\left|S_{nn}\right| dB \qquad \text{n = 1, 2.} \qquad (2.17)$$

### 2.5.3 Voltage Standing Wave Ratio

A standing wave may be formed when a wave is transmitted into one end of a transmission line and is reflected from the other end by an impedance mismatch. VSWR is the ratio of the maximum to minimum voltage in a standing wave pattern.

$$VSWR = \frac{1+\left|S_{nn}\right|}{1+\left|S_{nn}\right|} \qquad \text{n= 1, 2.} \qquad (2.18)$$

### 2.5.4 Phase Delay

Whenever we insert a sinusoid into a filter, a sinusoid must come out. The only thing that can change between input and output are the amplitude and the phase. Comparing a zero crossing of the input to a zero crossing of the output measures the so-called phase delay. To quantify this we define an input, sin (ωt) and an output sin(ωt-Ø). Then the phase delay τ is found by solving

$$\sin(\omega t - \phi) = \sin\omega(t - \tau_p) \qquad (2.19)$$

$$\omega t - \phi = \omega t - \omega\tau_p \qquad (2.20)$$

$$\tau_p = \frac{\phi}{\omega} \text{ Seconds} \qquad (2.21)$$

Where φ is in radians and ω is in radians per second. The phase delay is actually the time delay for a steady sinusoidal signal and is not necessarily the true signal delay because a steady sinusoidal signal doesn't carry information

### 2.5.5 Group Delay

Often the group delay is nothing more than the phase delay. This happens when the phase delay is independent of frequency. But when the phase delay depends on frequency, then a completely new velocity, the "group velocity" appears. Curiously, the group velocity is not an average of phase velocities.

The simplest analysis of group delay begins by defining a filter input $x_t$ as the sum of two frequencies:

$$x_t = \cos \omega_1 t + \cos \omega_2 t \tag{2.22}$$

By using a trigonometric identity,

$$x_t = 2\cos\left(\frac{\omega_1 - \omega_2}{2} t\right)\cos\left(\frac{\omega_1 + \omega_2}{2} t\right) \tag{2.23}$$

We see that the sum of two cosines looks like a cosine of the average frequency multiplied by a cosine of half the difference frequency.

Each of the two frequencies could be delayed a different amount by a filter, so take the output of the filter $y_t$ to be

$$y_t = \cos(\omega_1 t - \phi_1) + \cos(\omega_2 t - \phi_2) \tag{2.24}$$

In doing this, we have assumed that neither frequency was attenuated. (The group velocity concept loses its simplicity and much of its utility in dissipative media.) Using the same trigonometric identity, we find that

$$y_t = 2\cos\left(\frac{\omega_1 - \omega_2}{2} t - \frac{\phi_1 - \phi_2}{2}\right)\cos\left(\frac{\omega_1 + \omega_2}{2} t - \frac{\phi_1 - \phi_2}{2}\right) \tag{2.25}$$

Rewriting the beat factor in terms of a time delay $t_g$, we now have

$$\cos\left[\frac{\omega_1-\omega_2}{2}\left(t-t_g\right)\right]=\cos\left(\frac{\omega_1-\omega_2}{2}t-\frac{\phi_1-\phi_2}{2}\right) \tag{2.26}$$

$$\left(\omega_1-\omega_2\right)t_g=\phi_1-\phi_2 \tag{2.27}$$

$$t_g=\frac{\phi_1-\phi_2}{\omega_1-\omega_2}=\frac{\Delta\phi}{\Delta\omega} \tag{2.28}$$

For a continuum of frequencies, the group delay is

$$t_g=\frac{d\phi}{d\omega} \tag{2.29}$$

This represents the true signal (baseband signal) delay, and is also referred to as the envelope delay.

## 2.6 Immittance Inverter

Immittance Inverters are of two types, Impedance Inverter and Admittance inverter. The following Block diagram shows an Immittance Inverter.

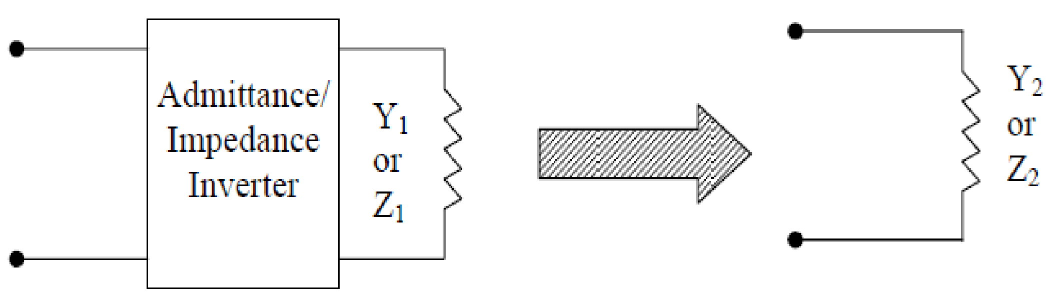


**Figure 2.2:** Immittance Inverter

An ideal Impedance Inverter is a two-port network that has a unique property at all frequencies, i.e. if it is terminated in impedance $Z_1$ on one port, the impedance $Z_2$ seen looking in the other port is

$$Z_2 = \frac{K^2}{Z_1} \tag{2.30}$$

Where $K$ is real and defined as characteristic impedance of the inverter. An impedance inverter converts a capacitance to inductance and vice versa. The ABCD matrix of the impedance inverter is

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} 0 & \mp jK \\ \pm \frac{1}{jK} & 0 \end{pmatrix} \tag{2.31}$$

Similarly an ideal Admittance Inverter is a two-port network that if terminated in admittance $Y_1$ on one port, the impedance $Y_2$ seen looking in the other port is

$$Y_2 = \frac{J^2}{Y_1} \tag{2.32}$$

where $J$ is real and defined as characteristic admittance of the inverter. Likewise an admittance inverter converts a capacitance to inductance and vice versa. The ABCD matrix of the admittance inverter is

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} 0 & \pm \frac{1}{jJ} \\ \mp jJ & 0 \end{pmatrix} \tag{2.33}$$

## 2.6.1 Properties of Immitance Inverter

If a series inductance is present between two Impedance Inverters, it looks like a shunt capacitance from its exterior terminals. [10]

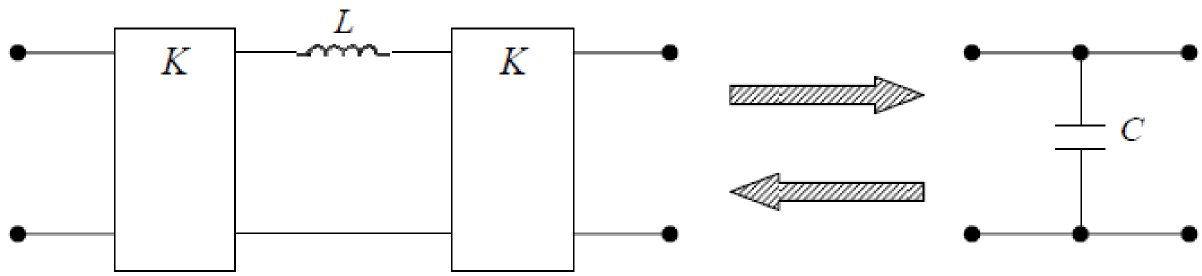


**Figure 2.3:** Immitance inverter used to convert a shunt capacitance into an equivalent circuit with series inductance.

Similarly if a shunt capacitance is present between two Admittance Inverters, it looks like a series inductance from its exterior terminals

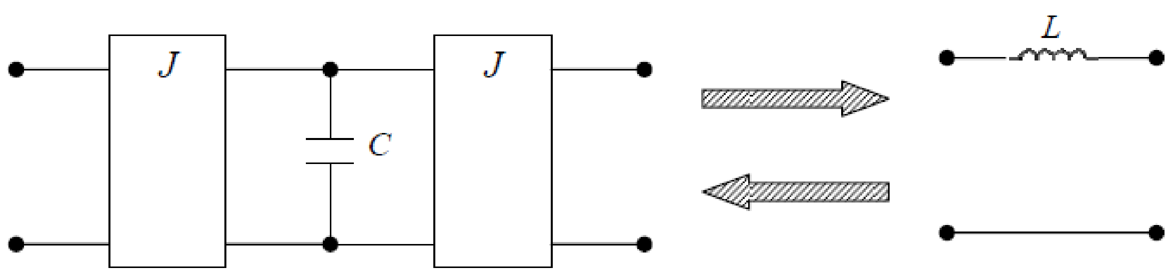


**Figure 2.4:** Immitance inverter used to convert a series inductance into a equivalent circuit with shunt capacitance.

Making use of the properties of Immittance Inverters, bandpass filters may be realized by series LC. Resonant circuits separated by Impedance Inverters (*K*) or shunt LC. Parallel resonant circuits separated by Admittance Inverters (*J*).

