INVESTIGATIONS ON WIDEBAND MICROSTRIP BANDPASS FILTERS USING DEFECTED GROUND STRUCTURES (DGS)

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Investigations on Wideband Microstrip Bandpass Filters Using Defected Ground Structures (DGS)

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To my family and friends.
ABSTRACT OF THE THESIS

Investigations on Wideband Microstrip Bandpass Filters Using Defected Ground Structures (DGS)

by

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San Diego State University, 2013

In the front-end transceiver of a communication system, a microwave filter is required to suppress some unwanted frequencies while pass some of the desired frequencies. One of the filter types is a bandpass filter. There are numerous types of bandpass filter design methods. One of the bandpass filters is a microstrip based parallel coupled line bandpass filter, which offers reasonable matching bandwidth.

This thesis presents the investigations results of the wideband microstrip bandpass filters using defected ground structures (DGS). The advantage of DGS is that it improves the matching bandwidth performance of the filter without adding any extended length to the filter design. The proposed bandpass filter using DGS on low loss Roger’s substrate (\( \varepsilon_r = 2.2, \tan \delta = 0.0004 \)) is excited using two edge launchers 50 \( \Omega \) coaxial probes. The reference parallel coupled bandpass filter shows a fractional bandwidth of 62\% (\( S_{11} \leq -10 \text{ dB} \)) with the passband frequency range of 2.9 \text{ GHz} to 5.6 \text{ GHz} and insertion loss better than -0.5 dB.

Various shapes of defected ground plane structures were investigated and their effect on the filter performance was recorded. In comparison to the reference filter, implementation of DGS with the filter offers an improved passband frequency range of 2.65 \text{ GHz} to 5.75 \text{ GHz} and insertion loss and reflection coefficient magnitude better than -0.3 dB and -15 dB, which accounts to a fractional bandwidth of 74\%. The effect of employing electromagnetic bandgap (EBG) structures with the defected ground plane based filter was also studied, which shows a fractional bandwidth of 80\%, the insertion loss and reflection coefficient magnitude are better than -0.3 dB and -15 dB, respectively. So, clearly the performance of the filter is significantly improved by employing the DGS and EBG structures.

This filter was fabricated on a low cost FR-4 substrate (\( \varepsilon_r = 4.5, \tan \delta = 0.04 \)) while physical dimensions of the filter remained unchanged. The prototype filter’s experimental verification was performed using a vector network analyzer. This DGS filter on FR-4 substrate exhibits an insertion loss, \( S_{21} \) more than -3 dB, and reflection coefficient magnitude, \( S_{11} \) less than -10 dB with the center frequency of the proposed filter 2.4 GHz and thereby an operating bandwidth of 64\%. For comparison, the operational bandwidth of the bandpass filter without DGS is only 44\%. 
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CHAPTER 1

INTRODUCTION

In the front ends of the communication systems, we require a filter network so that unwanted frequencies in a communication band can be attenuated while the desired frequencies can be transmitted. If the signal is spread over a large bandwidth, we need to use wideband filters [1]. There are numerous forms of bandpass filter design methods, which are used to meet these specifications like a fractal bandpass filter [2] and parallel coupled line bandpass filter [1, 3]. Usually, parallel coupled line filters have relatively narrower bandwidth, due to the fact that reductions of $w/d$ (width/length) and $s/d$ (spacing/length) using conventional fabrication methods were limited. However, this limitation can be overcome by scaling a gap to 2 $\mu m$.

There are some new structures such as Photonic or Electromagnetic Band Gap (EBG), through which we can enhance the quality of the communication system. In 1987, John [4] and Yablonovitch [5] proposed photonic band gap (PBG) structures [4, 5], which implodes and utilizes metallic ground plane and thus giving a breakthrough in the microwave circuit industry. Through PBG concept we got a structure called Defected Ground Structures (DGS) [6], which has a simpler structure and potentially great applicability to the design of microwave circuits. It provides an extra degree of freedom in the microwave circuit design. They also have the advantage of improving the performance of the systems without adding any extra length of microstrip on the top layer of substrate as they are applied on the ground plane without taking extra space.

