Ferroelectric Capacitor Based Tunable

Filter Circuits

by

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Filters are classified as: low pass, high pass, band-pass and band-stop. Band-stop filters are one of the key components in the RF front end of a microwave communication system. Band-stop filters or notch filters are useful to avoid unwanted signals and interference. The interfering signals may have shifting frequencies and therefore the tenability of bandpass is extremely important. Although a variety of tunable devices are available, recently ferroelectric capacitors are attracting the attention of many investigators, because of their high speed, polarity independent voltage tenability, small size, substrate independent integration, and absence of moving parts. In addition, present and future communication systems operate in multiple bands. Therefore there is necessity for dual and multiband filters.

The Dual Band Band-Stop Filter (DBBSF) is an important circuit component in microwave applications. DBBSFs are commonly employed in high power amplifiers and mixers to suppress the double sideband spectrum to reduce circuit size and cost. Design of Dual Band filters at microwave frequencies is still challenging since it has to take into consideration many parameters, including center frequency, bandwidth, tenability, return loss, and insertion loss. This dissertation focus on the implementation multi-band stop filters using ferroelectric capacitors.
In this work we will describe the motivation of our research, why we need to study tunable elements, and how we can use tunable rf-microwave devices. We will introduce tunable elements such as semiconductor varactors, Micro-Electro-Mechanical Systems (MEMS) devices, and Ferroelectric capacitors, such as Barium Strontium Titanate (BST).

Chapter two is a literature review of tunable circuits working in a single frequency-band, such as band-pass filter, and band-stop filter.

In chapter three, we design different circuits of tunable band-stop filters by using different methods to tune them. The advantage of using ferroelectric capacitor is that we can apply voltage in both polarities, but not over -5 to +5V. We design these circuits in first and third order, and we compare their performance in terms of return loss and insertion loss and bandwidth.

In chapter four, we explain tunable DBBSFs, and we design them by using two kinds of tunable elements, tuning diode and ferroelectric varactor. We use a different approach to connect those filters to get tuning for each branch independent to other branch and also tune both the bands simultaneously.
DEDICATION

I dedicate my dissertation work to

My loving parents,

whose words of encouragement have pushed me forward.

My wife, who has supported me to complete my studies.

My Kids, Saifalislam, Abdallah, and Taha, Who are my eyes in this life.

I also dedicate this dissertation

to my brothers, sisters and to all of my family members.

Finally, I would like to dedicate this dissertation to all of my friends.

A special dedication to my country, Libya.
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CHAPTER 1

TUNABLE ELEMENTS

1.1 Introduction

In recent years, there is an increasing interest in designing tunable microwave components with tunable devices. A tunable circuit is a circuit that enables the change of one or more of its parameters by using a voltage or a current control signal. For this reason, researchers developed several tunable components to get the best results. In this study, the researcher will explain their working principles, the advantages, and disadvantages of using them.

1.2 Varactor Technologies

The word varactor comes from variable reactor. It is also called a varactor diode or varicap (variable capacitance) diode, in semiconductor devices that are widely used in the electronics industry and are used in many applications where a voltage controlled variable capacitance is required [9]. Also it is, in principle, any two-terminal device whose capacitance varies with the dc bias.
Varactors can be used to tune key characteristics of a circuit: such as center frequency, impedance and channel bandwidth. These modifications allow the same circuit to be used with a number of different operational frequencies.

Research groups all over the world reported several different tunable components for various radio frequency (R.F.) applications using different materials and technologies. In this chapter, the researcher will focus on all of these technologies, including their advantages and disadvantages.

1.2.1 Semiconductor Devices

Semiconductors are materials that have electrical properties somewhere in the middle, between those of a "conductor" and an "insulator", such as silicon (Si), germanium (Ge) and gallium arsenide (GaAs). The atoms of these materials are closely grouped together in a crystalline pattern called a "crystal lattice”. However, their ability to conduct electricity can be changed dramatically by adding small numbers of a different element to the semiconductor crystal called “impurities”, producing more free electrons than holes or vice versa [9].

The most used electronic element in semiconductor material is silicon. It has four valence electrons in its outermost shell which it shares with its neighboring silicon atoms to form full orbitals of eight electrons. The structure of the bond between the two silicon atoms is such that each atom shares one electron with its neighbor, making the bond very stable.
Fig. 1 shows that the minority carrier concentration at the edge of the depletion layer will change with the application of voltage bias. If a forward voltage ($V_A=$ positive) is applied, the barrier will be lower and the carrier injection (diffusion part) will increase. The minority carrier concentration at the edge of the depletion layer will also increase. Likewise if a reverse voltage ($V_A =$ negative) is applied, the barrier for carrier injection (diffusion part) will increase, and the minority carrier concentration at the edge of the depletion layer will decrease.

Semiconductor varactors include different kinds of diodes, p-n junction diodes and p-i-n diodes. A p-n junction diode can be used as a varactor component in a reverse biased condition. Equation (1.1) shows the capacitance in reverse bias of a p-n junction diode [10].

![Fig. 1. Simple View of p-n Junction Diode](image-url)
Where \( C_{j0} \) is the junction capacitance at zero bias, \( V_R \) is the reverse biased voltage, \( \phi \) is built in potential of the p-n junction; (which is given in equation (1.2) and \( \gamma \) is the grading coefficient.

\[
\phi = \frac{kT}{q} \ln \frac{N_A N_D}{n_i^2}
\]  

(1.2)

where \( kT \) is Boltzmann’s constant, \( n_i \) is the intrinsic impurity density of the p or n material, and \( N_A, N_D \) are the doping of the acceptor and donor atoms.

Fig. 2(a) shows the current versus bias voltage curve (I-V) of the p-n junction diode, in both the direction of forward bias and reverse bias, and the varactor diode symbol shown in Fig. 2(b).

The capacitance versus bias curve (C-V) of the p-n junction varactor diode is shown in Fig. 3. Varactor diodes are always operated under reverse bias conditions, and in this way there is no satisfactory conduction. One of the main reasons for the early success of the tunnel diode was its high speed of operation and the high frequencies it could handle. This resulted from the fact that, while many other devices are slowed down by the presence of minority carriers, the tunnel diode only uses majority carriers, therefore tuning can be used for the high frequency operation. The tunnel diode has many disadvantages which is why it is rarely used these days.
Fig. 2 (a) Current Versus Bias Voltage Curve of p-n Junction Diode,

(b) Varactor Diode Symbol

Fig. 3 C-V Characteristics of a p-n Junction Varactor Diode
To begin with, they only have a low tunneling current and therefore are low power devices. While this may be acceptable for low noise and low power amplifiers, it is a significant drawback when they are used in oscillators, since further amplification is needed and this can only be undertaken by devices that have a higher power capability. Another disadvantage is that they have problems with the reproducibility of their characteristics, resulting in low yields and therefore higher production costs.

Semiconductor varactor diodes are commonly used throughout RF, microwave and millimeter-wave spectrums for a variety of applications, including the tuning of critical circuits such as Voltage-Controlled Oscillators (VCO), filters, matching networks and phase-shifters in RFIC and MMIC circuits. However, semiconductor varactor diodes are limited in their applications because of performance issues [6].

The addition of off-chip discrete components increases the size, complexity and cost of the circuits, so alternative solutions to this problem have been sought after. One technology being investigated as a potential solution to this problem is RF-microelectromechanical system (RFMEMS) variable capacitors (varactors) [6].

1.2.2 MEMS Devices

Micro-Electro-Mechanical Systems Devices (MEMS) are used widely in a variety of microwave devices, such as switches, capacitors, inductors etc. MEMS are small integrated devices, which combine electrical and mechanical components. This technology is being pursued by the industry and academia for applications in RF circuits [6].
MEMS fabrication uses many of the same techniques that are applied in the integrated circuit domain such as oxidation, diffusion, ion implantation, LPCVD, sputtering, etc., and combines these capabilities with highly specialized micromachining processes.

MEMS devices are currently utilized in a wide range of electronic products, including automotive electronics, hard disk drives, wireless devices, medical equipment, and other portable electronic devices, (PDAs)[10]. The mobile phone is required to operate over many frequency bands, (GPS, LAN, WiMAX, FM, Bluetooth, and more) [5].

1.2.2.1 MEMS Technologies

MEMS is a semiconductor-based technology that uses the selective deposition and etching of a series of thin films to create a range of micron-scale mechanical structures for use in applications ranging from airbag deployment to telecommunications. MEMS varactors operate by using electrostatic forces to vary the separation between two plates and hence change the capacitance of the mechanical structure [5].

In a MEMS varactor, the distance between capacitor plates is varied with a control voltage, changing the capacitance. Because air or inert gas is the dielectric, the capacitors can have a very high Q factor [4].

A simple mechanical model of a varactor is described in Fig. 4, where the capacitor is formed by a fixed plate and a movable plate. The fixed plate can be created
using the conducting Poly0 layer, or by a stacked configuration of the Poly0 and Poly1 layers. The movable plate can only be formed by etching away one or more of the oxide layers to allow vertical movement. The movable plate can be formed by depositing gold on top of the Poly2 layer, or it can be formed from a stacked configuration of the gold, Poly2 and Poly1 layers. Essentially, there are three layer configurations possible for building varactors in the PolyMUMPs process, which are shown in Fig. 5.

The three varactor configurations are mechanically tuned by changing the distance separating the plates. Therefore, the spacing between the plates must be mechanically controlled. The spacing between these plates can be adjusted in the PolyMUMPs process using electrostatic actuation as shown in Fig. 5, or using thermal actuation. The use of electrostatic actuation dominates due to its inherent speed and the low actuation power required. Thermal actuators rely upon resistive heating to deform a structure as a function of its thermal coefficient of expansion.

Thermal actuation is much slower than electrostatic actuation, and the power dissipated by resistive heating is much higher than that required for electrostatic actuation. Therefore, electrostatic was the only actuation concept considered.

Actuation occurs when an electrostatic force is created by applying a DC voltage between the capacitor electrodes. Increasing the DC voltage increases the electrostatic force between the electrodes, thereby displacing the movable plate toward the fixed plate.

The basic principles for the operation and design of MEMS varactors, are that when a control voltage ($V_{ctl}$) is applied, the suspended plate moves toward the fixed plate
until equilibrium is reached. Equilibrium occurs when the spring force (kx) equals the electrostatic force of the actuator.

\[
kx = 0.5 \varepsilon_{\text{air}} A_{\text{eff}} \left[ \frac{V_{\text{ctl}}}{d_0 \pm x} \right]^2
\]  

(1.3)

Where \( \varepsilon_{\text{air}} \) = the dielectric constant of air, \( A_{\text{eff}} \) = the effective area of the capacitor plates and \( d_0 \) = the distance between the plates when \( V_{\text{ctl}} = 0 \) [6].

Fig. 4. Mechanical Model of MEMS Varactor [4]
Fig. 5. Three Layer Configurations Possible for Building Varactors in The PolyMUMPs

1.2.2.2 Advantages and Disadvantages of MEMS Varactors

MEMS varactors are one of the most important passive MEMS devices. They have considerable advantages compared with other semiconductor devices, including low loss, very high Q at mm-wave frequencies, high power handling capability, low power consumption, and high IIP3[3]. They have a low tuning voltage, high stable tuning range, high quality factor, high base capacitance, and low harmonic distortion, as the device inertia prevents them from responding to RF frequencies. They also have low sensitivity to residual stress and temperature, inherently switched operation which allows for operation in arrays, low-temperature assembly, low profile, low-waste package, and
may be simultaneously fabricated with other IC-compatible passive or active components [1-2]

On the other hand, the switching speed of most electrostatic MEMS switches is 2-40µs, and thermal/magnetic switches are 200-3000µs. Certain communication and Radar systems require much faster switches. Most switches cannot handle more than 200mW although some switches have shown up to 500mW power handling (Terravicta and Raytheon). MEMS devices that can handle 1-10W with high reliability don't exist today. The reliability of a mature MEMS switch is 0.1-40 billion cycles. However, many systems require a reliability of 20-200 billion cycles. Also, the long term reliability has not yet been addressed [3].

It is now well known that the capacitive switches are limited by dielectric charging that occurs in the actuation electrode, while the metal contact switches are limited by the interface problems between the contact metals, which could be severe under low contact forces (in electrostatic designs, the contact forces are around 40-100 µN per contact). MEMS switches need to be packaged in inert atmospheres (Nitrogen, Argon etc.) and in very low humidity, resulting in hermetic or near hermetic seals. Hermetic packaging costs are currently relatively high, and the packaging technique itself may affect their liability of MEMS switch [8].

1.2.3 Ferroelectric Devices

Significant progress has been made in the development of ferroelectric composites that achieve higher dielectric constants and lower loss tangents at room
temperature. Ferroelectric and paraelectric materials have a tendency of changing its permittivity with the external applied electric field. Thus, they are a natural candidate for tunable devices. There is an increasing interest to design tunable microwave components with tunable ferroelectric capacitors such as Barium Strontium Titanate (BST), based on this technology exhibit high tuning speeds, good tunability, and high power handling capability and lower power requirements than ferrite phase shifters. Both bulk and thin film ferroelectrics have been studied so far by the researchers.