# Chapter 3

# Microstrip Basics

A general microstrip structure is shown in the figure 3.1; a microstrip transmission line consists of a thin conductor strip over a dielectric substrate along with a ground plate at the bottom of the dielectric.

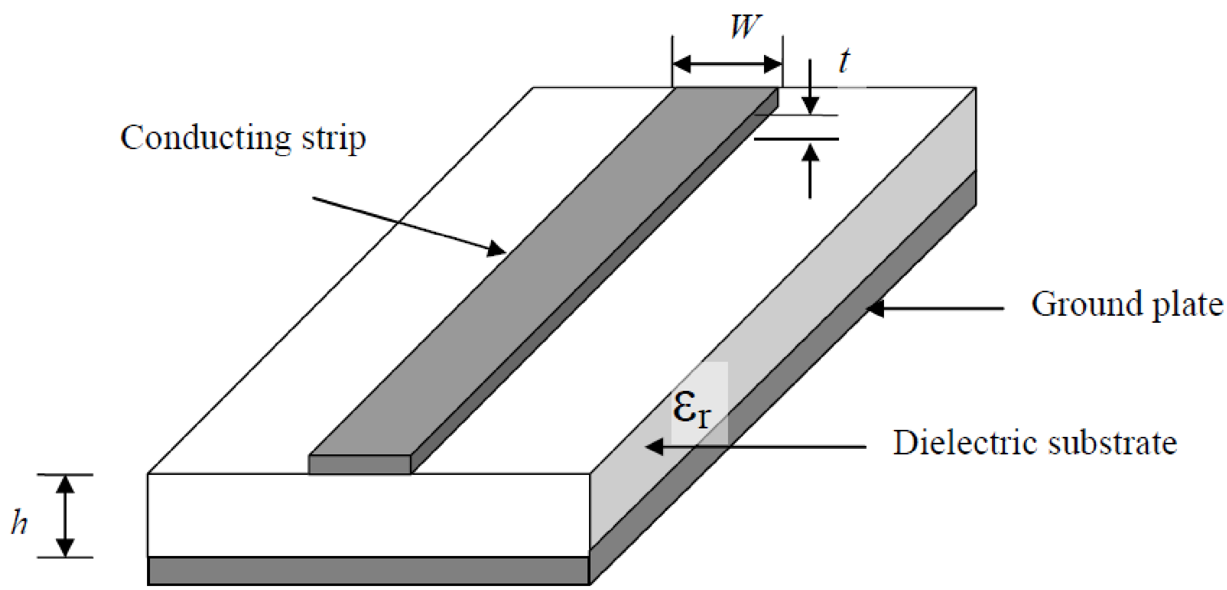


**Figure 3.1:** A basic Microstrip structure [2]

## 3.1 Waves in Microstrip Line

Wave traveling in microstrip line not only travels in the dielectric medium they also travel in the air media above the microstrip line. Thus they don't support pure TEM waves. In pure TEM transmission, the waves have only transverse component and the propagation velocity only depends on the permittivity and the permeability of the substrate. But in the case of microstrip line the magnetic and electric field also contains a longitudinal component, and their propagation velocity is dependent on the physical dimensions of the Microstrip as well.

If this longitudinal component is much smaller than the transverse component then the microstrip line can be approximated to TEM model. And this is called quasi TEM approximation.

## 3.2 Effective Dielectric Constant

Due to presence of two dielectric medium, air and the substrate, the effective dielectric constant replaces the relative dielectric constant of the substrate in the quasi TEM approximation. This effective dielectric constant is given in terms of $C_d$ capacitance per unit length with the dielectric substrate present and $C_a$, capacitance per unit length with dielectric constant replaced by air and is given by: [1]

$$\varepsilon_e = \frac{C_d}{C_a} \tag{3.1}$$

The effective dielectric constant of a microstrip line is given approximately by:

$$\varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(\frac{1}{\sqrt{1 + 12d/W}}\right) \tag{3.2}$$

## 3.3 Characteristic impedance

Characteristic impedance of the microstrip line is given by

$$z_c = \frac{1}{c\sqrt{C_a C_d}} \tag{3.3}$$

Where c is the velocity of electromagnetic waves in free space

Given the dimension of the microstrip line, the characteristic impedance can be calculated as:

$$Z_o = \frac{60}{\sqrt{\varepsilon_e}} \ln\left(\frac{8d}{W} + \frac{W}{4d}\right) \qquad \text{for } W/d \le 1 \qquad (3.4)$$

$$= \frac{120\pi}{\sqrt{\varepsilon_e\left[W/d + 1.393 + 0.667\ln\left(W/d + 1.444\right)\right]}} \qquad \text{for } W/d \ge 1 \qquad (3.5)$$

## 3.4 Some Other Formulae

- **W/d**

$$\frac{W}{d} = \frac{8e^{A}}{e^{2A} - 2} \qquad \text{For W/d < 2} \qquad (3.6)$$

$$= \frac{2}{\pi}\left[B - 1 - \ln(2B-1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left\{\ln(B-1) + 0.39 - \frac{0.61}{\varepsilon_r}\right\}\right] \qquad \text{for W/d >2} \qquad (3.7)$$

Where

$$A = \frac{Z_o}{60}\sqrt{\frac{\varepsilon_r + 1}{2}} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left(0.23 + \frac{0.11}{\varepsilon_r}\right) \qquad (3.8)$$

$$B = \frac{377\pi}{2Z_o\sqrt{\varepsilon_r}} \qquad (3.9)$$

- **Guided Wavelength**

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_e}} \qquad (3.10)$$

Where $\lambda_0$ is the free space wavelength at frequency f.

➢ **Propagation Constant**

$$\beta = \frac{2\pi}{\lambda_g} \tag{3.11}$$

➢ **Phase Velocity**

$$v_p = \frac{\omega}{\beta} = \frac{c}{\sqrt{\varepsilon_e}} \tag{3.12}$$

➢ **Electrical Length**

$$\theta = \beta l \tag{3.13}$$

$\theta$ is called the electrical length whereas $l$ is the physical length of the microstrip. Thus, $\theta = \frac{\pi}{2}$ when $l = \frac{\lambda_g}{4}$ and $\theta = \pi$ when $l = \frac{\lambda_g}{2}$ these are called quarter wavelength and half wavelength microstrip line and are important in the filter design. [1]

## 3.5 Surface Waves and Higher-Order Modes

Despite the absence of the top conductor there exists wave on ground plate guided by the air dielectric medium. These are called surface waves. The frequency at which these become significantly large is

$$f_s = \frac{c \tan^{-1} \varepsilon_r}{\sqrt{2}\pi h \sqrt{\varepsilon_r - 1}} \tag{3.14}$$

Where the phase velocity of the two modes become equal.

To avoid excitation of higher-order modes in Microstrip the frequency of operation is kept below the cut off frequency

$$f_c = \frac{c}{\sqrt{\varepsilon_r}\left(2W + 0.8h\right)} \tag{3.15}$$

## 3.6 Coupled Lines

The following figure shows the cross section of a coupled line. Widely used in the construction of filters, they support two modes of excitation, even and odd mode.