Defected Ground Structures have been implemented in many RF and microwave circuits as they provide additional inductive and capacitive components. One dimensional (1-D) periodic DGS have been reported for the reduction of the size in oscillators [7], and was used in amplifiers [8] for the reduction in noise and in coplanar waveguides [9] for the band rejection filter. A 2-D defected ground array was applied to planar circuits [10] to design a low pass filter with high power handling capability. Several DGS structures have been studied for like dumbbell shaped DGS [11], and implemented to the lowpass filter [12].
Furthermore the passband characteristics of the bandpass filter have been improved by studying various Dumbbell DGS in parallel [13].

In this thesis, the effect of the defected ground structures on the performance of wideband bandpass filters is investigated. Parallel coupled line geometry has been selected for the three-pole bandpass filter. Several DGS structures are applied on the bandpass filter to quantitatively investigate the effect of DGS structure on the bandpass filter performance. Furthermore, the effect of the EBG structures is also studied on the filter design. A microstrip bandpass filter was fabricated and measured, to validate the simulation results of the circuit and is compared with the bandpass filter without DGS to analyze the improvement in the filters performance with DGS

### 1.1 Simulation Tool Used

The full wave analysis simulation have been performed throughout the thesis using the Ansoft Designer version 3.5 by Ansoft Corp, which is a Method of Movement (MoM) based commercial program. In general, the method of moments employs the mixed-potential integral equation (MPIE) method, where the calculation is based on breaking up of the geometry onto the triangles and rectangles which collectively known as Mesh is inturn used for infinite dielectric. The tool generates a solution based initial mesh. Frequency sweep can be performed when we want to generate a solution across a range of frequencies. Here discrete and interpolating are the types of frequency sweeps that can be used.

### 1.2 Organization of Thesis

The thesis is organized as follows. Chapter 2 presents the theory used for the design of the microstrip parallel coupled line bandpass filter. Chapter 3 presents the design and the simulations results for microstrip bandpass filter. Chapter 4 presents the different DGS geometries studies including their parametric study. Chapter 5 presents the results of the DGS structures implemented on the bandpass filter. Chapter 6 presents the effect of the EBG structures on the bandpass filter using DGS. The bandpass filter with DGS is fabricated and the measured results are presented in Chapter 7. Finally, Chapter 8 presents the conclusion of the presented work.
CHAPTER 2

THEORY OF PARALLEL COUPLED LINE MICROSTRIP FILTERS

2.1 MICROSTRIP TRANSMISSION LINE

A microstrip transmission line consists of microstrip suspended above a ground plane on the dielectric material. Figure 2.1 shows a microstrip line of width \( W \) suspended over the substrate, which has a dielectric constant of \( \varepsilon_r \) and substrate height \( h \) [14].

![Figure 2.1. General microstrip structure.](image)

2.2 WAVES IN MICROSTRIPS

The electric and magnetic fields in the microstrips ranges over two media, air above and dielectric below, which makes the structure inhomogeneous. The microstrip line does not support a pure TEM wave as pure TEM wave has only transverse components, as the propagation velocity depends only on factors permittivity \( \varepsilon \) and the permeability \( \mu \) of the material. The propagation velocities and the longitudinal components and of electric and
magnetic fields will depend on the material properties and the physical dimensions of the microstrip [14].

### 2.3 QUASI-TEM APPROXIMATION

In the electric and magnetic fields, when the longitude components for the dominant mode of a microstrip line are considerably smaller as compared to the transverse components, then they can be disregarded. For this occurrence, the dominant mode acts like a TEM mode. Also, a TEM transmission line concept can be applied for the microstrip line. This is known as quasi-TEM approximation and is valid for most of the operative frequencies [14].

### 2.4 EFFECTIVE DIELECTRIC CONSTANT AND CHARACTERISTIC IMPEDANCE

In quasi-TEM approximation, the heterogeneous dielectric air media of microstrip lines is interchanged by uniform dielectric material with an effective dielectric permittivity. In microstrips the characteristics of the transmission line is explained by two parameters, first one is the effective dielectric constant $\varepsilon_{re}$ and other is characteristic impedance $Z_c$, which can be attained through quasi-static analysis. In quasi-static analysis of the microstrip, pure TEM is the presumed as the primary mode of wave propagation. The effective dielectric constant $\varepsilon_{re}$ and characteristic impedance $Z_c$ parameters are resolved from the values of two capacitances are,

\begin{align}
\varepsilon_{re} &= \frac{C_d}{C_a} \quad (2.1) \\
Z_c &= \frac{1}{c} \frac{1}{\varepsilon_a C_d} \quad (2.2)
\end{align}

where $C_d$ - Capacitance per unit length with the dielectric substrate present, $C_a$ - capacitance per unit length with the dielectric substrate replaced by air, $c$ - velocity of electromagnetic waves in free space.