Ferroelectric materials and devices have been paid much attention recently because of their unique voltage tunable capability. There is an increasing demand for ferroelectric thin film capacitors and devices because of its ability to integrate advanced IC’s and RF circuits. High dielectric constant thin film ferroelectric capacitors could reduce the device area at least by a factor of 10 than of the conventional dielectric capacitors. A voltage controlled capacitor gives the circuit designer a flexible element for the design of filters, oscillators, and tunable phase shifters. When coupled with high temperature superconducting wires and inductors, the designer has the opportunity to build low loss circuits. Table 1 shows the comparison among the various technologies [10].

1.3 Ferroelectric Capacitors

Researchers are increasing interest to design tunable microwave components with tunable ferroelectric capacitors such as Barium Strontium Titanate (BST). BST capacitors
Table (1) The Comparison Among The Various Technologies

<table>
<thead>
<tr>
<th></th>
<th>Semiconductor</th>
<th>MEMS</th>
<th>Ferroelectric</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Tunability</strong></td>
<td>High</td>
<td>Low</td>
<td>Moderate to High</td>
</tr>
<tr>
<td><strong>RF Loss</strong></td>
<td>moderate</td>
<td>Very good</td>
<td>moderate</td>
</tr>
<tr>
<td><strong>Control Voltage</strong></td>
<td>&lt;20 V</td>
<td>50-100 V</td>
<td>10-20 V</td>
</tr>
<tr>
<td></td>
<td>Unipolar</td>
<td>Bipolar</td>
<td>Bipolar</td>
</tr>
<tr>
<td><strong>Tuning Speed</strong></td>
<td>Fast</td>
<td>slow</td>
<td>Fast</td>
</tr>
<tr>
<td><strong>Power Handling</strong></td>
<td>Poor</td>
<td>Excellent</td>
<td>Excellent</td>
</tr>
<tr>
<td><strong>Technology dependence</strong></td>
<td>Dependent</td>
<td>Independent</td>
<td>Independent</td>
</tr>
<tr>
<td><strong>Chip assembly</strong></td>
<td>Easy</td>
<td>Very Hard</td>
<td>Easy</td>
</tr>
<tr>
<td><strong>Passivation</strong></td>
<td>Possible</td>
<td>Not Possible</td>
<td>Possible</td>
</tr>
<tr>
<td><strong>Stability</strong></td>
<td>Very good</td>
<td>Low</td>
<td>Very good</td>
</tr>
<tr>
<td><strong>Cost</strong></td>
<td>Moderate to high</td>
<td>Low</td>
<td>Low</td>
</tr>
</tbody>
</table>

are fabricated on 4" diameter substrates. The wafers are degreased by ultrasonic cleaning in acetone and methanol.

One of the methods to tune this varactor capacitor is by applying an electric field to change the dielectric constant ε. There are two phases of tunable materials, ferroelectric and paraelectric. In the ferroelectric phase, they exhibit hysteresis in polarization versus to electric filed curve as shown in Fig. 6(a). Fig. 6(b) shows
polarization versus electric filed curve for paraelectric phase. The advantages of using ferroelectric capacitor are:

1. High dielectric constant and as a result the self-capacitance is much higher than parasitic capacitance.
2. Low dissipation factor at high frequencies.
3. Good tunability over a wide voltage range.
4. Low temperature dependence of capacitance.
5. Less noisy than varactor diodes. Does not produce junction noise comparing to p-n junction.
6. Compatible with current CMOS technology, monolithic GaAs technology, low cost, low voltage tunability and high speed of operation.

In recent years, another ferroelectric material $\text{Ba}_{1-x}\text{Ca}_x\text{Ti}_{1-y}\text{Zr}_y\text{O}_3$ (BCTZ), which is similar to BST, has been studied for tunable applications at RF and microwave frequencies [10].

BCTZ has been widely studied and applied as a thick film formed by sintering for multilayer chip capacitors. BCTZ exhibits benefits similar to BST: high dielectric constant and low leakage current. In contrast, BCTZ is routinely processed in reducing atmospheres such as forming gas, while the dielectric properties of BST are known to degrade sharply with exposure to reducing anneals. Despite these facts, very little research has been performed on thin film BCTZ.
Fig. 6 The Difference Between Two Methods (PZT and BST)
Only two groups have studied thin film BCTZ, and both have used sol-gel processing for fabrication of their films. In our research facility at the University of Colorado at Colorado Springs, our research group has been studying thin film BCTZ using the rf sputtering system. Some other advantages of BCTZ over BST are that it is a relax or type ferroelectric material and exhibits broad dielectric constant-temperature curve near curie temperature $T_c$.

BST has a good tunability at low bias voltage, with moderate temperature annealing. It has some limitations due to its poor thermal conductivity and susceptibility to thermal stress.
CHAPTER 2

LITERATURE REVIEW ON TUNABLE CIRCUITS

2.1 Introduction

In this chapter, we will review some applications of tunable components discussed by other researchers. The research in this area has grown substantially in recent years for various frequencies. In this chapter, the study will discuss what those researchers are doing to develop these tunable elements and where they are used in different applications such as: band-pass filters and band-stop filters in single bands and multi bands.

2.2 Single Band

2.2.1 Band-Pass Filter

This filter allows the special band to pass only in the certain frequency, and to reject all other signals out of this band. Fig. 7 shows a lumped circuit of band-pass filter.

Fig. 7 Lumped Circuits of BandPass Filter
The important BPFs we can use and tune is microstrip tapped-line filters; the two most common types are the interdigital filter and the combline filter, as shown in Fig.8. They have more flexibility when they are built on microstrip substrate, and are of a very low cost. They also have a small size and they offer space advantages [11]. In this study, the researcher will focus on how an interdigital band-pass Filter can be designed.

![](image)

(a) Interdigital Filter

(b) Combline Filter

Fig. 8 Tapped-Line Filters [11]

2.2.1.1 Design Tapped-Line Bandpass Filter

Microstrip, interdigital and combline tunable filters have been described by many authors. A simple method to design a tunable tapped-line is to use the method shown below for the design of an interdigital filter [11]. Fig. 8(a) shows the schematics of an interdigital filter and Fig. 8(b) shows a combline filter.
a) The first step is to determine the coupling coefficients $K$ of the pair of resonators by following equation [7]:

$$K = \frac{f_2 - f_1}{f_0} = \frac{\Delta f}{f_0}$$  \hspace{1cm} (2.1)

where $f_0$ is the center frequency $= \frac{1}{2} (f_1 + f_2)$. $f_1, f_2$ are the frequencies at 3 dB before and after center frequency.

b) Measure the loaded $Q_L$ (Quality Factor) of the first and the last resonators in order to obtain tap-point locations.

$$Q_L = \frac{f_0}{BW_{3dB}}$$  \hspace{1cm} (2.2)

where $BW_{3dB} = BW/0.955 = (f_2 - f_1) / 0.955$

c) From Fig. 9 can be found the length of resonator edge to connecting point ($l$) as shown in Fig. 8; this figure shows the relation between $Q_L$ and $l/L$ ($L$ is length of resonator). Thus, can be found the $l/L$ from Fig. 9 or from the following equation:

$$\frac{Q_l}{Z_0/Z_{0l}} = \frac{\pi}{4\sin^2\left(\frac{\pi l}{2L}\right)}$$  \hspace{1cm} (2.3)

where $Z_0 = 50 \Omega$, $Z_{0l} = 70 \Omega$, $L = \frac{\lambda}{4}$, $\lambda = \frac{c}{f}$.

d) The final step is to find the dimensions of this microstrip line by using any software like ADS or Ansoft.
2.2.1.2 Design Tunable Bandpass Filter

Different tunable devices have been used to implement tunable BPFs. There are many strategies and technologies to achieve this goal, such as: semiconductor varactor diodes, micro electromechanical (MEMS), ferroelectric components and liquid crystals, among others [13].

Some researchers used varactor diodes to tune their BPFs circuit. These elements are used to load to coupled resonators. Varactor diodes (Infineon BB833) with a capacitance varying in range from 9.3-0.75 pF, to apply voltage among 1 and 28V, and they used tunable Open Complementary Split Ring Resonators (OCSRRs). They found

![Fig. 9 Singly Loaded $Q_L$ for a Tapped Interdigital Resonator [7]](image)

Fig. 9 Singly Loaded $Q_L$ for a Tapped Interdigital Resonator [7]
the tuning range of the band pass filter is over 30% for applied bias voltage in the range 5-25V [13].

Nath, et. al., are tuning filters by using BST thin film varactors in various topologies with two choices for each topology; a lumped element circuit or a distributed implementation [14].

Nath, et. al., applied this topology to the combline band-pass filter with a resonator length of 200 µm and width of 5 µm. The spacing among the fingers was 5µm and there were 3 resonators as depicted in Fig. 10. Each resonator is loaded with BST varactors. They use alumina (Al$_2$O$_3$) as the substrate for fabrication of the BST varactors and tunable filters [14].

Tunable BPF, fabricated by this approach, are tuned by a DC voltage from 0V to 200V. The center frequency filter changed by 25%, as shown in Fig. 11.
The insertion loss was 6.6 dB at zero bias and this decreased to 4.3 dB at 200V bias, and the return loss increased from 10 dB at zero bias to 14.7 dB at 200V bias, as depicted in Fig. 12[14].

![Fig. 11 Measured Insertion Loss of The Filter with Applied Bias](image1)

![Fig. 12 Measured Return Loss of The Filter with Applied Bias [14]](image2)

Varactor diodes or MESFET varactors were also used to tune the pass center frequency by Lopez, and Alonso [15]. The combline filter depicted in Fig. 13, has the
Dimensions: width 0.40 mm, length 1.59 mm, spacings (S1,S2) (0.331,0.439) mm and length from ground to input line (lt) 0.37 mm.

Fig. 14 shows comparison among simulation and measurement values when voltage from zero to 5V is applied; the center frequency equals 8 GHz at zero bias with the return loss of about 15 dB; however, it is around 12.5 GHz with 25 dB return loss at 5V [15].

MEMS is another technology used to implement tunable BPFs by Saha et al.[16]. The band-pass filter design uses both distributed transmission lines and RF MEMS capacities together to replace the lumped elements. The use of RF MEMS variable capacitors gives the flexibility of tuning both the center frequency and the band-width of the band-pass filter [16]. In this work, MEMS varactors work as a lumped element with off-chip inductors. The cut-off and center frequencies were tuned by changing the
inductance using MEMS switching. Their design of a band-pass filter with the combination of a low pass filter and a high pass filter is shown in Fig. 15.

Fig. 14 Return Losses for Various Control Voltage: Simulations (Dashed line) and Measurements [15]

Fig. 15 The Basic Band-Pass Filter, Which is Obtained From Low and High Pass Combination [16]
Fig. 16 shows the schematic view of the band pass filter with a transmission line. The MEMS capacitor and the filter were characterized up to 30 GHz. The actuation voltage was around 15 volts for shunt bridge and 8 volts for cantilevers. The S-parameters of this filter are shown in Fig. 17.

The smaller value of series capacitance may increase the return loss at higher center frequency. Capacitive tuning with an inductive inverter can be used to reduce the return loss for the entire tuning range. Saha, with other authors, conclude that the filter has the flexibility of tuning both center frequency and bandwidth simultaneously, and it is very compact and has very good control of the tuning frequency [16].

Table 2 shows a comparison among different kinds of methods used to tune BPFs. This comparison is shown in this table explaining the difference between these tunable elements in voltage, tuning range, frequency, BW, frequencies responses ($S_{11}$, $S_{21}$) type of tuning and the year of this work.
Table 2 Comparison of The Different Methods to Tune BPF

<table>
<thead>
<tr>
<th>Method</th>
<th>Varactor Diodes</th>
<th>BST Varactors</th>
<th>Varactor diode with capacitance</th>
<th>MEMS Varactor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>0 - 5V</td>
<td>0-200 V</td>
<td>5 - 25 V</td>
<td>8 - 15V</td>
</tr>
<tr>
<td>Tuning range</td>
<td>-----</td>
<td>25%</td>
<td>&gt; 30%</td>
<td>30%</td>
</tr>
<tr>
<td>Frequency</td>
<td>8 - 12.5GHz</td>
<td>1.6 - 2GHz</td>
<td>2.4 - 3.6GHz</td>
<td>8 - 10GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>630 -1350MHz</td>
<td>1.8 - 1.9 GHz</td>
<td>0.6GHz</td>
<td>913 - 1043MHz</td>
</tr>
<tr>
<td>Insertion Loss</td>
<td>4 - 13 dB</td>
<td>4.30 - 6.55 dB</td>
<td>2 - 3 dB</td>
<td>2 - 4 dB</td>
</tr>
<tr>
<td>Type of Tuning</td>
<td>f₀ , BW</td>
<td>f₀ , BW</td>
<td>f₀</td>
<td>f₀ , BW</td>
</tr>
</tbody>
</table>
2.2.2 Band Stop Filter

Band stop filters allow all the signals to pass except the specified band of frequencies. Fig. 18 shows the lumped circuit of the band stop filter.

In this section, the researcher will review how the researchers design tunable circuits and which different tunable components are used to tune BSFs to work in different frequencies.

Vanadium Dioxide (VO2) is an attractive material exhibiting an abrupt and reversible semi-conductor to metal (SC-M) transition at the critical temperature $T_c=341K$ [17]. That means this VO2 thin film works like a switch depending upon the temperature.