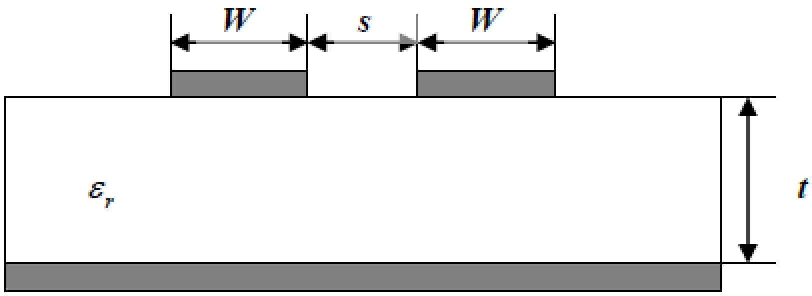


**Figure 3.2:** A Coupled Line Structure.

### 3.6.1 Even Mode

In even mode excitation both the microstrip coupled lines have the same voltage potential resulting in a magnetic wall at the symmetry plane

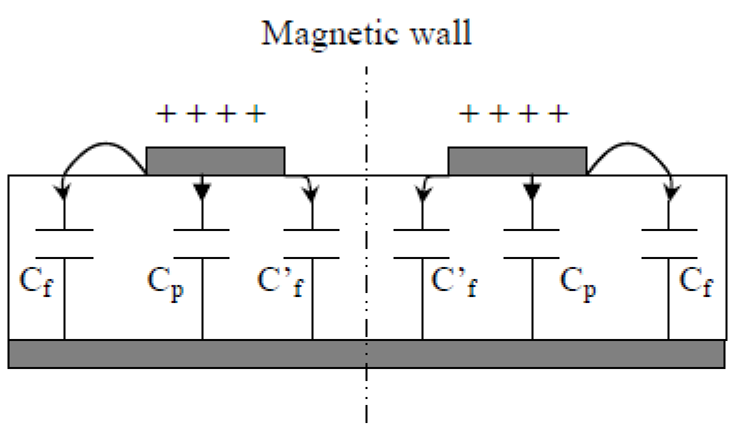


**Figure 3.3:** Quasi-TEM, Even Mode of a Pair of Coupled Microstrip Lines

Even mode capacitance is given by:

$$C_e = C_p + C_f + C_f^{'} \tag{3.16}$$

Where $C_p$ is the parallel plate capacitance between the microstrip line and the ground plate. Hence

$$C_p = \varepsilon_0 \varepsilon_r \frac{W}{d} \tag{3.17}$$

$C_f$ is the fringe capacitance and is given by

$$2C_f = \frac{\sqrt{\varepsilon_e}}{cZ_0} - C_p \tag{3.18}$$

And $C_f^{'}$ is the modified fringe capacitance, with the effect of the adjacent microstrip included

$$C_f^{'} = \frac{C_f}{1 + A\left(h/s\right)\tanh\left(8s/h\right)} \tag{3.19}$$

Where

$$A = \exp\left[-0.1\exp\left(2.33 - 2.53 W/h\right)\right] \tag{3.20}$$

The even mode characteristic impedance can also be obtained from the capacitance

$$Z_{ce} = \left(c\sqrt{C_e C_e^a}\right)^{-1} \tag{3.21}$$

Where $C_e^a$ is the even mode capacitance with air as a dielectric and the effective dielectric constant for even mode is given as:

$$\varepsilon_e^a = \frac{C_e}{C_e^a} \tag{3.22}$$

### 3.6.2 Odd Mode

In odd mode the coupled microstrip line possess opposite potential. This results into a electric wall at the symmetry. The following cross section diagram shows the same.

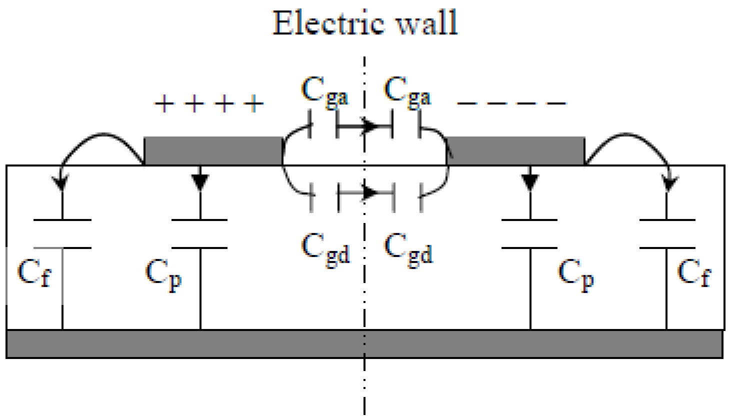


**Figure 3.4:** Quasi-TEM, Odd Mode of a Pair of Coupled Microstrip Lines

The resulting odd mode capacitance is given as

$$C_o = C_p + C_f + C_{gd} + C_{ga} \tag{3.23}$$

$C_{ga}$ and $C_{gd}$ represent fringe capacitance between the two microstrip line over the dielectric .

$$C_{ga} = \varepsilon_o \frac{K\left(k'\right)}{K\left(k\right)} \tag{3.24}$$

Where

$$k = \frac{s/h}{s/h + \frac{2W}{h}} \tag{3.25}$$

$$k' = \sqrt{1-k^2} \tag{3.26}$$

And the ratio of the elliptic function $\frac{K\left(k^{'}\right)}{K\left(k\right)}$ is given by

$$\frac{K\left(k^{'}\right)}{K\left(k\right)} = \frac{1}{\pi}\ln\left(2\frac{1+\sqrt{k^{'}}}{1-\sqrt{k^{'}}}\right) \qquad for\, 0 \le k^2 \le 0.5 \tag{3.27}$$

$$= \frac{\pi}{\ln\left(2\frac{1+\sqrt{k}}{1-\sqrt{k}}\right)} \qquad for\, 0.5 \le k^2 \le 1 \tag{3.28}$$

The odd mode characteristic impedance and effective dielectric constant is given as:

$$Z_{co} = \left(c\sqrt{C_o C_o^a}\right)^{-1} \tag{3.29}$$

Where $C_e^a$ is the even mode capacitance with air as a dielectric.

$$\varepsilon_e^o = \frac{C_o}{C_o^a} \tag{3.30}$$

# Chapter 4

## Filter basics

### 4.1 Classification of Filters

In general filters are the electronics devices which are used to select interest of application or reject any particular frequency band.

There are four types of filters they are

- Low Pass Filters
- High Pass Filters
- Band Pass Filters
- Band Stop Filters

Low pass filter allows low frequency signals to be transmitted from the input to the output with little attenuation. However, as the frequency exceeds a certain cut off point, the attenuation increases significantly with the result of delivering amplitude reduced signal to the output port.

The opposite behaviour is true for a high pass filter, where the low frequency signal components are highly attenuated in amplitude, while beyond a cut-off frequency point the signal passes the filter with little attenuation. Band pass and band stop filters restrict the pass band between specific lower and upper frequency points where the attenuation is either low (band pass) or high for (band stop) compared to the remaining frequency band. Figure summarizes their gain ‘v’ and normalized frequency ‘f’ behaviour.

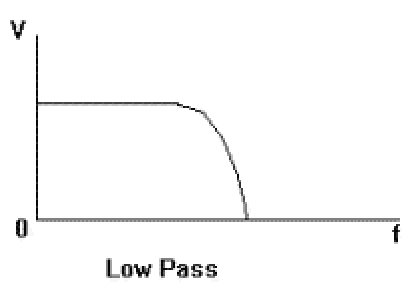


**Figure 4.1:** Low pass filer response

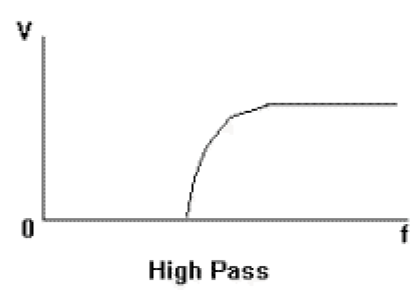


**Figure 4.2:** High pass filer response

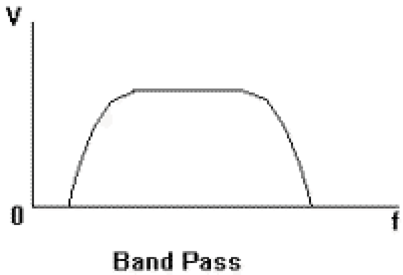


**Figure 4.3:** Band pass filer response

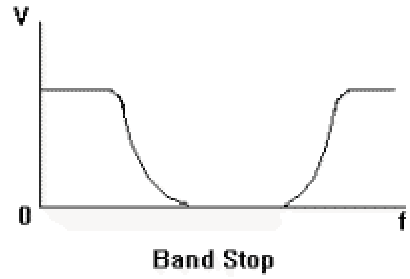


**Figure 4.4:** Band stop filer response

### 4.1.1 Butterworth Filter:

The transfer function of a two-port filter network is a mathematical description of network response characteristics, namely, a mathematical expression of $S_{21}$. On many occasions, an amplitude-squared transfer function for a lossless passive filter network is defined as:

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\varepsilon^2 F^2{}_n(\Omega)} \tag{4.1}$$

Where ε is a ripple constant, $F_n(\Omega)$ represents a filtering or characteristic function, and Ω is a frequency variable. For our discussion here, it is convenient to let Ω represent a radian frequency variable of a low pass prototype filter that has a cutoff frequency at $\Omega = \Omega_c$ for $\Omega_c = 1$ (rad/s). The amplitude-squared transfer function for Butterworth filters that have an insertion loss $L_{Ar}$= 3.01 dB at the cutoff frequency $\Omega_c$= 1 is given by

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\Omega^{2n}} \tag{4.2}$$

Where $n$ is the degree or the order of filter, which corresponds to the number of reactive elements, required in the lowpass prototype filter. This type of response is also referred to as maximally flat because its amplitude-squared transfer function defined has the maximum number of $(2n - 1)$ zero derivatives at Ω= 0. Therefore, the maximally flat approximation to the ideal lowpass filter in the pass band is best at Ω =0, but deteriorates as Ω approaches the cutoff frequency $\Omega_c$. Figure 4.5 shows a typical maximally flat response

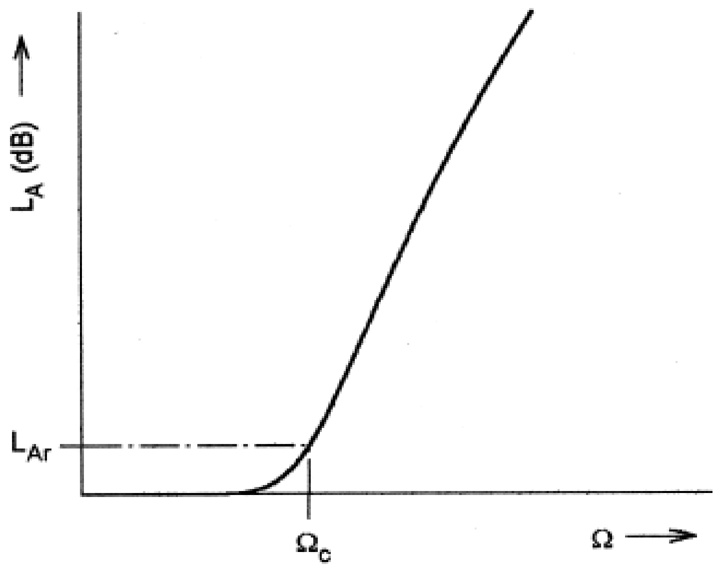


**Figure 4.5:** Butterworth low pass filter response [1]

### 4.1.2 Chebyshev Filter

The Chebyshev response that exhibits the equiripple pass band and maximally flat stop band is depicted in Figure 4.6. The amplitude-squared transfer function that describes this type of response is

$$|S_{21}(j\Omega)|^2 \quad = \quad \frac{1}{1+\varepsilon^2 T^2{}_n(\Omega)} \tag{4.3}$$

Where the ripple constant ε is related to a given pass band ripple $L_{Ar}$ in dB by

$$\mathcal{E} = \sqrt{10^{L_{Ar}/10}}\text{-}1 \tag{4.4}$$

$T_n(\Omega)$ is a Chebyshev function of the first kind of order *n*, which is defined as:

$$T_n(\Omega) = \begin{cases} \cos(cos^{-1}\Omega)\ |\Omega| \leq 1 \\ \cos(n\ cosh^{-1}\Omega)\ |\Omega| \geq 1 \end{cases} \tag{4.5}$$

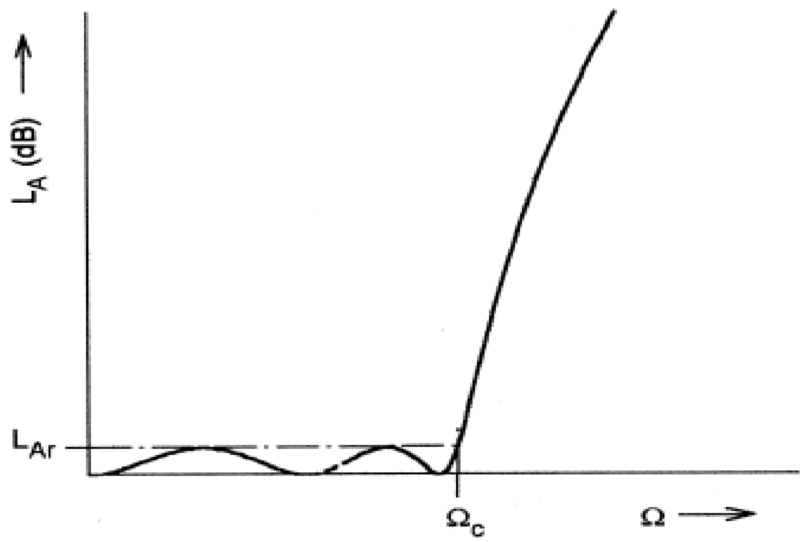


**Figure 4.6:** Chebyshev low pass filter response [1]

### 4.1.3 Elliptic Filter:

The response that is equal-ripple in both the passband and stopband is the elliptic function response, as illustrated in Figure 4.7. The transfer function for this type of response is

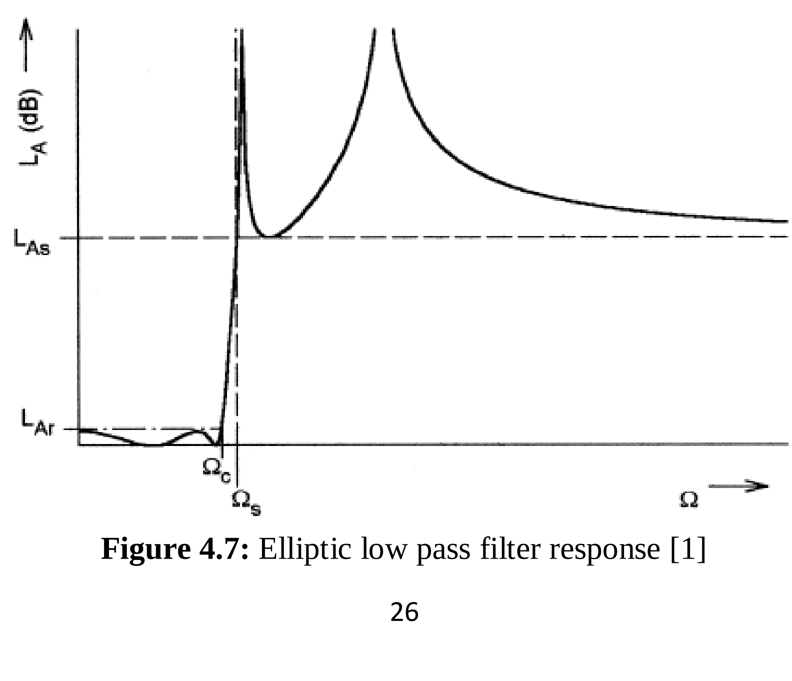


**Figure 4.7:** Elliptic low pass filter response [1]

$$|S_{21}(j\Omega)|^2 = \frac{1}{1+\varepsilon^2 F^2{}_n(\Omega)} \tag{4.6}$$

With

$$F^2{}_n(\Omega) = \begin{cases} M \dfrac{\prod_{t=1}^{\frac{n}{2}} \Omega_t^2 - \Omega^2}{\prod_{t=1}^{\frac{n}{2}} \frac{\Omega_s^2}{\Omega_t^2} - \Omega^2} & \text{for n even} \\ \\ N \dfrac{\prod_{t=1}^{\frac{n-1}{2}} \Omega_t^2 - \Omega^2}{\prod_{t=1}^{\frac{n-1}{2}} \Omega_s^2/\Omega_t^2 - \Omega^2} & \text{for n odd} \end{cases} \tag{4.7}$$

Where $\Omega_t (0 < \Omega_t < 1)$ and $\Omega_s > 1$ represent some critical frequencies; *M* and *N* are constants.