For very thin conductors (i.e., $t \to 0$), the equation that provides precision better than one percent is given by [15].
For $w/h \leq 1$:

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + 12 \frac{h}{w} -^{0.5} + 0.04 \ 1 - \frac{w}{h}^2 \right)
\]  \hspace{1cm} (2.3)

\[
Z_c = \frac{\eta}{2\pi \varepsilon_{re}} \ln \frac{8w}{h} + 0.25 \frac{w}{h}
\]  \hspace{1cm} (2.4)

where $\eta = 120\pi$ Ohms is the wave impedance in free space.

For $w/h \geq 1$:

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + 12 \frac{h}{w} -^{0.5} \right)
\]  \hspace{1cm} (2.5)

\[
Z_c = \frac{\eta}{\varepsilon_{re}} \frac{w}{h} + 1.393 + 0.677 \ln \frac{w}{h} + 1.444^{-1}
\]  \hspace{1cm} (2.6)

Accurate expression for the effective dielectric constant is

\[
\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + \frac{10}{u} \right)^{-ab}
\]  \hspace{1cm} (2.7)

where $u = w/h$, and

\[
a = 1 + \frac{1}{49} \ln \frac{u^4}{u^4 + 0.432^2} + \frac{1}{1.87} \ln 1 + \frac{u}{18.1}^3
\]  \hspace{1cm} (2.8)

\[
b = 0.564 \frac{\varepsilon_r - 0.9}{\varepsilon_r + 3}^{0.053}
\]  \hspace{1cm} (2.9)

The precision of this model is better than 0.2% for $\varepsilon_r \leq 128$ and $0.01 \leq u \leq 100$. The characteristic impedance is defined for the following model is

\[
Z_c = \frac{\eta}{2\pi \varepsilon_{re}} \ln \frac{F}{u} + 1 + \frac{2}{u}^2
\]  \hspace{1cm} (2.10)

where $u = w/h$, $\eta = 120\pi$ ohms, and

\[
F = 6 + 2\pi - 6 \exp \left( -\frac{30.666}{u}^{0.7528} \right)
\]  \hspace{1cm} (2.11)

The accuracy for $Z_c = \varepsilon_{re}$ is better than 0.01% for $u \leq 1$ and 0.03% for $u \leq 1000$ [15].
2.5 GUIDED WAVELENGTH, PROPAGATION CONSTANT, PHASE VELOCITY AND ELECTRICAL LENGTH

In quasi-TEM mode, the guided wavelength of microstrip is given by

$$\lambda_g = \frac{\lambda_0}{\varepsilon_{re}}$$  \hspace{1cm} (2.12)

where $\lambda_0$ - free space wavelength at operating frequency $f$.

If the frequency is known in gigahertz (GHz), the guided wavelength can be calculated in millimeters by the following expression

$$\lambda_g = \frac{300}{f\,(GHz)} \frac{1}{\varepsilon_{re}} \, mm$$  \hspace{1cm} (2.13)

and the propagation constant $\beta$ and phase velocity $v_p$ can expressed by

$$\beta = \frac{2\pi}{\lambda_g}$$  \hspace{1cm} (2.14)

$$v_p = \frac{\omega}{\beta} = \frac{c}{\varepsilon_{re}}$$  \hspace{1cm} (2.15)

where $c$ - velocity of light ($c \approx 3.0 \times 10^8 \, m/s$) in free space.

The electrical length ($\theta$) for a given physical length ($l$) of the microstrip is given by

$$\theta = \beta l$$  \hspace{1cm} (2.16)

Therefore, $\theta = \pi/2$ when $l = \lambda_g/4$ is called as quarter-wavelength and when $\theta = \pi$ when $l = \lambda_g/2$ called as half-wavelength microstrip lines. These parameters are considered very important while designing the microstrip filters [15].