Following the VO2 thin film structural and electrical characterizations, Dumas-Bouchiat et. al [17], fabricated switching devices for rf-microwave frequency domain in two types (a) Shunt Configuration, (b) Series Configuration as shown in Fig. 19. They found the average contrast, respectively, of -24 dB (shunt) and -25 dB (series), and the insertion losses of about 3 dB in both configurations [17].

![Fig. 18 Lumped Circuits of Band Stop Filter](fCELab.com)

This bandstop filter designed and simulated to operate in the 11-13 GHz uses VO2 switch [18]. Givernaud et al [18], used series switch from VO2 in the design of
bandstop filter. The filter showed a good results with state of room temperature and off state at 80° C. $S_{21}$ parameter of this filter is shown in Fig. 20.

Fig. 19 Microwave Switching Device Design Dased on VO2 Thin Films:
(a) Shunt Configuration and (b) Series Configuration [17]

Fig. 20 Electromagnetic Simulation (curves 1 and 2) and Measurement (curves 3 and 4) of The $S_{21}$ Parameter [18]
The tuning diode is most common tuning elements used in the design of band stop filter. Auffray, and Lacombe used a five-section coupled resonator as shown in Fig. 21 for the design of band stop filters [19]. This band stop filter can be tuned from 6.5GHz to 10 GHz, with a constant bandwidth. The tuning diodes are connected to the resonators, and tuning diode capacitance can be changed by DC voltage.

Fig. 21 Five Sections of-Coupled Resonators Band-Stop Filters [19]

Fig. 22 $S_{21}$ Responses for Different Values of Capacitances [19]
The advantages of this design are: very good performance, small dimensions of hybrid microstrip technology and very low cost. The filter can be tuned from 6.5 GHz to 10 GHz with a constant bandwidth of 250 MHz at 25 dB and with attenuation greater than 40 dB as shown in Fig. 22, and this filter can be used over the 2 - 28 GHz with insertion losses smaller than 1 dB [19].

Lumped element approach was also used in the design of band stop filter by Yu-Chin Ou, and Gabriel [20]. The lumped element filters use variable capacitors to achieve both frequency and bandwidth control. Fig. 23 shows the equivalent lumped element circuit used in this filter, and they used the following formulas to calculate the values of lumped elements of BSF.

\[ C_s = \sqrt{\frac{\Delta \theta_1}{L_p Z_0 \omega_0^3}} \]  
\[ C_p = \frac{1}{\omega_0^2 L_p} - C_s \]
The transmission line equivalent was used in the lumped element, as shown in Fig. 24. The filter was tuned by two methods: with single varactor diodes and with back to back varactor diodes. The measured 16 dB rejection S-parameters for tunable filters are shown in Fig. 25. A single diode filter tunes the center frequency of the filter from 470 to 720 MHz and the back to back diode filter tunes from 510 to 750 MHz. Both have a 0.3 dB to 0.5 dB insertion loss [20].

Vélez et. al. [13] designed tunable band stop filters based on Open Split Ring Resonators (OSRRs). The tuning range is roughly 120% for a varying applied voltage between 0 to 28V. The measured rejection level is around 30dB. The results of the co-simulations for different applied voltage and simulation are depicted in Fig. 26.

\[
L_L = \frac{Z_0}{\omega_0 \sqrt{g_0 g_3}} \quad (2.6)
\]

\[
C_L = \frac{\sqrt{g_0 g_3}}{Z_0 \omega_0} - C_s \quad (2.7)
\]

A.C. Guyette [21] presented a new tunable band stop resonator architecture which allows tuning of center frequency and bandwidth. Bandwidth tuning is achieved by attaching two varactors at opposite ends of the resonator. The center frequency tuning is achieved by tuning both varactors simultaneously. Fig. 27(a) and Fig. 27(b) show tunable circuits of frequency and bandwidth for band stop resonator filter. Figs. 27(c,d) show
Fig. 25 Measured S-Parameters of The Single Diode and The Back to Back Diode Tunable BSFs [20]

Fig. 26 Comparison Between CO-Simulation, Pure Electric Simulation and Measured Frequency Responses [13]
S-parameter of center frequency and bandwidth tuning respectively. Guyette designed a 2nd order notch filter with both a center frequency and a bandwidth tuning microstrip circuit. The measured results shown in Fig. 28 were tuned from 1.2 to 1.6 GHz of center frequency and the bandwidth is tuned from 70 to 140 MHz [21].

Ferroelectric materials such as BST have recently become more attractive for the development of electronically tunable microwave circuits. Young-Hoon, et. al. [22] presented a newly developed BST-varactor tunable bandstop filter (BSF). They fabricated and modeled an interdigital filter using a BST varactor as a tuning device. In this work, BST varactors are fabricated based on a structure of Fig. 29. Fig. 30 shows the fabricated tunable BSF with BST varactors, and the measured results of this filter are shown in Fig. 31, with center frequency from 1.2 to 1.4 GHz and bandwidth 100 MHz. Table 3 shows the comparison among different methods used to tune BSFs.

Fig. 27 Tunable Circuits (a,b) Band Stop Resonator Filter for Tuning Frequency and Bandwidth, (c,d) S-Parameter of Center Frequency and Bandwidth Tuning [21]
Fig. 28 Tunable 2nd Order Notch Filter with Both Center Frequency and Bandwidth Tuning [13]

Fig. 29 Profile of Layers for a Fabrication of BST Interdigital Varactors [22]
Fig. 30 The Fabricated Tunable BSF with BST Varactors [22]

Fig. 31 The Measured Results of Tunable Band Stop Filter [22]
### Table 3 Comparison of The Different Methods to Tune BSF

<table>
<thead>
<tr>
<th>Method</th>
<th>BST Capacitance</th>
<th>Capacitor Varactors</th>
<th>Varactor diode with capacitance</th>
<th>Lumped Element (Signal Diode)</th>
<th>Lumped Element (B. to B. Diode)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage</td>
<td>Change Capacitors 0.24-1.5 pF</td>
<td>0 - 35 V</td>
<td>Use different kinds of CV</td>
<td>0 - 28 V</td>
<td>4 - 30 V</td>
</tr>
<tr>
<td>Tuning range</td>
<td>14%</td>
<td>120 %</td>
<td></td>
<td>-----------------------------</td>
<td>-------------------------------</td>
</tr>
<tr>
<td>Frequency</td>
<td>6.5 - 10 GHz</td>
<td>1.2 - 1.4 GHz</td>
<td>1.2 - 1.6 GHz</td>
<td>0.4 - 2.4 GHz</td>
<td>470 - 720 MHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>250 MHz</td>
<td>100 MHz</td>
<td>70 - 140 MHz</td>
<td>0.9 GHz</td>
<td>8.5 - 12 MHz</td>
</tr>
<tr>
<td>Return Loss</td>
<td>&lt; 1 dB</td>
<td>2 - 2.5 dB</td>
<td>------</td>
<td>1 - 3 dB</td>
<td>0.3 - 0.5 dB</td>
</tr>
<tr>
<td>Type of Tuning</td>
<td>$f_0$</td>
<td>$f_0$</td>
<td>$f_0$, BW</td>
<td>$f_0$, BW</td>
<td>$f_0$, BW</td>
</tr>
</tbody>
</table>

#### 2.3 Dual Band Filters

Design of dual-band filters at microwave frequencies is still challenging since it has to take into consideration many parameters, including center frequency, bandwidth, and/or passband and bandstop functions. There have been many innovative dual-band
bandpass and bandstop filter designs [23]. Thus, in this section of this chapter the study will try to exhibit what other researchers have done in this area.

2.3.1 Band Pass Filter

Band Pass Filters are one of the key components in the RF front end of a microwave communication system. There are many methods to design dual-band pass filters. One way of designing a dual-band filter is to combine two bandpass filters designed for two different passbands. Another popular approach is to design a compact dual-band filter by utilizing the harmonic tunable property of a stepped-impedance resonator [24]. Also Sun et al. [25], using a simple microstrip ring resonator is presented for a novel design of dual band mode band pass filters with good isolation. This method is used widely because a ring resonator has the advantages of compact size and high quality factor (Q).

Fig. 32(a) shows the photograph of the fabricated circuit. The frequency responses of the measured minimum insertion losses achieved 0.65 dB in the first passband and 1.0 dB in the second passband, and 32 dB return loss from 2.88 to 3.34 GHz, with passband at 2.3 and 4.1 GHz as shown in Fig. 32(b).

Coupled line and coupled three line resonators are proposed to design dual wideband bandpass filters by Kuo, Fan, and Tang [23]. These resonators provide larger possible bandwidths and different frequency ratio of the two center passbands. Fig. 33(a) shows the layout of a coupled line resonators circuit, and the fabricated circuit is shown in Fig. 33(b). This circuit is designed to work in 900 MHz and 1575 MHz with
Fig. 32(a) The Photograph of The Fabricated Circuit, (b) Frequency Responses [25]

Fig. 33 (a) Layout of Coupled Line Resonators Circuit, (b) Fabrication Circuit
bandwidth $\Delta = 39\%$. Fig. 34 compares the measured with the simulated responses. The insertion losses are $-0.987$ dB and $-1.318$ dB, and the return loss is around $28.9$ dB.

For the second design, Fig. 35(a) shows the layout of a coupled three line resonators circuit. The fabricated circuit is shown in Fig. 35(b). It is designed to work at 900 and 1800 MHz with bandwidth $\Delta = 38\%$. The measured insertion losses are $-1.38$ dB and $-1.49$ dB, respectively, with a return loss of about $25$ dB as shown in Fig 36. In this work, the measured results match with the simulated responses quite well as shown in Figs. 34 and 36 [23].

All the previous researchers used different methods of designing the dual-bandpass filter without tuning these circuits. Abunjaileh, and Hunter [26], they tune bandpass filters based on dual band combline structures. A new approach can be used to realize tunable bandstop filters with constant bandwidth, and tunable bandpass
filters with tunable bandwidth and center frequency. A dual bandpass filter with wide passband and stopband is used to design a tunable bandpass filter. The measured results of the fabricated circuit are shown in Fig. 37, and in Fig. 38. The tuning mechanism was tuning screws located over the open circuited ends of the resonators.

Fig. 35 (a) Layout of Coupled Three Line Resonators Circuit, (b) Fabrication Circuit [23]

Fig. 36 Comparison of The Measured with The Simulation Responses [23]
Fig. 37 Fabricated Circuit of Tunable Dual Band Band Pass Filter [26]

Fig. 38 Frequency Responses of Tunable Band Pass Filter [26]
Dinghong Jia, et. al. [42], a novel dual-band bandpass filter with a tunable passband was proposed. The proposed filter structure offers a possibility of two tunable passbands with a fixed first passband and controllable second passband. Each passband can be independently controlled and has a tiny influence on the other. They divided the filter structure into independent filters with the original parameters, as shown in Fig. 39, which loaded with varactor capacitor. The single capacitances are 0.63 pF and 2.67 pF at 30 V and 0 V bias, respectively.

Fig. 40 shows the simulated and measured results, the first passband shows a center frequency of 1.472 to 1.886 GHz with fractional bandwidth range of $3.9 \pm 0.3\%$, an insertion loss of 4.2 to 5.7 dB, when they applied voltage from 30 to 3 V. At the first passband was tuning, the second passband exhibits a small variation with center frequency of 2.24 GHz with bandwidth of 53 MHz, an insertion loss of 3.517 dB.

Fig. 39 (a) Layout of Filter I, (b) Layout of Filter II [42]
2.3.2 Band Stop Filter

The dual band bandstop filters have become quite popular in recent years because we needed them for wireless mobile communication and various of microwave circuits. Dual-Band BandStop Filters (DBBSFs) are commonly employed in high power amplifiers and mixers to suppress the double sideband spectrum to reduce circuit size and cost. The dual section or tri-section stepped impedance resonators as shown in Fig. 41 can resonate at two different frequencies and be used in dual band bandstop filter design [27 - 29]. They designed three second order maximally flat dual band bandstop filters with various frequency ratios.

The filters were designed at 1.5 and 3.15 GHz center frequency with bandwidths of $\Delta f = 70\%$ and $\Delta s = 35\%$. Fig. 42(a) is a photograph of the fabricated filter. The frequencies measured were 1.495 and 3.11 GHz with bandwidths of 55% and 27%, respectively as shown in Fig. 42(b). These results agree closely. The second filters were
designed at 2.4 and 5.8 GHz with bandwidths of 50% and 25%, the measured results are close to the simulated results as shown in Fig. 43. The third filter was designed at 1.5/4.275 GHz with bandwidths of 50% and 25%, and the measured results were very close as shown in Fig. 44 [27].

Ajay and Subrata [29], designed dual band bandstop filters at GSM bands to stop the interference due to the mobile bands at the operating frequency of 1.4 GHz. Stepped impedance resonators are used to achieve the excellent out-of-band and in-band

![Diagram](image1)

**Fig. 41** Configurations of (a) Two Section SIR (b) Tri- Section SIR [27]

![Graph](image2)

**Fig. 42** (a) A Photograph of The Fabricated Filter, (b) The Frequencies Responses [27]
Fig. 43 The Frequency Responses and Fabricated Circuit of Second filter [27]

Fig. 44 The Frequency Responses and Fabricated Circuit of Third Filter [27]

Fig. 45 The Microstrip Layout of The Filter [29]
performance respectively. The proposed filter showing the high rejection in the two stop bands up to 35dB. The microstrip layout of the filter is given in the Fig. 45.