# Chapter 5

# Design of Parallel coupled line filter

## 5.1 Design of Third Order Chebyshev Bandpass Parallel Coupled Filter

For an $N^{th}$ order Chebyshev filter, the transfer function is given as

$$|H(s)|^2 = \frac{H_0}{1+\varepsilon^2 C_N^2\left(\frac{\omega}{\omega_c}\right)} \tag{5.1}$$

Where $H_0$ is the dc attenuation, $\varepsilon$ is the ripple magnitude, $\omega_c$ is the 3- dB corner frequency and the Chebyshev polynomial is given by

$$C_N(\omega) = \cos\left[N\cos^{-1}(\omega)\right] \tag{5.2}$$

The poles of the Chebyshev transfer function lie on an ellipse and are given as

$$s_k = \omega_c(\sigma_k + j\omega_k) \tag{5.3}$$

Where k= 1, 2 …2N for an Nth order filter. $\sigma_k$ and $\omega_k$ are given as

$$\sigma_k = -\sinh(a)\sin\left[\frac{(2k-1)\pi}{2N}\right] \tag{5.4}$$

$$\omega_k = \cosh(a)\cos\left[\frac{(2k-1)\pi}{2N}\right] \tag{5.5}$$

$$a = \frac{1}{N}\sinh^{-1}\left(\frac{1}{\varepsilon}\right) \tag{5.6}$$

The component values for a Chebyshev lowpass prototype are determined using the following equations: [2-4]

$$g_0 = 1 \tag{5.7}$$

$$g_1 = \frac{2a_1}{\gamma} \tag{5.8}$$

$$g_k = \frac{4a_{k-1}a_k}{b_{k-1}g_{k-1}}, k = 2, 3, ........N \tag{5.9}$$

$$g_{N+1} = 1 \qquad \text{for } N \text{ odd} \tag{5.10}$$

$$g_{N+1} = \coth^2\left(\beta/4\right) \quad \text{for N even} \tag{5.11}$$

Where

$$a_k = \sin\left[\frac{(2k-1)\pi}{2N}\right], \; k = 1, 2, .......N \tag{5.12}$$

$$b_k = \gamma^2 + \sin^2\left(\frac{k\pi}{N}\right), \quad k = 1, 2, .......N \tag{5.13}$$

$$\beta = \ln\left[\coth\left(\frac{A}{17.372}\right)\right] \tag{5.14}$$

$$\gamma = \sinh\left(\frac{\beta}{2N}\right) \tag{5.15}$$

Where $A$ is the passband ripple in decibel given by

$$A = 10\log\left(1+\varepsilon^2\right) \tag{5.16}$$

Hence, the information required while designing a filter is the order of the filter, the ripple factor in the passband, and the center frequency. The bandwidth is required while designing bandpass filters. We refer to the filter tables given in D.M. Pozar [1] to find the following coefficients for a third order Chebyshev filter.

$$g_1 = 1.5963$$
$$g_2 = 1.0967$$
$$g_3 = 1.5963$$
$$g_4 = 1.000$$

These values are for a lowpass prototype design with source and load impedances equal to unity. The coefficients are nothing but inductances and capacitances of a lowpass filter ladder network as shown in figure 5.1. For filter with $50\Omega$ impedances at the source and load, the values of inductances and Capacitances need to be scaled. The design is then transformed to a bandpass version by converting s to

$$s_{bp} = \frac{s^2 + \omega_0^2}{sBW} \tag{5.17}$$

$s_{bp}$ is the converted complex angular frequency of the bandpass filter, $BW$ is the filter bandwidth, and $\omega_0 = 2\pi f_0$ , where $f_0$ is the center frequency. Hence, the transfer function

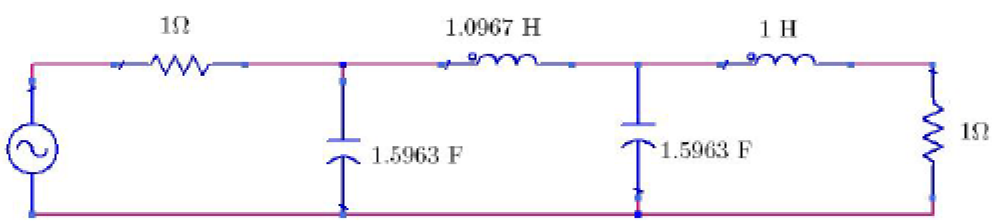


**Figure 5.1:** A ladder network for a third order lowpass Chebyshev filter prototype beginning with a shunt element.

Of this bandpass filter $H_{bp}$ is

$$H_{bp}\left(s_{bp}\right)=H\left(\frac{s^2+\omega_0^2}{sBW}\right) \tag{5.18}$$

The fractional bandwidth of the filter given by

$$\Delta=\frac{f_2-f_1}{f_0} \tag{5.19}$$

Where

$f_1$ is the lower cut-off frequency,

$f_2$ is the higher cut-off frequency

Using the concept of admittance inverters, for an Nth order filter, the admittance inverter parameters given by Pozar [1] are reproduced as follows [4-7]:

for $n=1$

$$Z_0J_1=\sqrt{\frac{\pi\Delta}{2g_1}}; \tag{5.20}$$

for $n=2,3,....N$

$$Z_0J_n=\frac{\pi\Delta}{2\sqrt{g_{n-1}g_n}}; \tag{5.21}$$

and for $n=N+1$

$$Z_0J_{N+1}=\sqrt{\frac{\pi\Delta}{2g_Ng_{N+1}}}; \tag{5.22}$$

Where $J_n$ is the admittance inverter constant for the nth section and $Z_0$ is the characteristic impedance of the filter.

Now, the even and odd mode impedances of the coupled line $Z_{0e}$ and $Z_{0o}$, respectively, are computed using following equations:

$$Z_{0e} = Z_0\left[1 + J_n Z_0 + \left(J_n Z_0\right)^2\right] \tag{5.23}$$

$$Z_{0o} = Z_0\left[1 - J_n Z_0 + \left(J_n Z_0\right)^2\right] \tag{5.24}$$

for $n = 1, 2, ....N+1$ . From these values of $Z_{0e}$ and $Z_{0o}$, the widths of the coupled microstrip line are calculated for each section. The lengths are calculated from the frequency and effective dielectric constant information.

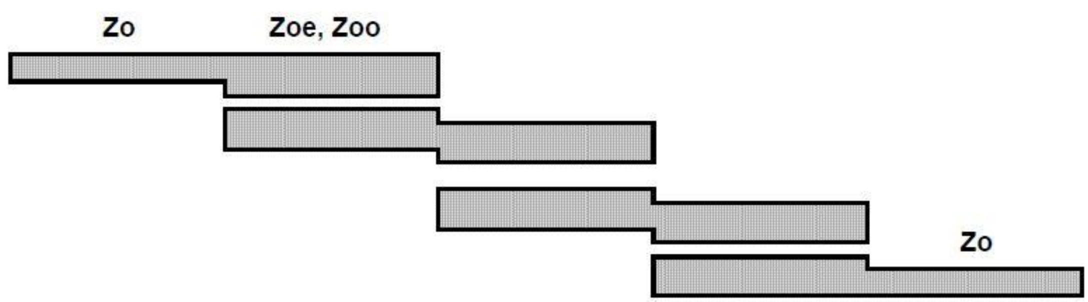


**Figure 5.2:** General structure of parallel-coupled microstrip bandpass filter [1]

## 5.2 The Quarter-wave Transformer

The input impedance of a filter may not be $50\Omega$ at the desired center frequency. If it is terminated into 50 ohm transmission lines at either ends, there would be a lot of reaction, deteriorating the receiver performance. To match a filter to $50\Omega$ termination impedances, a microstrip line of quarter wavelength or an electrical length of $90^\circ$ is used. Consider a quarter-wave transformer having a characteristic impedance of $Z_0$ terminated in load impedance $Z_L$, which is the input impedance of the filter. The input impedance of the overall network $Z_{in}$ is then given by

$$Z_{in} = Z_0 \frac{Z_L \cos(\beta l) + jZ_0 \sin(\beta l)}{Z_0 \cos(\beta l) + jZ_L \sin(\beta l)} \tag{5.25}$$

Since the length of the transformer is $\lambda / 4$ , substituting $\beta l = \pi / 2$ , we get

$$Z_{in} = Z_0 \frac{Z_0}{Z_L} \tag{5.26}$$

$$Z_0 = \sqrt{Z_{in} Z_L} \tag{5.27}$$

which is the characteristic impedance of the matching transformer. The microstrip line of this impedance can be designed using the information about the properties of the board. [2]

# Chapter 6

# DESIGN AND SIMULATION

## 6.1 Filter Specifications:

- Centre frequency $f_c$ of 1.575 GHz.
- Insertion loss: less than 2.5 dB
- Bandwidth: 20%
- Passband Ripple: 0.5dB
- Source Impedance = Load impedance = 50 ohm.