2.6 SYNTHESIS OF W/H

Approximate expressions for w/h in terms of $Z_c$ and $\varepsilon_r$ are available.

For $w/h \leq 2$

$$\frac{w}{h} = \frac{8 \exp \frac{A}{2} \exp A}{\exp 2A - 2}$$  \hspace{1cm} (2.17)

with

$$\varepsilon_{re} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \frac{h}{w} \frac{1}{w} \right) + 0.04 \left(1 - \frac{w}{h} \right)^2$$  \hspace{1cm} (2.18)
\[ A = \frac{Z_c}{60} \frac{\varepsilon_r+1}{2} \ln \left(1 + \frac{\varepsilon_r+1}{\varepsilon_r-1} \right) + \frac{0.23}{\varepsilon_r} + \frac{0.11}{\varepsilon_r} \]  

(2.19)

and for \( w/h \geq 2 \)

\[ \frac{w}{h} = \frac{2}{\pi} B - 1 - \ln 2B - 1 \frac{\varepsilon_r-1}{2\varepsilon_r} \ln B - 1 + 0.39 - \frac{0.61}{\varepsilon_r} \]  

(2.20)

With

\[ B = \frac{60 \pi^2}{Z_c \varepsilon_r} \]  

(2.21)

These equations give better precision but to achieve more accuracy, an optimization method can be used based on more accurate analysis models explained previously [15].

### 2.7 Effect of Strip Thickness

The effect of conducting strip thickness \((t)\) was not taken into account in the previous sections (referred to Figure 2.1). The conducting strip thickness \((t)\) is generally negligible when the microstrip line is realized by conducting thin films. Hence, the effect of the conducting strip thickness can be disregarded though its effect on the characteristic impedance \(Z_c\) and effective dielectric constant \(\varepsilon_{re}\) can be included.

For \( w/h \leq 1 \):

\[ Z_c \ t = \frac{\eta}{2\pi \varepsilon_{re}} \ln \frac{8}{w_e(t)/h} + 0.25 \frac{w_e(t)}{h} \]  

(2.22)

For \( w/h \geq 1 \):

\[ Z_c \ t = \frac{\eta}{\varepsilon_{re}} \frac{w_e \ t}{h} + 1.393 + 0.667 \ln \frac{w_e \ t}{h} + 1.444 \]  

(2.23)

Where

\[ w_e(t) = \frac{w}{h} + \frac{1.25 t}{\pi h} \frac{1 + \ln \frac{4\pi w}{t}}{t} \]  

\[ \frac{w}{h} \leq 0.5 \pi \]  

(2.24)

\[ w_e(t) = \frac{w}{h} + \frac{1.25 t}{\pi h} \frac{1 + \ln \frac{2h}{t}}{t} \]  

\[ (\frac{w}{h} \geq 0.5 \pi) \]

\[ \varepsilon_{re} \ t = \varepsilon_{re} - \frac{\varepsilon_r-1}{4.6} \frac{t}{h} \]  

(2.25)
It is noted that $\varepsilon_{re}$, is the effective dielectric constant for $t=0$. From the above expressions, the effect of strip thickness is noted irrelevant on both the characteristic impedance and effective dielectric constant for small values of $t/h$. The effect of strip thickness is important when the conductor loss is taken into account for the microstrip line [14].

### 2.8 Dispersion in Microstrip

The dispersion in microstrips is the phenomenon in which its phase velocity is not a constant but also depends on frequency. The effective dielectric constant $\varepsilon_{re}$ is a function of frequency and can be expressed as the frequency-dependent effective dielectric constant $\varepsilon_{re}(f)$. The effective dielectric constant $\varepsilon_{re}$ attained in earlier sections is based on the quasi-TEM or quasi-static approximation. The functions defined earlier give us a good estimation at low microwave frequencies. The $\varepsilon_{re}(f)$ is expressed as shown in equation 2.26 with consideration of the effect of dispersion,

$$
\varepsilon_{re} f = \varepsilon_{re} - \frac{\varepsilon_r - \varepsilon_{re}}{1 + \frac{f}{f_{50}}} \tag{2.26}
$$