Another study by Abu Safia, et. al, used split ring resonators to design DBBSFs at a center frequency of 9.8 and 11.4 GHz [30]. CPW technology is one of the other methods used to design DBBSFs by Kalkur, et. al [32]. They use series connected parallel LC resonators patterned in the center conductor of CPW. The measured results are in agreement with the simulated results.

The previous studies have different methods of designing dual-band bandstop filters (DBBSFs) without any tuning for these filters. Abunjaileh, and Hunter [26], tuned bandstop filters with constant bandwidth by using the following equation to transfer the network to lumped dual band structure with combline structures.

\[ \omega = \alpha \left[ \frac{\omega}{\omega_0 A} - \frac{\omega_0 A}{\omega} \right] - \beta \left[ \frac{\omega}{\omega_0 B} - \frac{\omega_0 B}{\omega} \right] \]  (2.8)

They used the flexibility of the dual-band approach to realize tunable bandstop filters with constant bandwidth, and tunable bandpass filters with tunable bandwidth and center frequency. They designed center frequency tuning from 3.1 to 3.8 GHz with 30 MHz bandwidth. Fig. 46 shows the fabricated circuit. The tuning mechanism was tuning screws located over the open circuited ends of the resonators. The measured results for 3 different center frequencies were 3.225, 3.465 and 3.66 GHz shown in Fig.47 [26].

Jun Wang et. al. [43] designed a novel miniaturized dual-band bandstop filter by using the T-shaped Defected Microstrip Structures (DMSs) and the U-shaped Defected Ground Structures (DGSs). Fig. 48 shows the configuration of the dual-band
bandstop filter resonator for both shaped T and U shaped. Simulated and measured S-
parameters of the fabricated DBBSR as shown in Fig. 49 without any tuning.

Fig. 46 Fabricated Circuit of Tunable DBBSF

Fig. 47 Frequency Responses of DBBSFs
Jun Wang et. al. changed their design to get tuning for the first order of this kind of filter as shown in Fig. 50 for T-shaped and Fig. 51 for U-shaped. They designed a second order of this filter to work at 3.66GHz and 5.07GHz, but without any tuning for them as shown in Fig. 52 with FBW₁=7.91% and FBW₂=9.17%.

Fig. 48 Configuration of The Dual-Band Bandstop Filter Resonator

Fig. 49 Simulated and Measured S-parameters of The Fabricated DBBSR
Fig. 50 Variations of The Transmission Response for Different T-Shaped DMS [43]
Fig. 51 Variations of the Transmission Response for Different T-Shaped DMS [43]

Fig. 52 Full-Wave Simulated Responses DBBSF, T-Shaped DMSs and U-Shaped DGSs
CHAPTER 3

DESIGN, SIMULATION AND IMPLEMENTATION OF

TUNABLE BAND STOP FILTER

3.1 Introduction

Recently, adaptive electronics has attracted the attention of many investigators for front end RF communication applications to cover multiple frequency bands. The replacement of fixed frequency circuits with reconfigurable adaptive circuits reduces space and cost. Tunable devices such as p-n junction varactors, Micro Electro Mechanical Systems (MEMS), liquid crystal cells and ferroelectric varactors have been under investigation for the fabrication of tunable RF circuit blocks such as voltage controlled oscillators, matching networks for power amplifiers, tunable filters, active and passive phase shifters, and tunable antennas [32-35]. Tunable ferroelectric capacitors play an important role in tunable RF blocks because of their small size, ease of handling, voltage polarity independent tuning characteristics, low voltage operation and high quality factor. Tunable band stop filters (notch filters) are extremely important in eliminating unwanted frequencies in communication systems [36-38]. In this study, the results of a single pole and a multi pole band stop filters were presented by implementing integrated discrete chip inductors and discrete tunable ferroelectric capacitors with
coplanar microstrip with FR-4 substrates. The performances of these filters are compared with filters fabricated with tunable p-n junction varactors.

BandStop Filters (BSFs) or Notch Filters are useful to avoid unwanted signals. Tunability in the filters have been studied to enable their flexibility. In recent years, there was a demand in the wireless communication industry for smart and reconfigurable radio transceivers. Furthermore, tunable bandstop filters become more important to reject unwanted signals for wideband applications like ultra-wideband (UWB) systems [22].

In this chapter, two different components will be used to tune BSF, i.e., tuning diode and ferroelectric capacitor. These circuits will be designed by using microstrip with 50 Ω impedance with lumped elements for one order and three orders. Advanced Design System software (ADS) is going to be used to design these circuits, and will fabricate and compare measured results with a simulation result. The advantages of this design: small dimensions, easy to design, good performance, and very low cost.

### 3.2 Fabrication of Tunable BST Capacitors

BST capacitors are fabricated on 4” diameter substrates. The wafers are degreased by ultrasonic cleaning in acetone and methanol. The substrates are dehydrated by baking in an vacuum oven for 30 minutes. Platinum films of thickness 200 nm with an adhesion layer of titanium of thickness 20 nm are deposited on these sapphire wafers by DC magnetron sputtering in argon environment. These films were patterned by standard photolithographic techniques and ion milling to form bottom electrodes for the capacitors. The BST films (Ba0.5Sr0.5TiO3) are deposited by spin-on Metal Organic Decomposition (MOD) technique. The films are annealed in an oxygen environment at
800°C for 30 minutes in oxygen environment. Top electrode platinum film of thickness 200 nm is deposited by DC magnetron sputtering. The top electrodes were also patterned by photolithography and ion-milling. Fig. 53(a) shows the cross section and Fig. 53(b) shows the plan view of the tunable capacitor. The area of the parallel plate capacitor is determined by the overlap area of the bottom and top electrodes.

### 3.3 Electrical Characterization of BST Film

The capacitance versus voltage characteristics of BST capacitors were determined by HP4275A LCR meter. The capacitance versus voltage characteristics of the capacitor is measured with small signal AC voltage of frequency 1 MHz and amplitude 100 mV.

The capacitance versus voltage characteristics of the capacitor is shown in Fig. 54. The tunability of the BST capacitor is defined as:

$$Tunability = \frac{\varepsilon_{r_0} - \varepsilon_{r_b}}{\varepsilon_{r_0}} \times 100\%$$  \hspace{1cm} (3.1)

where $\varepsilon_{r_0}$ is the dielectric constant at zero bias, and $\varepsilon_{r_b}$ is the dielectric constant at the maximum bias [33]. Fig. 55 shows the variation of tunability with applied DC bias. With increase in applied voltage, the tunability increases almost linearly and a maximum tunability of about 50% that was obtained for an applied bias of 5 V [40].
Fig. 53 (a) Cross and Section, (b) Plan View of BST Capacitors on Sapphire

Fig. 54 Capacitance Versus Voltage Characteristics of a BST Capacitor
3.4 Principal Equation and Design of Band Stop Filter

Band Stop Filters reject frequency in certain band width. Also called band elimination, band reject or notch filters. There are different methods to design a notch filter, the simplest method is to use a standard two zero, two pole. Other methods are Double Notch, Chebyshev and Elliptical filters [36], and can be designed from low pass filter by using the frequency transformation.

\[
\frac{1}{\omega'} = \frac{\omega_0}{\omega_2 - \omega_1} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)
\]  

(3.2)

The low pass prototype circuit, band stop filter circuit and their attenuation responses are shown in Fig. 56, where \(\omega_x\) represents a frequency for high attenuation in the stop band [39].

Fig. 55 Variation of Tunability with Applied DC Bias Voltage
Equation (3.2) can be modified to the following equations:

\[
\omega \leftarrow \frac{1}{\Delta} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \tag{3.3}
\]

\[
\Delta = \frac{\omega_2 - \omega_1}{\omega_0} \tag{3.4}
\]

\[
\omega_0 = \sqrt{\omega_1 \omega_2} \tag{3.5}
\]
where $\omega_0$ is the center frequency, filter bandwidth is equal:

$$\text{BW} = \omega_2 - \omega_1$$  \hspace{1cm} (3.6)

The lumped elements for Fig. 56(b) can be found from the following equations [37]:

$$L_k = \frac{Z_0}{\omega_0 g_k \Delta}, \quad C_k = \frac{g_k \Delta}{\omega_0 Z_0}, \quad k = 1,3,5,...$$  \hspace{1cm} (3.7)

$$L_k = \frac{g_k \Delta Z_0}{\omega_0}, \quad C_k = \frac{1}{g_k \omega_0 \Delta Z_0}, \quad k = 2,4,6,...$$  \hspace{1cm} (3.8)

The main important methods used to design these circuits are Butterworth or Maximally flat and Chebyshev. The element values of low pass Butterworth prototype network are determined by using the following equations:

$$g_0 = 1$$

$$g_k = 2 \sin \left[ \frac{(2K - 1)\pi}{2n} \right], \quad k = 1,2,\ldots,n$$  \hspace{1cm} (3.9)

$$g_{n+1} = 1$$

For the Chebyshev, using the following equations:

$$g_0 = 1$$

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

$$g_k = \frac{4a_{k-1}a_k}{b_{k-1}b_{k-1}}, \quad K = 2,3,\ldots,n$$  \hspace{1cm} (3.10)

$$g_{n+1} = 1, \quad \text{for } n \text{ odd}$$
\[ g_{n+1} = \coth^2\left(\beta/4\right), \quad \text{for } n \text{ even} \]

where
\[ a_k = \sin\left[\frac{(2k-1)\pi}{2n}\right], k = 1, 2, \ldots, n \]

\[ b_k = \gamma^2 + \sin^2\left[\frac{K\pi}{n}\right], k = 1, 2, \ldots, n \quad (3.11) \]

\[ \beta = \ln\left[\coth\left(\frac{A}{2 \cdot 8.686}\right)\right], \quad \gamma = \sinh\left(\frac{\beta}{2n}\right) \]

where \( A \) is the bandpass ripple in decibels, and the relation between \( A \) and \( \varepsilon \) is given by the following expression [40]:

\[ A = 10\log(1 + \varepsilon^2) \quad (3.12) \]

### 3.5 Design Tunable BSF with Tuning Diode

BSF is designed in the first order with series inductor and capacitor with transmission line has a 50 \( \Omega \) impedance as shown in Fig. 57, and center frequency is 3 GHz with 20% bandwidth. Fig. 58 shows the equivalent lumped circuit of Fig. 57.

By using the Butterworth method, the element values for the low pass prototype can be calculated by using equations (3.9):

The findings are: \( g_1 = 1.000 = L_1, \quad g_2 = 2.000 = C_2, \quad g_3 = 1.000 = L_3 \)

By using the equations (3.7, 3.8) to transfer these values to BSF,

\[ L_1 = \frac{g_1\Delta Z_0}{\omega_0} = \frac{1 \times 0.2 \times 50}{2\pi 3^9} = 5.305e^{-10} \]
The microstrip resonator was used with length $BL = \frac{\lambda}{2}$ from port to port, but $BL = \frac{\lambda}{4}$ was used from the port to the middle point, at the point that needs to contact the filter on it, as shown in Fig. 57.

From ADS software the length and width of microstrip resonator can be calculated with substrate FR-4 copper board, $5.813 \times 10^7$ conductivity, 0.002 loss tangent, 60 mil thick dielectric and 0.50 mil thick copper. Thus, the length of $BL = \frac{\lambda}{4}$ was equal to 408.767 mil, and the width $= 117.13$ mil.

Fig. 59 shows a circuit were designed by ADS, it used two transmission lines with 50 $\Omega$ implantation by microstrip lines to connect as terminals. Fig. 60 shows fabrication circuit designed by lumped components.
Fig. 57 BSF Circuit with Microstrip Resonator

Fig. 58 Equivalent BSF Lumped Circuit

Fig. 61 shows the simulated variation in $S_{11}$ and $S_{21}$ with frequency for the p-n junction varactor tuned band stop filter with tuning voltage from 0 to 5 V. The notch frequency of the filter can be changed from 550 MHz to 830 MHz by the application of a tuning voltage of 5 V.

The layout from ADS is downloaded to LPKF machine to fabricate the filter on FR-4 substrate. For some filters, discrete inductors and p-n junction varactors were
assembled by soldering. The band stop filter is characterized using an Agilent Network Analyzer.

Voltages were applied of 0 – 5 V in the circuit in Fig. 60 to tune it. Fig. 62 shows the measured variation of $S_{11}$ and $S_{21}$ with frequency for band stop filter tuned with p-n junction varactor. The notch frequency can be varied from 390 MHz to 720 MHz by applying a tuning voltage of 5 V and the notch depth is about 20 dB.