## 6.2 Design Procedure:

### 6.2.1 Low Pass Filter Prototype:

First step is to take standard lowpass prototype parameters

| | 0.5 dB Ripple | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| $N$ | $g_1$ | $g_2$ | $g_3$ | $g_4$ | $g_5$ | $g_6$ | $g_7$ | $g_8$ | $g_9$ | $g_{10}$ | $g_{11}$ |
| 1 | 0.6986 | 1.0000 | | | | | | | | | |
| 2 | 1.4029 | 0.7071 | 1.9841 | | | | | | | | |
| 3 | 1.5963 | 1.0967 | 1.5963 | 1.0000 | | | | | | | |
| 4 | 1.6703 | 1.1926 | 2.3661 | 0.8419 | 1.9841 | | | | | | |
| 5 | 1.7058 | 1.2296 | 2.5408 | 1.2296 | 1.7058 | 1.0000 | | | | | |
| 6 | 1.7254 | 1.2479 | 2.6064 | 1.3137 | 2.4758 | 0.8696 | 1.9841 | | | | |
| 7 | 1.7372 | 1.2583 | 2.6381 | 1.3444 | 2.6381 | 1.2583 | 1.7372 | 1.0000 | | | |
| 8 | 1.7451 | 1.2647 | 2.6564 | 1.3590 | 2.6964 | 1.3389 | 2.5093 | 0.8796 | 1.9841 | | |
| 9 | 1.7504 | 1.2690 | 2.6678 | 1.3673 | 2.7239 | 1.3673 | 2.6678 | 1.2690 | 1.7504 | 1.0000 | |
| 10 | 1.7543 | 1.2721 | 2.6754 | 1.3725 | 2.7392 | 1.3806 | 2.7231 | 1.3485 | 2.5239 | 0.8842 | 1.9841 |

**Figure 6.1**: Element value for equal- Ripple Low-Pass filter prototype ( $g_0 = 1, \omega_c = 1, N = 1$ to 10) 0.5 dB ripple. [1]

| N | $g_1$ | $g_2$ | $g_3$ | $g_4$ | $g_5$ | $g_6$ | $g_7$ | $g_8$ | $g_9$ | $g_{10}$ | $g_{11}$ |
|---|---|---|---|---|---|---|---|---|---|---|---|
| 1 | 2.0000 | 1.0000 | | | | | | | | | |
| 2 | 1.4142 | 1.4142 | 1.0000 | | | | | | | | |
| 3 | 1.0000 | 2.0000 | 1.0000 | 1.0000 | | | | | | | |
| 4 | 0.7654 | 1.8478 | 1.8478 | 0.7654 | 1.0000 | | | | | | |
| 5 | 0.6180 | 1.6180 | 2.0000 | 1.6180 | 0.6180 | 1.0000 | | | | | |
| 6 | 0.5176 | 1.4142 | 1.9318 | 1.9318 | 1.4142 | 0.5176 | 1.0000 | | | | |
| 7 | 0.4450 | 1.2470 | 1.8019 | 2.0000 | 1.8019 | 1.2470 | 0.4450 | 1.0000 | | | |
| 8 | 0.3902 | 1.1111 | 1.6629 | 1.9615 | 1.9615 | 1.6629 | 1.1111 | 0.3902 | 1.0000 | | |
| 9 | 0.3473 | 1.0000 | 1.5321 | 1.8794 | 2.0000 | 1.8794 | 1.5321 | 1.0000 | 0.3473 | 1.0000 | |
| 10 | 0.3129 | 0.9080 | 1.4142 | 1.7820 | 1.9754 | 1.9754 | 1.7820 | 1.4142 | 0.9080 | 0.3129 | 1.0000 |

**Figure 6.2:** Element values for maximally Flat Low- Pass Filter Prototype ( $g_0 = 1, \omega_c = 1, N = 1$ to 10) [1]

We have taken N=3 for this low pass prototype parameter are $g_0 = 1 = g_4$ , $g_2 = 1.0967$ for Chebyshev and $g_0 = 1 = g_4$ , $g_1 = 1.0000 = g_3$ , $g_2 = 2.0000$ for Butterworth response. Also the characteristic impedance of the connecting feed line is taken as 50 ohm.

### 6.2.2 Even and Odd Mode Calculations:

Now the normalized J-inverter values are calculated using the following equations [4-7]

For $n = 1$

$$Z_0 J_1 = \sqrt{\frac{\pi\Delta}{2g_1}}; \tag{6.1}$$

For $n = 2,3,....N$

$$Z_0 J_n = \frac{\pi\Delta}{2\sqrt{g_{n-1}g_n}}; \tag{6.2}$$

And for $n = N+1$

$$Z_0 J_{N+1} = \sqrt{\frac{\pi\Delta}{2g_N g_{N+1}}}; \tag{6.3}$$

Where

$\Delta = \dfrac{f_2 - f_1}{f_0}$  $\Delta$ = fractional bandwidth of the filter

Where

$f_1$ is the lower cut-off frequency,

$f_2$ is the higher cut-off frequency,

Then to realize the J-inverter values obtained from the above equations, the even and odd mode impedances are calculated using the following
Equations

$$Z_{0e} = Z_0\left[1 + J_n Z_0 + \left(J_n Z_0\right)^2\right] \tag{6.4}$$

$$Z_{0o} = Z_0\left[1 - J_n Z_0 + \left(J_n Z_0\right)^2\right] \tag{6.5}$$

By calculating even and odd impedances of all the section of the parallel coupled line band pass filter we get the following results:

**TABLE 6.1: Circuit design parameters of the three-pole, Parallel Coupled Chebyshev filter**

| $J$ | Admittance Inverter $J_{j,j+1}/Y_0$ | Even- mode impedance $Z_{0e}$ $(\Omega)$ | Odd-mode impedance $Z_{0o}$ $(\Omega)$ |
|---|---|---|---|
| 0 | 0.4436 | 82.02 | 37.659 |
| 1 | 0.2374 | 64.69 | 40.947 |
| 2 | 0.2374 | 64.69 | 40.947 |
| 3 | 0.4436 | 82.02 | 37.659 |

**TABLE 6.2: Circuit design parameters of the three- pole, Parallel Coupled Butterworth filter**

| $J$ | Admittance Inverter $J_{j,j+1}/Y_0$ | Even- mode impedance $Z_{0e}$ $(\Omega)$ | Odd-mode impedance $Z_{0o}$ $(\Omega)$ |
|---|---|---|---|
| 0 | 0.5605 | 93.7330 | 37.6830 |
| 1 | 0.2220 | 63.5642 | 41.3642 |
| 2 | 0.2220 | 63.5642 | 41.3642 |
| 3 | 0.5605 | 93.7330 | 37.659 |

### 6.2.3 Geometrical parameters of the coupled line filter

Next the physical parameters of the quarter wavelength resonant filter are calculated by the LineCalc tool present in ADS software: a microstrip line component is added to both sides of the filter whose characteristic impedance is 50 ohm. Also the length and width of that is also calculated with the LineCalc tool. The LineCalc tool embedded in ADS can help calculate the actual dimension of the microstrip if the odd and even resistances were given in a certain frequency as well as the substrate was specified [8-12].

**Substrate Properties:**

Substrate is taken having relative dielectric of 9.6 and thickness 1.27 mm.