Where

$$
f_{50} = \frac{f_{TM0}}{0.75 + 0.75 - 0.332 \frac{\varepsilon_r - \varepsilon_{re}}{w/h}} \tag{2.27}
$$

$$
f_{TM0} = \frac{c}{2\pi h} \tan^{-1} \frac{\varepsilon_r - 1}{\varepsilon_r - \varepsilon_{re}} \tag{2.28}
$$

$$
m = m_0 m_c \leq 2.32 \tag{2.29}
$$

$$
m_0 = 1 + \frac{1}{1 + \frac{w}{h}} + 0.32 \left(1 + \frac{w}{h}\right)^3 \tag{2.30}
$$

$$
m_c = 1 + \frac{1.4}{1 + w/h} \begin{cases}
0.15 - 0.235 \exp\left(-\frac{0.45f}{f_{50}}\right) & \text{for } \frac{w}{h} \leq 0.7 \\
1 & \text{for } \frac{w}{h} \leq 0.7
\end{cases} \tag{2.31}
$$

where $c$ - velocity of light in free space.
The dispersion model shows, when frequency increases than $\varepsilon_{\text{re}}(f)$ also increases. Thus, $\varepsilon_{\text{re}}(f) \rightarrow \varepsilon_r$ as $f \rightarrow \infty$. The effect of dispersion on the characteristic impedance can be expressed by

$$Z_c f = Z_c \frac{\varepsilon_{\text{re}}(f) - 1}{\varepsilon_{\text{re}} - 1} \frac{\varepsilon_r}{\varepsilon_{\text{re}}(f)}$$  \hspace{1cm} (2.32)

where, $Z_c$ is the quasi-static value of characteristic impedance obtained earlier [15].

### 2.9 MICROSTRIP COUPLED LINES

Microstrip coupled lines are broadly used for realizing microstrip filters. The cross section diagram of microstrip-coupled lines is shown in Figure 2.2 [14]. Here the microstrip lines are in a parallel arrangement with width $W$ and separated by a distance $s$.

![Cross section of coupled microstrip lines](image)

**Figure 2.2.** Cross section of coupled microstrip lines. Source: Hong, Jia-Sheng, and M. J. Lancaster. *Microstrip Filters for RF/Microwave Applications*. New York: John Wiley & Sons, Inc., 2001.

This parallel-coupled microstrip line arrangement shown above supports two quasi-TEM modes, the even mode and the odd mode shown in Figure 2.3 [14]. Figure 2.3(a) shows the even-mode excitation; here a magnetic wall builds up at the symmetry plane because of the same voltage potentials (positive) on microstrip lines. Figure 2.3(b) shows the
odd mode excitation; in this case an electric wall is formed on symmetric plane as microstrip lines have the opposite voltage potentials. Usually, the even mode and odd mode are excited at same period. Since they are not pure TEM modes therefore the propagation take place at distinctive phase velocities and have different permittivities. Hence for modes, the characteristic impedances and the effective dielectric constants play an important role in the characterization of the microstrip-coupled lines [14].

### 2.9.1 Even and Odd Mode Capacitances

The even and odd mode capacitances, $C_e$ and $C_o$ can be expressed as

\[ C_e = C_p + C_f + C'_f \]  \hspace{1cm} (2.33)

\[ C_o = C_p + C_f + C_{gd} + C_{ga} \]  \hspace{1cm} (2.34)

Where, $C_p$ is the parallel plate capacitance between the strip and the ground plane and $C_f$ is the fringe capacitance as if for an uncoupled single microstrip line

\[ C_p = \varepsilon_o \frac{\varepsilon_r w}{h} \]  \hspace{1cm} (2.35)

\[ 2C_f = \frac{\varepsilon_r e}{cz_c} - C_p \]  \hspace{1cm} (2.36)

$C'_f$ is the modification of fringe capacitance $C_f$ of a single line due the presence of another line. An expression for $C'_f$ is given by
\[ C'_f = \frac{C_f}{1 + (A h/s) \tanh(8 h/s)} \]  

(2.37)

where,

\[ A = \exp[-0.1 \exp(2.33 - 2.53 w/h)] \]  

(2.38)

For the odd mode, \( C_{ga} \) is the fringe capacitances for the air and \( C_{gd} \) is the fringe capacitances for dielectric regions across the coupling gap. A closed-form expression for \( C_{gd} \) is

\[ C_{gd} = \frac{\varepsilon_r \varepsilon_f}{\pi} \ln \coth \left( \frac{\pi s}{4 h} \right) + 0.65 \ C_f \ \frac{0.02 \ \varepsilon_r}{s/h} + 1 - \frac{1}{\varepsilon_r^2} \]  