The bandwidth of this BSF is 169 MHz. The quality factor (Q) effect on filter response, and the Q factor can be found from the following equation:

$$Q = \frac{f_0}{\Delta f}$$

(3.13)

Thus, by using the previous equation, Q factor for this filter was found between 13.57 to 15.565.
Fig. 60 Fabricated BSF with Tuning Diode Circuit

Fig. 61 Simulated Variation of $S_{11}$ and $S_{21}$ with Frequency for Band Stop Filter Tuned with p-n Junction Varactor
Fig. 62 Measured Variations of $S_{11}$ and $S_{21}$ for Band Stop Filters Tuned with p-n Junction Varactor

3.6 Design Tunable BSF with Ferroelectric Capacitor

The same previous circuit is used but with change in the tuning diode to 6.7 pf ferroelectric capacitor, and this capacitor has character as shows in Fig.54. Ferroelectric varactors are electrically connected to FR-4 board using indium ribbons by pressure bonding. Fig. 63 shows, the prototype of the filter fabricated on a FR-4 substrates with BST tunable capacitor. Fig. 64 shows, the measured return loss and insertion loss for the tunable band pass filter implemented with ferroelectric capacitor. The insertion loss in the pass band is about 2.3 dB and at the notch frequency is about 13 dB. The notch frequency can be tuned from 600 MHz to 810 MHz by applying tuning voltage from 0 to 5 V. Similar results are obtained by applying negative tuning voltage for the BST varactor. By using equation (3.13) to find Q factor for this filter. Q factor equals from 4.983 to 6.697.
Fig. 63 Fabricated BSF with BST Varactor

Fig. 64 Variation of Return Loss and Insertion Loss with Frequency for Tunable Band Pass Filter with BST Capacitor

Fig. 65 shows the comparison of simulated and measured results for single stop band filter fabricated with tunable BST capacitors. The insertion loss for the simulated filters is higher than the measured value. This may be because of lower “Q” factor for the ferroelectric capacitors due to its series resistance. The series resistance of the
ferroelectric capacitor is not accounted for in the simulation. The simulated and measured shift in notch frequency with applied tuning voltage is in close agreement.

### 3.7 Design 3\textsuperscript{rd} Order Tunable BSF with Tuning Diodes

The 3\textsuperscript{rd} order of BSF is designed with 3 tuning diodes and three inductors. Third order multi-section band pass filters were also simulated with ADS. Fig. 66 shows the fabrication of 3\textsuperscript{rd} order filter. In this circuit, $S_{21}$ parameter was better than previous circuit of single diode, it becomes around -25 dB. Voltage was applied from 0 to 5V, that makes this circuit tuned from 883 MHz to 999 MHz (Band Width = 116 MHz), as shown in Fig. 67. The return loss in this design was around 0.55dB, it is also better than a single diode circuit. Q factor for this filter is between 46.474 to 75.682 when voltage was applied from 0 to 5V.

![Comparison of Measured and Simulated Response for the First Order BSF](image)

Fig. 65 Comparison of Measured and Simulated Response for the First Order BSF
3.8 Design 3rd Order Tunable BSF with Ferroelectric Capacitor

This filter is designed like the previous filter, but in 3rd order, Fig. 68 shows the prototype of the filter fabricated on a FR-4 substrates with BST tunable capacitor.
Voltage was applied from -5 to 5V. Fig. 69 shows the variation of return loss and insertion loss with frequency for a third order tunable filter. Comparison of Fig. 64 and Fig. 69 show that the return loss in the bandstop filter that has improved, and the maximum return loss is equal to -20 dB obtained at a tuning voltage of 5 V. Q factor is changed from 4.267 to 9.968.

Fig. 68 The Prototype of the Filter Fabricated on a FR-4 Substrates with BST Tunable Capacitor

Fig. 69 Variation of Insertion Loss and Return Loss for Third Order Band Stop Filter with BST Capacitor
CHAPTER 4

DESIGN, SIMULATION, AND IMPLEMENTATION OF TUNABLE DUAL BAND BANDSTOP FILTER

4.1 Introduction

A new approach to designing tunable Dual Band BandStop Filters (DBBSFs) will be presented in this chapter. The Dual-Band Band-Stop Filters (DBBSFs) and Band-Pass Filters (DBBPFs) have become the most important circuit components in modern communication devices. Recently, significant numbers of research papers have been published on this topic. Most of these papers on the topic of fixed frequency design for dual bands primarily are implemented without incorporating tuning. Tuning is, however, very necessary for optimum efficiency in some cases. Tuning is important in the circuits identified because we can change the center frequency and width, in multiple bands of frequencies, by applying voltage without the need to redesign circuits. Thus, this leads the circuits to become smaller, thereby increasing the efficiency and respectively incurring reduction in cost.

Fig. 70 shows a schematic of the lumped element Dual-Band Band-Pass Filter, with center frequency 1 GHz and 3 GHz. Their Advanced Design System Software (ADS) simulated insertion loss and return loss variation in frequency is shown in Fig.
The lumped element circuit for dual band stop filters is shown in Fig. 72. The ADS simulated insertion loss and return loss frequency response of the circuit is shown in Fig. 73.

4.2 Principles of Design Coupled-Line Microstrip Filters

As was explained in chapter two, there are different methods to design dual-band passband or bandstop filters. Coupled-Line Microstrip is one of these methods. The principles of coupled-line microstrip filters are now introduced. Cascaded sections of coupled resonators to implement this filter is used. The filter response is synthesized by modifying the even and odd-mode characteristic impedances of each resonator. Fig. 74 shows open and short circuited coupled-line resonators and their equivalent distributed models [41].

Open-circuited coupled-line resonators are more commonly used because they do not require ground via-holes. The following equations illustrate how to design an open-circuited coupled-line section, which is most used to design band pass filters, and can utilize the same formulas to operate with a short-circuited coupled-line sections.

The following equation gave us the two-port Z-parameters of the open-circuited coupled-line section in Fig. 74 [41].

\[
\begin{bmatrix}
Z_{11} & Z_{21} \\
Z_{21} & Z_{22}
\end{bmatrix} = -\frac{j}{2} \begin{bmatrix}
Z_{0e} \cot(\theta_e) + Z_{0o} \cot(\theta_o) & Z_{0e} \csc(\theta_e) - Z_{0o} \csc(\theta_o) \\
Z_{0e} \csc(\theta_e) - Z_{0o} \csc(\theta_o) & Z_{0e} \cot(\theta_e) + Z_{0o} \cot(\theta_o)
\end{bmatrix}
\] (4.1)

where e and o are even and odd modes.

For \( \theta_e \approx \theta_o \approx \beta l \), the above expression reduce to

\[
\begin{bmatrix}
Z_{11} & Z_{21} \\
Z_{21} & Z_{22}
\end{bmatrix} = -\frac{j}{2} \begin{bmatrix}
(Z_{0e} + Z_{0o}) \cot(\beta l) & (Z_{0e} - Z_{0o}) \csc(\beta l) \\
(Z_{0e} - Z_{0o}) \csc(\beta l) & (Z_{0e} + Z_{0o}) \cot(\beta l)
\end{bmatrix}
\] (4.2)
Fig. 70 Schematics of Dual-Band Band Pass Filter Lumped Circuit

Used in ADS Simulation

Fig. 71 The variation of Return Loss and Insertion Loss with Frequency for DBBPF
Fig. 72 The Schematics of Dual-Band Band Stop Filter Lumped Circuit

Used in ADS Simulation

Fig. 73 Variation of Return Loss and Insertion Loss with Frequency for DBBSF
The Z-parameters of the stub-network are found by the following equations:

\[ Z'_{11} = \frac{Z_1}{j \tan(\beta l)} + \frac{Z_2}{j \tan(\beta l)} \] (4.3)

\[ Z'_{22} = \frac{Z_1}{j \tan(\beta l)} + \frac{Z_3}{j \tan(\beta l)} \] (4.4)

\[ Z'_{12} = Z'_{21} = \sqrt{\left(\frac{Z'_{22} - \frac{1}{Y'_{22}}}{Z'_{22}}\right) Z'_{11}} \] (4.5)

where

\[ \frac{1}{Y'_{22}} = \frac{Z_2}{Z_1 + j \frac{Z_2}{Z_1 + j \tan(\beta l)}} + \frac{Z_3}{j \tan(\beta l)} \] (4.6)

which leads to

\[ Z'_{12} = \sqrt{-Z_1^2 [\tan^2(\beta l) + 1] \frac{Z_1 + Z_2}{\tan^2(\beta l)(Z_1 + Z_2)}} \]

\[ = -j \frac{Z_1}{\sin(\beta l)} \] (4.7)

where \( Z_1 \) is the characteristic impedance of the inverter section, \( Z_2 \) and \( Z_3 \) are the characteristic impedances of the stubs shown in Fig. 74.

When equations (4.3), (4.4), and (4.5) are compared with (4.2), the characteristic of an inverter and stubs in terms of the even and odd impedances of the coupled microstrip line as follows for open-circuited coupled lines is yielded:

\[ Z_1 = \frac{Z_{oo} - Z_{oe}}{2} \]

\[ Z_2 = Z_3 = Z_{oo} \] (4.8)
Fig. 74 Open and Short-Circuited Microstrip Coupled-Line with Equivalent Transmission-Line Circuits.

Short-circuited coupled lines are represented by the following equations:

\[ Y_1 = \frac{Y_{0e} - Y_{00}}{2} \]

\[ Y_2 = Y_3 = Y_{00} \quad (4.9) \]

The previous equations are for symmetrical coupled lines, but the following formulas are the equation for asymmetrical coupled lines for open-circuited coupled lines:

\[ Z_1 = \frac{1}{2} \sqrt{\left( z_{0e}^a - z_{00}^a \right) \left( z_{0e}^b - z_{00}^b \right)} \]

\[ Z_2 = \frac{1}{2} (z_{0e}^a - z_{00}^a) - Z_1 \]

\[ Z_3 = \frac{1}{2} (z_{0e}^b - z_{00}^b) - Z_1 \quad (4.10) \]
For short-circuited coupled lines:

\[
Y_1 = \frac{1}{2} \sqrt{(Y_{oe}^a - Y_{oo}^a)(Y_{oe}^b - Y_{oo}^b)}
\]

\[
Y_2 = \frac{1}{2} (Y_{oe}^a - Y_{oo}^a) - Y_1
\]

\[
Y_3 = \frac{1}{2} (Y_{oe}^b - Y_{oo}^b) - Y_1
\]

(4.11)

where \(a\) and \(b\) are the first and second lines and of different width.

Explained are the closed-form expressions to determine even and odd-mode characteristic impedances of the resonators in the case of Chebyshev-type attenuation characteristics:

\[
Z_{oe}^{N+1} = Z_{oo}^{N+1} = Z_0 \left[ 1 + \frac{K_{10}}{Z_0 \sqrt{S}} \sin \left( \frac{\pi \omega_1}{2 \omega_0} \right) \right]
\]

\[
Z_{oo}^{N+1} = Z_{oo}^{N+1} = Z_0 \left[ 1 - \frac{K_{10}}{Z_0 \sqrt{S}} \sin \left( \frac{\pi \omega_1}{2 \omega_0} \right) \right]
\]

(4.12)

\[
Z_{oe}^{k+1} = Z_{oo}^{N-k+1} = \frac{Z_0}{S} \left( M_{k+1,k} + Y_0 K_{k+1,k} \right)
\]

\[
Z_{oo}^{k+1} = Z_{oo}^{N-k+1} = \frac{Z_0}{S} \left( M_{k+1,k} - Y_0 K_{k+1,k} \right)
\]

(4.13)

where

\[
K_{10} = K_{N+1,N} = \frac{Z_0}{\sqrt{g_0 g_1}} = \frac{Z_0}{\sqrt{g_{N+1} g_N}}
\]

\[
K_{k+1,k} = \frac{Z_0}{\sqrt{g_k g_{k+1}}}
\]

\[
M_{k+1,k} = \sqrt{\left( \frac{K_{k+1,k}}{Z_0} \right)^2 + \frac{1}{4} \tan^2 \left( \frac{\pi \omega_1}{2 \omega_0} \right)}
\]
Where \( Z_0 \) is the reference impedance (usually 50\( \Omega \)), \( Z_{0e} \) is the even-mode characteristic impedance, \( Z_{0o} \) is the odd-mode characteristic impedance, \( \omega_0 \) is the center frequency, \( \omega_1 \) is the frequency at the lower edge of the passband, \( N \) is the order of the filter, and \( g_k \) is the value of the components of the lowpass prototype filter that can be calculated by using equations (3.9 or 3.10). The number of coupled-line sections are \( N+1 \), and all coupled-line sections are a quarter-wavelength long at the center frequency [41].