Following are the values calculated:

**TABLE 6.3: Geometrical Parameters of the Coupled Line Chebyshev filter**

| Line Description | Width(mm) | Length(mm) | Gap(mm) |
|---|---|---|---|
| 50 Ω line | 1.223650 | 18.807500 | |
| Coupled Line 1 | 0.933981 | 19.514900 | 0.300508 |
| Coupled Line 2 | 1.102160 | 19.098200 | 0.744214 |
| Coupled Line 3 | 1.102160 | 19.098200 | 0.744214 |
| Coupled Line 4 | 0.933981 | 19.514900 | 0.300508 |

**TABLE 6.4: Geometrical Parameters of the Coupled Line Butterworth filter**

| Line Description | Width(mm) | Length(mm) | Gap(mm) |
|---|---|---|---|
| 50 Ω line | 1.223650 | 18.807500 | |
| Coupled Line 1 | 0.853784 | 19.72600 | 0.211043 |
| Coupled Line 2 | 1.113900 | 19.06760 | 0.809201 |
| Coupled Line 3 | 1.113900 | 19.06760 | 0.809201 |
| Coupled Line 4 | 0.853784 | 19.72600 | 0.211043 |

## 6.3 The Simulation of Coupled Microstrip Filter with ADS

ADS design software by Agilent Technology is used for simulation purpose

### 6.3.1 Schematic of coupled line Chebyshev Filter

In ADS Coupled line are selected and using LineCalc tool physical dimension of the corresponding Coupled line obtained, after that designed is modified by the calculated physical dimension. Also two 50 ohm transmission lines are also added on either side of the filter to match the filter with the other components. Physical parameter for this transmission line is also calculated by the LineCalc tool. [16], [17] The schematic of the above filter in ADS is as follows:

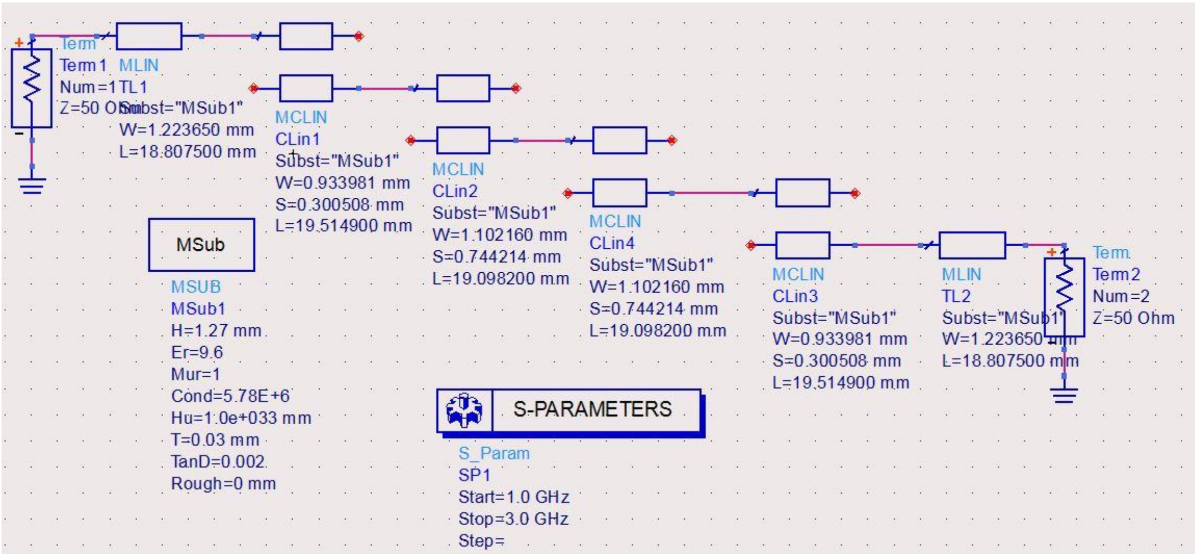


**Figure 6.3:** Schematic circuit of Coupled line Chebyshev filter

## 6.3.2 Schematic of coupled line Butterworth filter:

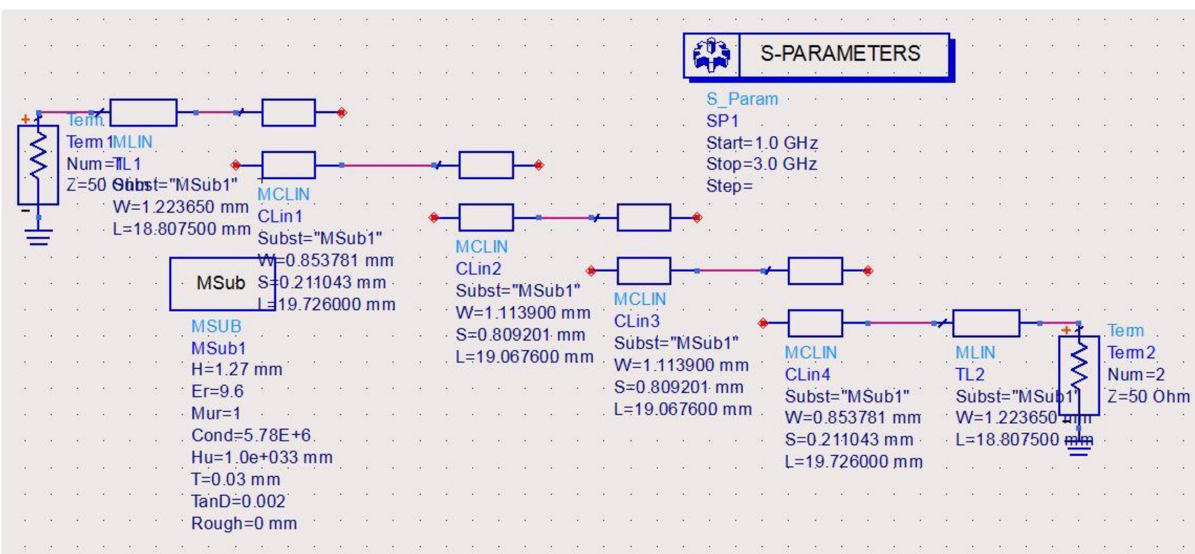


**Figure 6.4:** Schematic circuit of Coupled line Butterworth filter.

## 6.4 Results of Simulation:

### 6.4.1 Simulation result for Chebyshev filter:

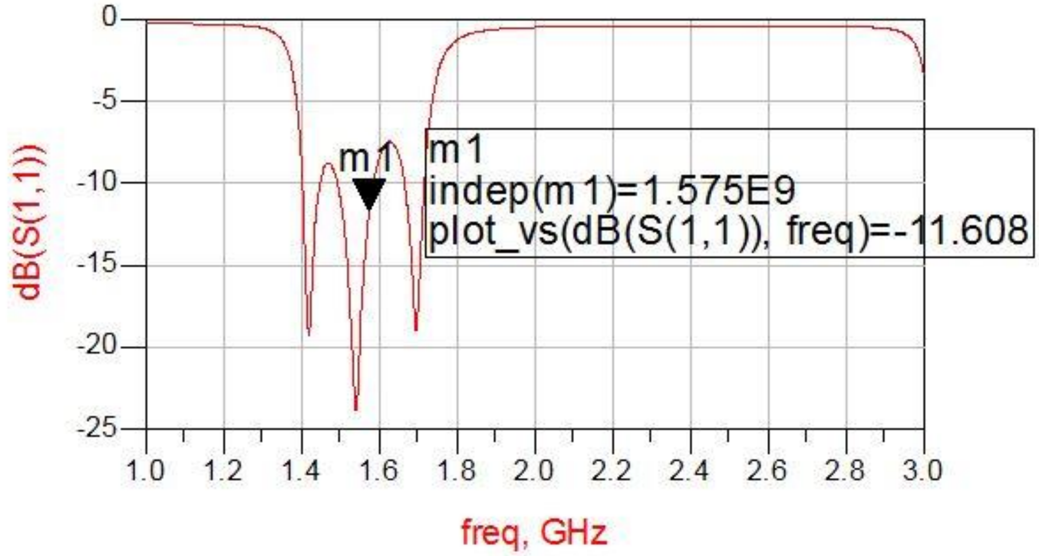


**Figure 6.5:** Simulated output of $S_{11}$ - parameter of the filter for Chebyshev response.

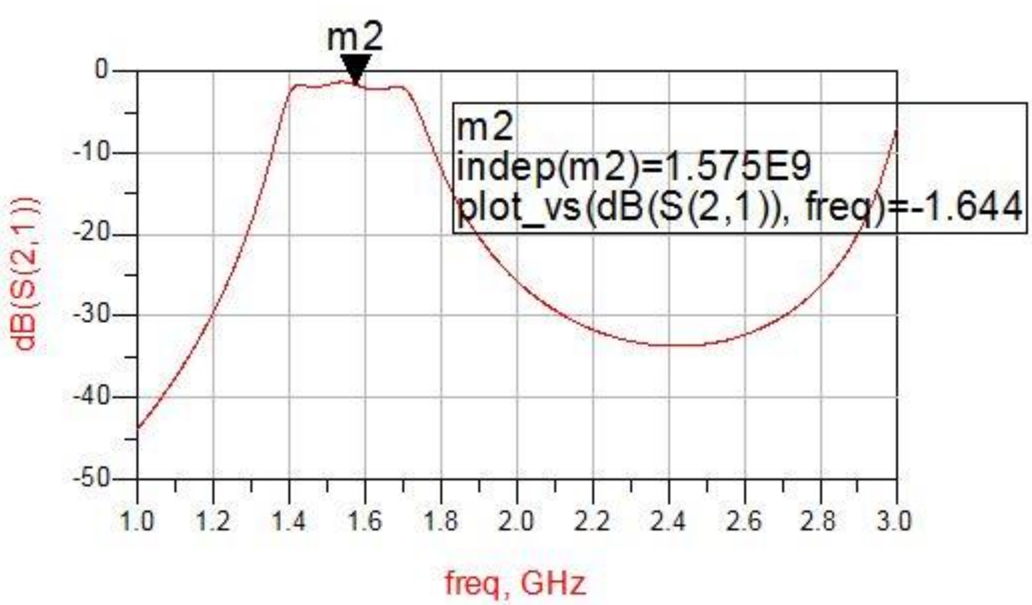


**Figure 6.6:** Simulated output of $S_{21}$ - parameter of the filter for Chebyshev response.

### 6.4.2 Simulation result for Butterworth Filter:

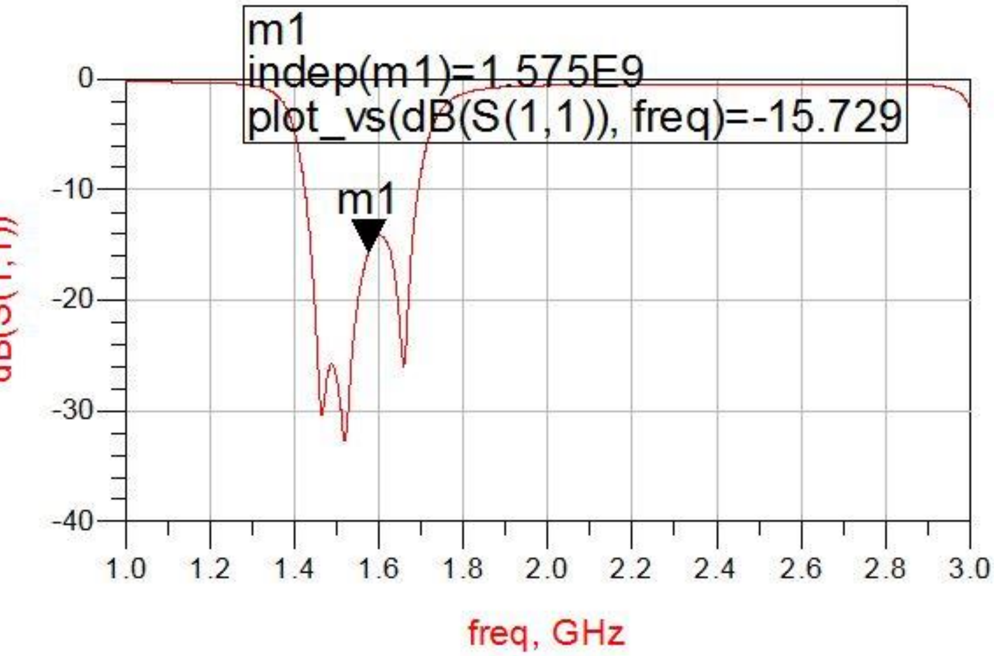


**Figure 6.7:** Simulated output of $S_{11}$ - parameter of the filter for Butterworth response

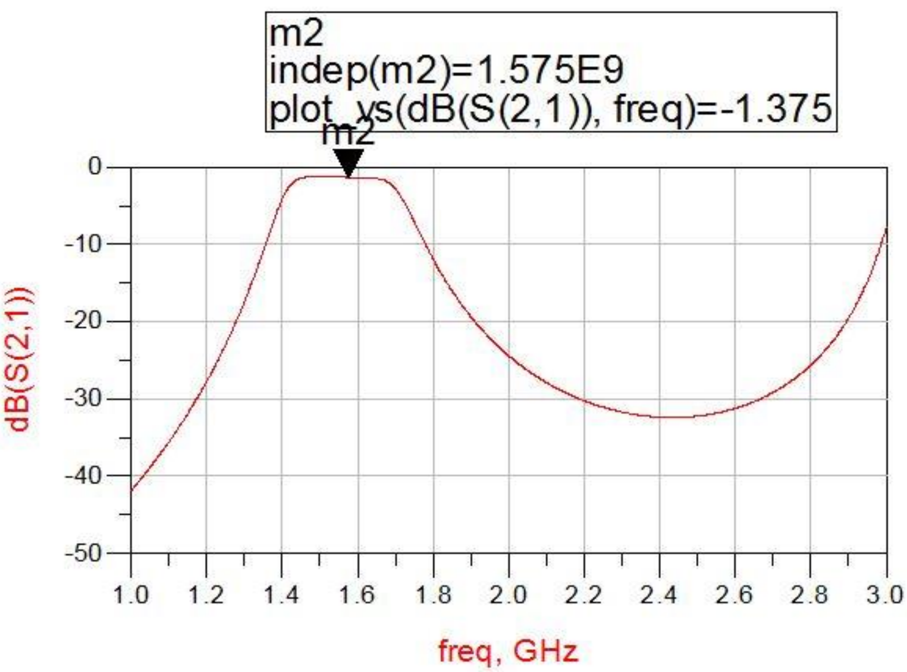


**Figure 6.8:** Simulated output of $s_{21}$ parameter of the filter for Butterworth response

# Chapter 7

# Result and Conclusion

## 7.1 Results of Designed Filters

Parallel coupled bandpass filter is designed for center frequency of 1.575 GHz. The simulated structure and S-parameter graphs are shown in previous chapter. In the S-parameter graphs frequency is plotted along x-axis in GHz and gain is plotted along y-axis in dB, from the graph it can be seen that the mid-band frequency is 1.575 GHz and fractional bandwidth is approximately 20%. In this thesis parallel coupled band pass filter is designed for Butterworth as well as Chebyshev response, for Chebyshev response $S_{11}$ and $S_{21}$ parameters are -11.608 dB and -1.644 dB respectively, and for Butterworth response $S_{11}$ and $S_{21}$ parameters are -15.729 dB and -1.375 dB respectively.

The S- parameter graph of figure 6.5 and figure 6.6 are for Chebyshev Response and hence we can see ripple in pass band region, while figure 6.7 and figure 6.8 are for Butterworth response hence graph is flat in the pass band region. Butterworth filter offers maximally flat characteristic, hence it is more suitable for filter operation. But in some cases Chebyshev response can be more helpful.

## 7.2 Conclusion

The filters are one of the primary and essential parts of the microwave system and any communication system. Any communication system cannot be designed without filters. Our designed parallel coupled microstrip bandpass filter operates at a center frequency of 1.575 GHz, which falls in the Microwave L-band applications especially for the Global positioning system. Global positioning system is a worldwide radio navigation system. GPS receivers are becoming miniaturized and are becoming very economical and this makes the technology accessible to virtually everyone. GPS receiver needs compact, low cost high performance bandpass filter. The parallel coupled band pass filter designed above fulfills the entire requirement for GPS system.

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