(2.39)

The capacitance \( C_{ga} \) can be modified from the capacitance of the corresponding coplanar strips and given as

\[ C_{ga} = \varepsilon_0 \frac{K(k')}{K(k)} \]  

(2.40)

\[ k = \frac{s/h}{s/h+2w/h} \]  

(2.41)

\[ k' = 1 - k^2 \]  

(2.42)

and the ratio of the elliptic functions is given by

\[ \frac{K(k')}{K(k)} = \frac{\frac{1}{\pi} \ln \left( \frac{2 + 1 + \frac{k'}{1 + k'}}{1 + \frac{k'}{1 + k}} \right)}{\frac{\pi}{\ln \left( \frac{2 + 1 + \frac{k'}{1 + k'}}{1 + \frac{k'}{1 + k}} \right)}} \quad \text{for } 0 \leq k^2 \leq 0.5 \]  

(2.43)

\[ \frac{K(k')}{K(k)} = \frac{\frac{1}{\pi} \ln \left( \frac{2 + 1 + \frac{k'}{1 + k'}}{1 + \frac{k'}{1 + k}} \right)}{\frac{\pi}{\ln \left( \frac{2 + 1 + \frac{k'}{1 + k'}}{1 + \frac{k'}{1 + k}} \right)}} \quad \text{for } 0.5 \leq k^2 \leq 1 \]

The capacitances shown in equations 2.39 and 2.40 are accurate to 3% over the ranges \( 0.2 \leq w/h \leq 2 \), \( 0.05 \leq s/h \leq 2 \), and \( \varepsilon_r \geq 1 \) [15].

### 2.9.2 Even and Odd Mode Characteristic Impedances and Effective Dielectric Constants

The even and odd mode characteristic impedances \( Z_{ce} \) and \( Z_{co} \) can be obtained from the capacitances, which yields in

\[ Z_{ce} = c \ \frac{1}{C_e \ C_e^{-1}} \]  

(2.44)

\[ Z_{co} = c \ \frac{1}{C_o \ C_o^{-1}} \]  

(2.45)
where $C_e^a$ and $C_o^a$ are even and odd mode capacitances for the coupled microstrip line configuration with air as dielectric.

Effective dielectric constants $\varepsilon_{re}^e$ and $\varepsilon_{re}^o$ for even and odd modes, respectively, can be obtained from $C_e$ and $C_o$ by using the relations

$$
\varepsilon_{re}^e = \frac{C_e}{C_e^a}
$$

and

$$
\varepsilon_{re}^o = \frac{C_o}{C_o^a}
$$

### 2.9.3 More Accurate Design Equations of Coupled Microstrip Lines

More accurate closed-form expressions for the effective dielectric constants and the characteristic impedances of coupled microstrip without considering dispersion are given below

$$
\varepsilon_{re}^e = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2} + \frac{10}{v} \exp(-g)
$$

with

$$
v = \frac{u(20+g^2)}{10+g^2} + g \exp(-g)
$$

$$
a_e = 1 + \frac{1}{49} \ln \frac{v^4 v/52^2}{v^4 + 0.432} + \frac{1}{18.1} \ln 1 + \frac{v}{18.1}^{3}
$$

$$
b_e = 0.564 \frac{\varepsilon_r-0.9}{\varepsilon_r+3}^{0.0053}
$$

where $u = w/h$ and $g = s/h$. The error in $\varepsilon_{re}^e$ is within 0.7% over the ranges of $0.1 \leq u \leq 10$, $0.1 \leq g \leq 10$ and $1 \leq \varepsilon_r \leq 18$.

$$
\varepsilon_{re}^o = \varepsilon_{re} + 0.5 \varepsilon_r + 1 - \varepsilon_{re} + a_o \exp -c_o g^{ab}
$$

with

$$
a_o = 0.7287 \varepsilon_{re} - 0.5 \varepsilon_r + 1 - \exp -0.179u
$$

$$
b_o = \frac{0.747\varepsilon_r}{0.15+\varepsilon_r}
$$

$$
c_o = b_o - b_o - 0.207 \exp -0.414u
$$
where \( \varepsilon_{re} \) is the static effective dielectric constant of single microstrip of width \( W \) as discussed previously. The error in \( \varepsilon_{re}^0 \) is stated to be on the order of 0.5%.