### 4.3 Theory and Design Equations of DBBSF

Microstrip Transmission Line Resonators (MTLRs) have been used widely to build an abundance of RF-Microwave circuits because they are easy to implement, are of small size, and low cost. Fig. 75 has two transmission lines with two impedances, i.e., \( Z_1 \) and \( Z_2 \), and each impedance has different electrical lengths of \( \beta_{t1} \) and \( \beta_{t2} \), respectively [46]. From the theory of the Stepped Impedance Resonators (SIRs), the inductor as a high impedance (\( Z_1 \)) and the capacitor as a low impedance (\( Z_2 \)) can be implemented. These are depicted in equations 4.14 and 4.15:

\[
L_i = \frac{1}{\omega_i g_i \Delta_i} \quad (4.14)
\]

\[
C_i = \frac{g_i \Delta_i}{\omega_i} \quad (4.15)
\]

To determine the input impedance of the Microstrip Transmission Line Resonator as shown in Fig. 75(a) with an open circuit line and the other as a short circuit line, it is
necessary to find $Z_{in}$ for each MTLR as shown in Fig. 76. Input impedance $Z_{in}$, for a transmission line, in Fig. 76 can be written as [45]:

$$Z_{in} = \frac{Z_L + jZ_0\tan\beta_l}{Z_0 + jZ_L\tan\beta_l}$$  \hspace{1cm} (4.16)
When the load impedance $Z_L$ becomes zero incurring a short circuit, the input impedance in equation (4.16) will become:

$$Z_{in} = jZ_0 \tan \beta_l$$  \hspace{1cm} (4.17)

For the open circuit, when $Z_L = \infty$. Then $Z_{in}$ equals:

$$Z_{in} = -jZ_0 \cot \beta_i$$  \hspace{1cm} (4.18)

To find the input impedance in Fig.75 (a), the input impedance need be divided into two sections as shown in Fig. 77, and determine each $Z_{in}$, and then find the total input impedance:

![Fig. 77 Two Transmission Lines Circuit](image)

For the first input impedance $Z_{in1}$:

$$Z_{in1} = Z_1 \frac{Z_L + jZ_1 \tan \beta_{l1}}{Z_1 + jZ_L \tan \beta_{l1}}$$  \hspace{1cm} (4.19)

For the second input impedance $Z_{in2}$:

$$Z_{in2} = Z_2 \frac{Z_{in1} + jZ_2 \tan \beta_{l2}}{Z_2 + jZ_{in1} \tan \beta_{l2}}$$  \hspace{1cm} (4.20)
By substituting the equation (4.19) into equation (4.20) the yield is:

\[
Z_{in2} = Z_2 \frac{Z_1 Z_L + jZ_1^2 \tan \beta_{l1} + jZ_2 \tan \beta_{l2}}{Z_2 + \frac{Z_1 Z_L + jZ_1^2 \tan \beta_{l1}}{Z_1 + jZ_L \tan \beta_{l1}} (j \tan \beta_{l2})} \quad (4.21a)
\]

\[
Z_{in2} = Z_2 \frac{Z_1 Z_L + jZ_1^2 \tan \beta_{l1} + (Z_1 + jZ_L \tan \beta_{l1})(jZ_2 \tan \beta_{l2})}{Z_2(Z_1 + jZ_L \tan \beta_{l1}) + jZ_1 Z_L \tan \beta_{l2} - Z_1^2 \tan \beta_{l1} \tan \beta_{l2}} \quad (4.21b)
\]

\[
Z_{in2} = Z_2 \frac{Z_1 Z_L + jZ_1^2 \tan \beta_{l1} + jZ_1 Z_L \tan \beta_{l2} - Z_2 Z_L \tan \beta_{l1} \tan \beta_{l2}}{Z_1 + jZ_L \tan \beta_{l1}} \quad (4.21c)
\]

\[
\therefore Z_{in2} = Z_2 \frac{Z_1 Z_L + jZ_1^2 \tan \beta_{l1} + jZ_1 Z_L \tan \beta_{l2} - Z_2 Z_L \tan \beta_{l1} \tan \beta_{l2}}{Z_1 Z_2 + jZ_2 Z_L \tan \beta_{l1} + jZ_1 Z_L \tan \beta_{l2} - Z_1^2 \tan \beta_{l1} \tan \beta_{l2}} \quad (4.22)
\]

Where \( \beta_{l1} \) and \( \beta_{l2} \) are the electric lengths of the microstrip, \( Z_1 \) and \( Z_2 \) are the characteristic impedances of each transmission line. The impedance ratio \( (R_Z) \) equals the following:

\[
R_Z = \frac{Z_2}{Z_1} = \frac{Y_1}{Y_2} \quad (4.23)
\]

There are two special cases for the input impedance \( Z_{in2} \) which were outlined in equation (4.22). One case was a short circuit and the other case was an open circuit. For the short circuited line, the load impedance equaled zero \( (Z_L = 0) \). Thus, the input impedance in equation (4.22) will become the following:

\[
Z_{in2}^{\text{short}} = Z_2 \frac{jZ_1^2 \tan \beta_{l1} + jZ_1 Z_L \tan \beta_{l2}}{Z_1 Z_2 - Z_1^2 \tan \beta_{l1} \tan \beta_{l2}} \quad (4.24)
\]
\[ Z_{\text{in2}}^{\text{short}} = jZ_2 \frac{Z_1 \tan \beta_{l1} + Z_2 \tan \beta_{l2}}{Z_2 - Z_1 \tan \beta_{l1} \tan \beta_{l2}} \]  

(4.25) 

by substituting \( R_Z \) in equation (4.25) 

\[ Z_{\text{in2}}^{\text{short}} = \frac{jZ_2 \tan \beta_{l1} + jZ_2 R_Z \tan \beta_{l2}}{R_Z - \tan \beta_{l1} \tan \beta_{l2}} \]  

(4.26) 

For the open circuited line, the load impedance will equal infinity \( (Z_L = \infty) \). Thus, the input impedance in equation (4.22) will become the following:

\[ Z_{\text{in2}}^{\text{open}} = Z_2 \frac{Z_1 - Z_2 \tan \beta_{l1} \tan \beta_{l2}}{jZ_2 \tan \beta_{l1} + jZ_1 \tan \beta_{l2}} \]  

(4.27) 

\[ Z_{\text{in2}}^{\text{open}} = -jZ_2 \frac{Z_1 - Z_2 \tan \beta_{l1} \tan \beta_{l2}}{Z_2 \tan \beta_{l1} + Z_1 \tan \beta_{l2}} \]  

(4.28) 

by substituting \( R_Z \) in equation (4.28) 

\[ Z_{\text{in2}}^{\text{open}} = -jZ_2 + jZ_2 R_Z \tan \beta_{l1} \tan \beta_{l2} \frac{R_Z \tan \beta_{l1} + \tan \beta_{l2}}{R_Z \tan \beta_{l1} + \tan \beta_{l2}} \]  

(4.29) 

the input admittance \( Y_{\text{in}} = 1/Z_{\text{in}} \), further assuming \( Y_{\text{in}} = 0 \) [44]. Thus, the equations (4.26) and (4.29) can be obtained as follows:

For the short circuited line:

\[ R_Z - \tan \beta_{l1} \tan \beta_{l2} = 0 \]  

(4.30) 

\[ \therefore R_Z = \tan \beta_{l1} \tan \beta_{l2} \]  

(4.31) 

For the open circuited line:

\[ R_Z \tan \beta_{l1} + \tan \beta_{l2} = 0 \]  

(4.32) 

\[ \therefore R_Z = -\cot \beta_{l1} \tan \beta_{l2} \]  

(4.33) 

The series inductors will be replaced with high impedance line sections \( (Z_0 = Z_{\text{high}}) \), and the shunt capacitors will be replaced with low impedance line sections.
\( Z_0 = Z_{low} \). In this kind of design, the impedance ratio (\( R_Z \)) should be as high as possible.

The electric lengths for low and high impedance line sections are determined by using the following equations of (4.34) and (4.35) respectively, assuming (\( \beta_l < \pi/4 \))[37].

For low impedance lines:

\[
\beta_l = \frac{g_k Z_l}{Z_0}, k = 1,3,5, ...
\]  \hspace{1cm} (4.34)

For high impedance lines:

\[
\beta_l = \frac{g_k Z_0}{Z_h}, k = 2,4,6, ...
\]  \hspace{1cm} (4.35)

Where \( \beta_l \) is the electric length, \( Z_0 \) is the filter impedance, \( g_k \) are the normalized element values of the low pass prototype, \( Z_l \) and \( Z_h \) are the low and high impedance line sections.

Fig. 78 shows the lumped circuit of the dual band stop band filter at frequencies \( \omega_1 \) and \( \omega_2 \). This circuit will reject unwanted signals at the frequencies \( \omega_1 \) and \( \omega_2 \) that exhibit the response of Dual Band Band Stop Filter.

![Schematics of Lumped Circuit of DBBSF](image-url)
In this work, two micro-strip transmission lines were used to design DBBSF at two frequencies 1.5 and 3.5 GHz, as shown in Fig. 7, with two resonators for input and output with impedance $Z_0 = 50 \, \Omega$. In this circuit, a new approach of implementing this filter occurs by using two MTLRs to design band stop filters with specific frequencies. The low and high impedance line sections can be calculated for each transmission line by modifying equations (4.34) and (4.35) to the following equations:

For **low impudence** lines:

$$\beta_l = \frac{g_k Z_l}{2.05 \times Z_0}, \, k = 1, 3, 5, \ldots$$  \hspace{1cm} (4.36)

For **high impudence** lines:

$$\beta_l = \frac{g_k Z_0}{2.05 \times Z_h}, \, k = 2, 4, 6, \ldots$$  \hspace{1cm} (4.37)

From Fig. 7(b) both $Z_1$ and $Z_2$ impedances can be calculated, and then $Z_{\text{TOTAL}}$ can be found from the following equations:

$$Z_1 = j\omega_0 L_2 + \frac{1}{j\omega_0 C_a}$$  \hspace{1cm} (4.38)

$$Z_2 = j\omega_0 L_4 + \frac{1}{j\omega_0 C_b}$$  \hspace{1cm} (4.39)

$$Z_T = \frac{Z_1 Z_2}{Z_1 + Z_2} \quad \text{eq. (4.40a)}$$

$$Z_T = \frac{(j\omega_0 L_2 + \frac{1}{j\omega_0 C_a})(j\omega_0 L_4 + \frac{1}{j\omega_0 C_b})}{(j\omega_0 L_2 + \frac{1}{j\omega_0 C_a}) + (j\omega_0 L_4 + \frac{1}{j\omega_0 C_b})} \quad \text{eq. (4.40b)}$$
Fig. 79 (a) DBBSF Circuit Using Two MTLR's. (b) Equivalent Lumped Circuit for Two MTLR's.
where \( \omega_0 \) is the center frequency of band stop notch filter, \( L_2, L_4 \) are the inductors for first and second band stop notch filter, \( C_a = C_1//C_5 \), and \( C_b = C_4//C_6 \).

The capacitors \( C_5 \) and \( C_6 \) are the tunable capacitors such as tuning diode or ferroelectric capacitors as shown in Fig. 79(b).

Fig. 80 shows the new approach of connecting two MTLRs parallel, with each one working in a specific frequency band. The input admittances are \( Y_{in1} \) and \( Y_{in2} \), and each one faces opposite sides of each other. Thus, the total admittance is equal [45]:

\[
Y_{Total} = Y_{in1} + Y_{in2}
\]  

\[\text{(4.42)}\]
From equation (4.28), the calculation can be drawn as $Y_{in1}$:

$$Y_{in1}^{open} = jY_2 \frac{Y_2 \tan \beta_{l1} + Y_t \tan \beta_{l2}}{Y_1 - Y_2 \tan \beta_{l1} \tan \beta_{l2}}$$  \hspace{1cm} (4.43)$$

From equation (4.23), the equation (4.43) will be as follows:

$$Y_{in1}^{open} = jY_2 \frac{\tan \beta_{l1} + R_Z \tan \beta_{l2}}{R_Z - \tan \beta_{l1} \tan \beta_{l2}}$$  \hspace{1cm} (4.44)$$

By the same method, the second input admittances $Y_{in2}$ can be find as:

$$Y_{in2}^{open} = jY_4 \frac{Y_4 \tan \beta_{l3} + Y_3 \tan \beta_{l4}}{Y_3 - Y_4 \tan \beta_{l3} \tan \beta_{l4}}$$  \hspace{1cm} (4.45)$$

The relationship between the electric length ($\beta_l$) and the impedance ratio ($R_Z$) can be established by the following formula:

$$\beta_{lT} = \beta_{l1} + \beta_{l2}$$  \hspace{1cm} (4.46)$$

From equation (4.33), $\beta_{l2}$ can be determined:

$$\beta_{l2} = \tan^{-1} \left(- \frac{R_Z}{\cot \beta_{l1}} \right)$$  \hspace{1cm} (4.47)$$

then

$$\beta_{lT} = \beta_{l1} + \tan^{-1} \left(- \frac{R_Z}{\cot \beta_{l1}} \right)$$  \hspace{1cm} (4.48)$$

where $\beta_{lT}$ is the total of the electric length. Fig. 81 shows the relationship between the electrical length with impedance ratio $R_Z$ [45].
Fig. 81 The Relationship Between The Electrical Length with Impedance Ratio $R_z$

4.4 Calculation and Design DBBSF

Tunable DBBSFs that work at two different bands 1.5 GHz and 3.5 GHz, as shown in Fig. 77, were designed. The Butterworth method is selected with $N = 3$ to find the element values for the low pass prototype which is used in equation (3.9). The findings are $g_1 = 1$, $g_2 = 2$ and $g_3 = 1$.

By using Advanced Design System Software (ADS) for both MTLRs in each frequency, the width (W) and the length (L) can be calculated for each microstrip impedance. The dimension of impedance $Z_{rf}$ from $\beta_l = \pi/4$ and $\omega_{rf}$ can be calculated by using ADS. Where $\omega_{rf}$ is the average-frequency of $\omega_1$ and $\omega_2$, frequency equals 2.5 GHz, while $Z_{rf}$ is the characteristic impedance which equals 50 $\Omega$. By using ADS, which dimension was found in MTLR being 3 mm for width (W) and the length (L)
equals 16.657 mm for each one as shown in Fig. 82. From equations (4.14) and (4.15), the values of the inductor $L_2$ and capacitor $C_2$ can be determined for the first branch, and $L_4, C_4$ for the second branch as shown in Fig. 77(b) with bandwidth ($\Delta$) at 10%. The function works at center frequencies that equal 1.5 GHz and 3.5 GHz respectively. The findings were the value of $L_2 = 26.53 \, nH$, $C_2 = 0.42 \, pF$, $L_4 = 11.37 \, nH$ and $C_4 = 0.18 \, pf$.

**Fig. 82** Dual Band BandStop Filter Circuit Used in ADS Simulation

In this work, two kinds of tuning elements are used, i.e, tuning diode and ferroelectric capacitor. Hence, the values of the capacitors $C_2$ and $C_4$ are variables when connecting the tuning diode or the ferroelectric capacitor to $C_a$ and $C_b$, as will be further explained below and shown in Fig. 77(b). These filters were fabricated by using
substrate FR-4 copper board with $5.813 \times 10^7$ conductivity, 0.002 loss tangent, 60 mil thick dielectric and 0.50 mil thick copper. To implement the element components of these filters to the microstrip lines, 17 $\Omega$ for low impedance and 130 $\Omega$ for high impedance will be used, and used to calculate the electric length $\beta_l$ for low and high impedances of capacitor and inductor from equations (4.23) and (4.24). $\beta_l$ of the low impedance (capacitor) equals $19.588^\circ$ and the high impedance (inductor) equals $17.816^\circ$. ADS is used to find the dimension of each microstrip line for the inductor and the capacitor of each branch of the filter. For the first branch (1.5 GHz), the dimension of the inductor (high impedance) is $W=0.294$ mm and $L=7.104$ mm, and for the capacitor, (low impedance) $W=13.2$ mm and $L=5.47$ mm. For the second branch (3.5 GHz), the dimension of the inductor (high impedance) is $W=0.3$ mm and $L=3.05$ mm, and for the capacitor, (low Impedance) $W=13.4$ mm and $L=2.33$ mm as shown in Fig. 82.

Fig. 83 exhibits simulated return loss ($S_{11}$) and insertion loss ($S_{21}$) for the filter without tuning capacitors connected to the resonator parameter outputs of the layout circuit which was designed in Fig. 82 without using any tuning elements.

### 4.5 Calculation S Parameters by Using MATLAB Model Circuit

Return Loss (RL) and Insertion Loss (IL) are called S-Parameters ($S_{11}$ and $S_{21}$). The return loss is the ratio among the power reflected back to the input source and the input power when the load is mismatched. There is less reflection when the RL has higher value and the load is compatibly matched and is defined (in dB) as the following equation [37]:

$$RL = 20 \log\left(\frac{P_{out}}{P_{in}}\right)$$

where $P_{out}$ is the output power and $P_{in}$ is the input power.
Fig. 83 Simulated Variation of $S_{11}$ and $S_{21}$ with Frequency for the Filter

Layout Shown in Fig. 82.

$$RL = -20 \log \left( \frac{P_r}{P_{in}} \right)$$

(4.49)

where $P_r$ is the power reflect from the input source and $P_{in}$ is the input power (source).

$$\therefore RL = -20 \log(\Gamma) \text{ dB}$$

(4.50)

Where $\Gamma$ is the voltage reflection coefficient which equals:

$$\Gamma = \frac{V_0^-}{V_0^+} = \frac{Z_L - Z_0}{Z_L + Z_0}$$

(4.51)
The insertion loss is the ratio between the output power element to the input power, therefore the insertion loss should be as low as possible, and is defined (in dB) as the following equation [37]:

\[ IL = -20 \log \left( \frac{P_i}{P_{in}} \right) \]  \hspace{1cm} (4.52)

where \( P_i \) is the load power and \( P_{in} \) is the input power (source).

\[ RL = -20 \log (T) \text{ dB} \]  \hspace{1cm} (4.53)

Where \( T \) is the transmission coefficient which equals:

\[ T = 1 + \Gamma = 1 + \frac{Z_L + Z_0}{Z_L + Z_0} = \frac{2Z_L}{Z_L + Z_0} \]  \hspace{1cm} (4.54)

From the S matrix [37], the determination of \( S_{11} \) and \( S_{21} \) is made by using the following general equation:

\[ S_{ij} = \frac{V_i^-}{V_j^+} \bigg|_{V_k^+ = 0 \ for \ k \neq j} \]  \hspace{1cm} (4.55)

\[ \therefore S_{11} = \frac{V_1^-}{V_1^+} \bigg|_{V_2^+ = 0} \]  \hspace{1cm} (4.56)

\[ S_{21} = \frac{V_2^-}{V_1^+} \bigg|_{V_2^+ = 0} \]  \hspace{1cm} (4.57)

From the circuit design in Fig. 80, the branch of each filter is connected parallel with \( Z_0 \). \( S_{11} \) and \( S_{21} \) for one branch of this filter will be calculated, which equation can be used for other branch of the filter.

From equations (4.56) and (4.51), \( S_{11} \) will be determined as follows [46]:

\[ Z_L = Z_0 + \left( \frac{1}{\mathcal{Y} / / Z_0} \right) \]  \hspace{1cm} (4.58)
Substituting for $Z_L$ in $S_{11}$, the yield is:

$$S_{11} = -Z_0Y + (2 + YZ_0)$$  \hspace{1cm} (4.59)

where $Y$ is admittance which equals

$$Y = \frac{1}{Z_1} \text{ or } \frac{1}{Z_2}$$  \hspace{1cm} (4.60)

$Z_1$ and $Z_2$ were determined in equations (4.38) and (4.39).

From equation (4.57) $S_{21}$ [47], can be found by applying a source voltage $E_1$ to port 1, as following:

$$I = \frac{E_1}{(Z_L + Z_0)}$$  \hspace{1cm} (4.61a)

$$V_2 = V_2^- = (1/Y/Z_0)I$$  \hspace{1cm} (4.61b)

$$V_1 = V_1^+ + V_1^- = V_1^+(1 + S_{11}) = Z_LI \implies V_1^+ = Z_LI/(1 + S_{11})$$  \hspace{1cm} (4.61c)

Dividing equation (4.61b) by (4.61c), the yield is:

$$S_{21} = \frac{V_2^-/V_1^+}{(1/Y/Z_0)(1 + S_{11})/Z_L}$$  \hspace{1cm} (4.61d)

$$\therefore S_{21} = \frac{2}{2 + YZ_0}$$  \hspace{1cm} (4.62)

By applying the equations (4.38), (4.39), (4.59) and (4.62) in the MATLAB program, with the values of lumped elements $L_2$, $C_2$, $L_4$ and $C_4$ received from Fig. 79, a model of S parameters can be found as shown in Fig. 84. When Figs. 83 and 84 are compared for S Parameters for DBBSF, an effective relationship of agreement was established between them.
4.6 Design Tunable DBBSF by Using Tuning Diode

The layout of the circuit designed by ADS will be used as shown in Fig. 82 to fabricate this circuit using a tuning diode as shown in Fig. 85. Used was the LPKF machine to fabricate this filter on FR-4 substrate. The tuning element that was used to tune this circuit was the tuning diode. Fig. 86 shows the characteristics of this tuning diode (ZC930) that was used to tune this filter.

The simulation results of S parameters, by using the tuning diode without any tuning (without applying any volt), are shown in Fig 87. Fig. 88 illustrates the measured and empirical results of S parameters without any tuning for both bands. When both
Figs. 87 and 88 are compared, there is optimum agreement is reached between them.

From Fig. 86, the capacitance value is acquired of the tuning diode change from about 10 pF to 1.5 pF when applying voltage from 0V to 5V as shown in Fig. 86 and is indicated in Table 4 from the ZC930 datasheet. In this design, the filter is able to be tuned and effect either independently or together.

![Fig. 85 Fabricated DBBSF with Tuning Diode Circuit](image)

![Fig. 86 The Capacitance Versus Voltage Characteristics of the Tuning Diode ZC930](image)
Fig. 87 Simulated Variation of S Parameters Using Tuning Diode Without Tuning

Fig. 88 Measured Variation of S-Parameters Using Tuning Diode Without Tuning.
Table 4. Capacitance Value of Tuning Diode ZC930

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<th>PART</th>
<th>Capacitance V_R=1V</th>
<th>Capacitance V_R=2.5V</th>
<th>Capacitance V_R=4V</th>
<th>Minimum Q v_R=4V f=50MHz</th>
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<td>MAX. pF</td>
<td>MIN. pF</td>
<td>MAX. pF</td>
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</table>

4.6.1 Tuning First Branch of Band Stop Filter

Voltage was applied on the first branch of the filter and the second branch was held as an open circuit. The simulated and measured variation, in S_{11} and S_{21} for the first band frequency with a tuning voltage of 0 to 5 V, is shown in Fig. 89 and Fig 90 respectively. For the simulation, the center frequency will change from 600 MHz to 1.12 GHz with attenuation, more than -30 dB with fixed second frequency at 3.44 GHz.

The measured notch frequency filter can be changed from 760 MHz to 1.098 GHz by applying a tuning voltage to 5V with attenuation more than -25 dB as shown in Fig. 90. The tuning range is equal to 338 MHz (24% tuning range) and, at the same time, the second band will stay constant at 3.378 GHz, and the average of a return loss is about -0.822 dB.

4.6.2 Tuning Second Branch of Band Stop Filter

The next step is the application of voltage to the tuning diode on the second band stop filter, and the first band stop filter will be kept as an open circuit. Fig. 91 shows the
simulated results of $S_{11}$ and $S_{21}$ for the second band with tuning. As seen in Fig. 91, 0V ($\pm 10$ pF) is applied to the filter, interference occurred between first and second band as shown in Fig. 89. Thus, when the second band filter needs to be tuned, choosing the capacitance value of the tuning diode should be less than 2.9 pF. This procedure indicates that voltage should not be applied at less than a 4 V level in order to block any interference between the bands of the filter, as shown in Fig. 92. The values of the voltage, or the capacitance, depend on the kind of tuning diode you used.

Fig. 89 Simulated Variation of $S_{11}$ and $S_{21}$ for The First Band Stop Filter Tuned With Tuning Diode.
Fig. 90 Measured Variations of $S_{11}$ and $S_{21}$ for The First Band Stop Filters Tuned with p-n Junction Varactor.

Fig. 93 shows us the measure of S-parameters when voltage is applied to the tuning diode. In this case, any level applied which is less than 5 V will not avoid the interference between the filter bands, as was mentioned previously. Therefore, to avoid this issue of potential interference, an experiment was done in absence of the tuning
Fig. 91 Simulated Variation of $S_{11}$ and $S_{21}$ for The Second Band Stop Filter Tuned with Tuning Diode.

Fig. 92 Simulated Variation of $S_{11}$ and $S_{21}$ for The Second Band Stop Filter Tuned with Tuning Diode from 4V to 7V.
diode. In this etude, different values of capacitances were tested at the respective levels of half-oriented range and whole unit(s) increments of 0.1, 0.2, 0.3, 0.5, 1, 1.5, and 2 pF in view of the implementation and disintegrated affect of the voltage via the tuning diode. Fig. 94 shows the measurement results when these capacitors were used. The insertion loss of the second band filter can be changed from 1.647 to 3.336 GHz with a tuning range equal to 1.689 GHz (50.6% tuning range) and attenuation of more than -35 dB, without any tuning for the first branch, keeping constant the center frequency which equals to 1.288 GHz. The tuning range for the second branch is greater, or larger, than the tuning first branch. The return loss ($S_{11}$) for the second band is about -0.468 dB, as shown in Fig. 94.
Fig. 94 Measured Variations of $S_{21}$ and $S_{11}$ for The Second Band Stop Filter Tuned by Applying Capacitors from 0.1pF to 2pF
4.6.3 Tuning First and Second Branches of Band Stop Filter

Fig. 95 shows the measurement of $S_{21}$ and $S_{11}$ for both frequency bands first and second tuning, by utilizing respective tuning diodes. For the first branch voltage from 0V to 5V was applied singularly, but for the second branch from 5V to 7V was applied concurrently. This figure is mixed between Fig. 90 and Fig. 93.

4.7 Design Tunable DBBSF by Using Ferroelectric Capacitor

The same previous circuit was used with the same dimensions and were about 25 mm width and 26 mm length, but this time the ferroelectric capacitor was connected instead of the tuning diode. In this design, two different values of ferroelectric capacitors were used, one to tune the first band and the other for the second band stop filter. Both of them have different and varying characteristics as shown in Figs. 96(a) and (b). Ferroelectric varactors were electrically connected to a FR-4 board by using indium ribbons by pressure bonding. Fig. 97 showed the prototype of the filter fabricated on a FR-4 substrate with two BST tunable capacitors. Each ferroelectric varactor had a different value from the other.

4.7.1 Tuning First Branch of Band Stop Filter

To tune the lower band stop filter, a high capacitance value of ferroelectric varactor was used, as shown in Fig. 96(a). The varactor valued between 0.722 pF at 0V to 0.427 pF at 5V. As was determined, the values of the ferroelectric varactors were too low, and this kind of capacitor was too sensitive. Hence, the parasitic capacitance needed
Fig. 95 Measured Variations of $S_{21}$ for The First and The Second Band Stop Filters

Tuned with p-n Junction Varactor
Fig. 96 Capacitance Versus Voltage Characteristics of a BST Capacitor

(a) For Low Frequency (1.5 GHz), (b) For High Frequency (3.5 GHz)
be taken into consideration in the model. The parasitic capacitance could be ignored at low frequencies without adverse effect, but in high frequencies the parasitic capacitance was definitively proven to be a detriment to the smooth running of the system and a major problem and challenge with which to be dealt.