The even and odd mode characteristic impedances given by the following closed-form expressions are accurate to within 0.6% over the ranges \( 0.1 \leq u \leq 10 \), \( 0.1 \leq g \leq 10 \) and \( 1 \leq \varepsilon_r \leq 18 \).

\[
d_o = 0.593 + 0.694 \exp(-0.526u) \tag{2.56}
\]

where \( \varepsilon_{re} \) is the static effective dielectric constant of single microstrip of width \( W \) as discussed previously. The error in \( \varepsilon_{re}^0 \) is stated to be on the order of 0.5%.

The even and odd mode characteristic impedances given by the following closed-form expressions are accurate to within 0.6% over the ranges \( 0.1 \leq u \leq 10 \), \( 0.1 \leq g \leq 10 \) and \( 1 \leq \varepsilon_r \leq 18 \).

\[
Q_1 = 0.8685u^{0.194} \tag{2.57}
\]

\[
Q_2 = 1 + 0.7519g + 0.189g^{2.31} \tag{2.58}
\]

\[
Q_3 = 0.1975 + 16.6 + \frac{8.4}{g} \frac{6^{-0.387}}{g^{10}} + \frac{1}{241} \ln \frac{g^{10}}{1 + g/3.4^{10}} \tag{2.59}
\]

\[
Q_4 = \frac{2Q_1}{Q_2} \cdot u^{0.3} \exp -g + [2 - \exp(-g)]u^{-Q_3} \tag{2.60}
\]

\[
Q_5 = 1.794 + 1.141 \ln 1 + \frac{0.638}{g + 0.517g^{2.43}} \tag{2.61}
\]

\[
Q_6 = 0.2305 + \frac{1}{281.3} \ln \frac{g^{10}}{1 + \frac{g^{10}}{5.1^{10}}} + \frac{1}{5.1} \ln (1 + 0.598g^{1.154}) \tag{2.62}
\]

\[
Q_7 = \frac{10 + 190g^2}{1 + 82.3g^2} \tag{2.63}
\]

\[
Q_8 = \exp -6.5 - 0.95 \ln g - (g/0.15)^5 \tag{2.64}
\]

\[
Q_9 = \ln Q_7 \cdot Q_8 + \frac{1}{16.5} \tag{2.65}
\]

\[
Q_{10} = Q_4 - \frac{Q_8}{Q_2} \exp \frac{Q_8 \ln u}{uQ_9} \tag{2.66}
\]

then,

\[
Z_{ce} = \frac{Z_c \varepsilon_{re}/\varepsilon_{re}^0}{1 - Q_4 \varepsilon_{re}Z_c/377} \tag{2.67}
\]

\[
Z_{co} = \frac{Z_c \varepsilon_{re}/\varepsilon_{re}^0}{1 - Q_{10} \varepsilon_{re}Z_c/377} \tag{2.68}
\]
where $Z_{co}$ is the characteristic impedance of $Z_c$ in odd-mode

$Z_{ce}$ is the characteristic impedance of $Z_c$ in even-mode

$Z_c$ is the characteristic impedance of single microstrip of width $w$.

The characteristic impedance and effective dielectric constant expressions, are given above, which can be used to obtain accurate values of capacitances for the even and odd modes defined in equation 2.44-2.47 [14].

### 2.10 Parallel Coupled-Line Bandpass Filters

The general configuration of parallel coupled-line microstrip bandpass filter is shown in Figure 2.4. The microstrips lines are placed parallel to each other, which provides large coupling between lines because of their spacing and gives a larger bandwidth than other structures. The design equations for parallel coupled line bandpass filter are given by

\[
\frac{J_{01}}{Y_0} = \frac{\pi FBW}{2g_0g_1}
\]

\[
\frac{J_{j,j+1}}{Y_0} = \frac{\pi FBW}{2} \frac{1}{g_jg_{j+1}} \quad j = 1 \text{ to } n - 1
\]

\[
\frac{J_{n,n+1}}{Y_0} = \frac{\pi FBW}{2g_ng_{n+1}}
\]

$J_{j,j+1}$ - characteristic admittances of $J$-inverters

$Y_0$ - characteristic admittance of the terminating lines.