Fig. 98 follows and indicates the simulated results of the first branch, which can be tuned from 1.18 to 1.24 GHz. Fig. 99 shows the measurement of insertion loss and return loss for the tunable dual band bandstop filter implemented with a single resonator loaded with FE capacitors. The first notch of the filter can be tuned from 570 MHz to 781 MHz. The tuning range is equal to 211 MHz (15% tuning range). As expected, the second notch cannot be tuned, and it stays at 3.18 GHz. The main reason of the difference among the simulation and measurement results, cumulatively on numerous tests, is the parasitic capacitance as was mentioned previously.

![Fig. 97 Fabricated DBBSF with BST Varactor](image)

Fig. 97 Fabricated DBBSF with BST Varactor
Fig. 98 Simulated Variation of $S_{11}$ and $S_{21}$ for The First Band Stop Filter Tuned By Ferroelectric Varactor

Fig. 99 Variation of $S_{21}$ and $S_{11}$ with Tuning Voltage by Loading One Resonator with FE Capacitor
4.7.2 Modeling of Tuning First Branch of Band Stop Filter by Using MATLAB

Fig. 100 also follows and illustrates the modeling of tuning first branch of band stop filter with the values of ferroelectric varactor which was received from Fig. 96(a). The first notch filter tuned from 800 MHz to 1.15 GHz and the second notch filter stayed constant at 3.55 GHz.

![Simulation Results of S21 and S11 from MATLAB Program by Using FE for The First Band Stop Filter](image)

4.7.3 Tuning Second Branch of Band Stop Filter

In this circuit, the second branch of band stop filter was tuned by applying voltage to FE, and the first notch filter was kept as an open circuit with which any voltage could have been applied. The second kind of FE capacitance used for high frequency, as shown in Fig. 96(b), where the capacitance values were between 0.35 pF for 0V to 0.239 pF for 5V. A small value of the FE was chosen to block any interference.
This reduced value was chosen as to insure against any type of negative intervention, or interference, between the first and second band. Simulation for the circuit designed by ADS, in Fig. 80, was founded on the use of previous values of FE historically recognized data. The insertion loss and return loss is reflected in Fig. 101. This band was tuned from 2.72 to 2.83 GHz, and the first band was fixed, or constant, at 1.47 GHz. The measured results, upon loading the second band only, caused a frequency tuning range from 2.16 GHz to 2.55 GHz, as shown in Fig. 102. The tuning range equals 390 MHz (11.7% tuning range), with an attenuation about -25 dB. At the same time, the first notch stayed constant at 1.47 GHz, as shown in Fig. 102.

When the simulation and measurement and quantitative results for the this tuning were compared, the tuning range of fabricated circuit with FE measurement was in agreement to the simulation circuit result. However, there were some differences in the values of center frequencies when voltage was applied from 0V to 5V. The reason remains the same and is attributed to parasitic capacitance.

4.7.4 Modeling Tuning Second Branch of Band Stop Filter by Using MATLAB

As seen in Fig. 103, the second filter tuned from 2 to 2.4 GHz when the low value of FE capacitance was used. Thus, the tuning range was equal to 400 MHz, which calculated higher than the simulated and measured results. Simultaneously, the second notch filter stayed constant at 1.5 GHz, as shown in Fig. 103. Thus, there were a good agreement between the measurement result and this result.
Fig. 101 Simulated Variation of $S_{11}$ and $S_{21}$ for The Second Band Stop Filter Tuned by Ferroelectric Varactor

Fig. 102 Variation of $S_{21}$ and $S_{11}$ with FE Capacitor by Loading Second Resonator
4.7.5 Tuning First and Second Branch of Band Stop Filter

The simulation results of S parameters for tuning both bands are shown in Fig. 104. Fig. 105 illustrates the measured insertion loss of both bands when the double resonators were loaded with FE capacitors. Two bands could be tuned. The first notch could be tuned from 566 MHz to 786 MHz, with an attenuation about -17 dB and the second notch could be tuned from 2.23 GHz to 2.55 GHz, with an attenuation about -25 dB. The frequency tuning ranges for the first and second center frequency bands were 220 MHz (15.6% tuning range) and 320 MHZ (9.6% tuning range), respectively. When the measurement and simulation results were compared, there was a minimal difference in the first and the second branches of the filter.
Fig. 104 Simulated Variation of $S_{11}$ and $S_{21}$ for The First and The Second Band Stop Filter Tuned by Ferroelectric Varactor

Fig. 105 Variation of $S_{21}$ with Bias Voltage by Two Resonators Loaded with Two FE Capacitors
4.7.6 Modeling Tuning First and Second Band Stop Filter by Using MATLAB

Fig. 106 shows us the simulation for the first and the second bands with tuning. The first band tuned from 800 MHz to 1.15 GHz, whereas the second band filter tuned from 2 to 2.4 GHz. Agreement was achieved similarly among these results. This proved the two approaches of simulation and measurement yielded near like outcome. The second band was a clear example of simulation by ADS and the process of measurement reliable for concurring assessment tools in this model. Although, the overall consensus was in close agreement, the first band offered minimal deviation.

Fig. 106 Simulation Results of $S_{21}$ and $S_{11}$ from MATLAB Program by Using FE for The First and The Second Band Stop Filter
CHAPTER 5

CONCLUSIONS AND FUTURE WORK

5.1 Conclusion:

To conclude with, ferroelectric BST capacitors are fabricated on sapphire substrates using a MOSD technique. Single-pole and multi-pole band-stop filters were fabricated using capacitors with coplanar waveguides fabricated on FR-4 substrates, and also using the tuning diode. Measured response of the first order filter shows that the notch frequency can be tuned from 600 MHz to 810 MHz by applying a bias voltage of 5V with an average notch depth of 13dB. For the third order filter, the maximum notch depth obtained is about -20dB.

DBBSFs are implemented on FR-4 substrates, and the BST capacitors are electrically connected to the resonators by indium ribbon pressure bonding. The dual bands can be tuned independently using independent bias sources. A new approach was used to connect two filters in order to be able to tune each band separately. Measured response of the first band stop filter shows that the notch frequency can be tuned from 570 MHz to 781 MHz by applying a bias voltage of 5V with an average notch depth of -17 dB. For the second band stop filter, the maximum notch depth obtained is about -30dB with tuning from 2.16 GHz to 2.55 GHz.
5.2 Future work:

A DBBSFs were fabricated on FR-4 copper board substrate with a ferroelectric varactor by using indium ribbons to connect the ferroelectric varactor to the board. As a result, the effect of a parasitic capacitance will be increased. In future, in order to reduce this effect and also increase the efficiency of this filter, it might be needful to design and integrate all the components on the same substrate to get less loss, less parasitic capacitance and less attenuation.

All the parameters of DBBSFs were measured by using an Agilent network analyzer, set on low level power, about -10 to -15 dB. Any circuit will have a nonlinear output when applying the high power on it. Hence, to operate the DBBSF in high level of the power with liner output, more study needs to be done to improve the design of the DBBSF.

A band-pass filter is just as important as band-stop filter in the RF-Microwave communication circuit. However, the same method should be used to design and model the structure of Dual Band Band-Pass Filter (DBBPF), in order to tune each band separately or simultaneously by using the tuning components, i.e., p-n junction capacitors and ferroelectric capacitors.

Bandwidth is one of the important parameters in the communication circuit system. Thus, DBBSF and DBBPF circuits should be designed to tune the bandwidth, not to tune the center frequency only, and try to design same circuit to tune the center frequency and the bandwidth together.
REFERENCES


[45] Lei Zhu, Sheng Sun, Rui Li "Microwave Bandpass Filters for Wideband Communications" (Wiley Series in Microwave and Optical Engineering), John Wiley&Sons, Inc, 2012


APPENDIX

7.1 MATLAB Code for Calculating S-Parameters of DBBSF Without Tuning

c1 = 0.42e-12;
l1 = 26.53e-9;
c2 = 0.18e-12;
l2 = 11.37e-9;

fa = (0:75e6:5000e6);
S11 = zeros(1,length(fa));
S21 = zeros(1,length(fa));

for m = 1:length(fa)
    wa = 2*pi*fa(m);
    z11= (1i*wa*l1);
    z12=(1/(1i*wa*c1));
    z1=z11+z12;

    z21= (1i*wa*l2);
    z22=(1/(1i*wa*c2));
    z2=z21+z22;

    ZT=(z1*z2)/(z1+z2);
    YT= 1/ZT;

    S111=(-50*YT)/(2+(50*YT));
    S112=abs(S111)
    S11(m) = 20*log10(S112);

    S211=2/(2+(50*YT));
    S221=abs(S211)
    S21(m) = 20*log10(S221);

end
plot(fa,S11)
Hold all;
plot(fa,S21)
axis ([0 5000e6 -50 5])
grid;
7.2 MATLAB Code for Calculating S-Parameters of First Branch of Band Stop

Filter with Tuning by FE

c1 = 0.843e-12:0.0598e-12:1.142e-12
l1 = 26.53e-9;
c2 = 0.18e-12;
l2 = 11.37e-9;

fa = (0:75e6:5000e6);
S21 = zeros(1,length(fa));
S11 = zeros(1,length(fa));

for m = 1:length(fa)
    for n = 1:length(c1)
        wa = 2*pi*fa(m);
        z11= (1i*wa*l1);
        z12=(1/(1i*wa*c1(n)));
        z1=z11+z12;

        z21= (1i*wa*l2);
        z22=(1/(1i*wa*c2));
        z2=z21+z22;

        ZT=(z1*z2)/(z1+z2);
        YT= 1/ZT;
        S111=(-50*YT)/(2+(50*YT));
        S112=abs(S111);
        S11(m,n)= 20*log10(S112);

        S211=2/(2+(50*YT));
        S212=abs(S211);
        S21(m,n)= 20*log10(S212);
    end
end

plot(fa,S11)
hold on
plot(fa,S21)
axis ([0 5000e6 -40 5])
grid;
legend ('C1=5V','C2=4V','C3=3V','C4=2V','C5=1V','C6=0V')
7.3 MATLAB Code for Calculating S-Parameters of Second Branch of Band Stop Filter with Tuning by FE

```matlab
c1 = 0.42e-12;
l1 = 26.53e-9;
c2 = 0.419e-12:0.0222e-12:0.53e-12;
l2 = 11.37e-9;

fa = (0:75e6:5000e6);
S21 = zeros(1,length(fa));
S11 = zeros(1,length(fa));

for m = 1:length(fa)
    for n = 1:length(c2)
        wa = 2*pi*fa(m);
        z11 = (1i*wa*l1);
        z12 = (1/(1i*wa*c1));
        z1 = z11 + z12;

        z21 = (1i*wa*l2);
        z22 = (1/(1i*wa*c2(n)));
        z2 = z21 + z22;

        ZT = (z1*z2)/(z1+z2);
        YT = 1/ZT;

        S111 = (-50*YT)/(2+(50*YT));
        S112 = abs(S111);
        S11(m,n) = 20*log10(S112);

        S211 = 2/(2+(50*YT));
        S212 = abs(S211);
        S21(m,n) = 20*log10(S212);
    end
end

plot(fa,S11)
hold on
plot(fa,S21)
axis ([0 5000e6 -40 5])
grid;
legend ('C1=5V','C2=4V','C3=3V','C4=2V','C5=1V','C6=0V')
```
7.4 MATLAB Code for Calculating S-Parameters of First and Second Branches of Band Stop Filter with Tuning by FE

```matlab
% MATLAB Code for Calculating S-Parameters of First and Second Branches of Band Stop Filter with Tuning by FE

% Initial values
C1 = 0.847e-12:0.059e-12:1.142e-12;
L1 = 26.53e-9;
C2 = 0.416e-12:0.023e-12:0.531e-12;
L2 = 11.37e-9;

fa = (0:25e6:5000e6);
S21 = zeros(1,length(fa));
S11 = zeros(1,length(fa));

for m = 1:length(fa)
    for n = 1:length(C2)
        wa = 2*pi*fa(m);

        z11 = (wa*L1);
        z12 = (1/(wa*C1(n)));
        z1 = z11 + z12;

        z21 = (wa*L2);
        z22 = (1/(wa*C2(n)));
        z2 = z21 + z22;

        ZT = (z1*z2)/(z1+z2);
        YT = 1/ZT;

        S111 = (-50*YT)/(2+(50*YT));
        S112 = abs(S111);
        S11(m,n) = 20*log10(S112);

        S211 = 2/(2+(50*YT));
        S212 = abs(S211);
        S21(m,n) = 20*log10(S212);
    end
end

plot(fa,S11)
hold on
plot(fa,S21)
axis ([0 5000e6 -40 5])
grid;
legend ('C1=5V', 'C2=4V', 'C3=3V', 'C4=2V', 'C5=1V', 'C6=0V')
```

The code calculates the S-parameters for the band stop filter with tuning by FE. It defines the capacitive and inductive values, initializes the S-parameter matrices, and iterates through the frequency values to calculate the S-parameters. Finally, it plots the S11 and S21 parameters against frequency.