Here $n$ tells us the order of filter. For characterization of the $J$-inverters attained earlier, the even and odd mode characteristic impedances of the coupled microstrip line are given by [14].

\[
(Z_{0e})_{j,j+1} = \frac{1}{Y_0} 1 + \frac{J_{j,j+1}}{Y_0} + \frac{J_{j,j+1}}{Y_0}^2 \quad j = 0 \text{ to } n
\]

\[
(Z_{0o})_{j,j+1} = \frac{1}{Y_0} 1 - \frac{J_{j,j+1}}{Y_0} + \frac{J_{j,j+1}}{Y_0}^2 \quad j = 0 \text{ to } n
\]

### 2.11 Summary of the Chapter

Here we presented about the microstrip coupled lines and the parallel coupled line bandpass filters and the equations used for the filter design. In the next chapter the design of

the microstrip bandpass filter and its results are discussed, with the design equations used in this chapter.
CHAPTER 3

MICROSTRIP COUPLED LINE BANDPASS FILTER

3.1 INTRODUCTION

In this thesis, the aim is to design the wideband microstrip bandpass filter based on the conventional parallel-coupled lines and then the implementation of defected ground structures (DGS) on it. In this chapter, a parallel-coupled bandpass filter (3-pole) is designed on the substrate and then the proposed defected ground structure is implemented to the filter, which is discussed in chapter 4. The general structure of the proposed parallel-coupled line microstrip bandpass filter (3-pole) is shown in Figure 3.1.

Figure 3.1. General structure of parallel edge coupled microstrip bandpass filters.

3.2 DESIGN OF THE MICROSTRIP BANDPASS FILTER

The general structure of the parallel coupled line bandpass filter with 3rd order is shown in Figure 3.2, is placed on the substrate Rogers RT/Duroid 5880 (dielectric constant, $\varepsilon_r = 2.2$, loss tangent, $\tan \delta = 0.0004$) with the height of 20 mil is simulated with the
The method of movement (MoM) based commercial tool Ansoft Designer. The Chebyshev prototype with a passband ripple of 0.218 dB and the center frequency $f_0$ of 4 GHz is used for the design for the parallel-coupled line microstrip bandpass filter. Using the design equations 68-72 gives us the design parameters, which are listed in Table 3.1.

Here the microstrip coupled transmission lines are placed on the top of substrate, with the two feeds given by 50Ω edge launching probe-fed as shown in Figure 3.3. Table I shows the parametric values for the designing of the microstrip bandpass filter, which were defined for Figure 3.1. For filters, only individual scattering parameters $|S11|$ and $|S21|$ are studied because of their symmetrical design, since $|S11|$ is very nearly equal to $|S22|$ and $|S21|$ is very nearly equal to $|S12|$ [1]. Figure 3.3 shows circuit model of the 3rd order edge coupled microstrip bandpass filter on RT Duroid 5880 substrate.

The 3-D view of the microstrip bandpass filter in Ansoft Designer is shown in Figure 3.4.
Table 3.1. Parametric Values for Bandpass Filter Defined in Figure 3.1 for RT Duroid 5880 Substrate

<table>
<thead>
<tr>
<th>S.No.</th>
<th>Parameter</th>
<th>Value (in mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>$l_1$</td>
<td>13.44</td>
</tr>
<tr>
<td>2.</td>
<td>$l_2$</td>
<td>13.46</td>
</tr>
<tr>
<td>3.</td>
<td>$W_1$</td>
<td>.21</td>
</tr>
<tr>
<td>4.</td>
<td>$W_2$</td>
<td>.14</td>
</tr>
<tr>
<td>5.</td>
<td>$S_1$</td>
<td>0.089</td>
</tr>
<tr>
<td>6.</td>
<td>$S_2$</td>
<td>0.098</td>
</tr>
</tbody>
</table>

Figure 3.3. Three-pole edge coupled microstrip bandpass filter on RT Duroid 5880 Substrate.

Figure 3.4. Three-pole edge coupled line microstrip bandpass filter.
3.3 Simulation Results of the Microstrip Bandpass Filter

From the simulation response of 3-pole parallel coupled-line bandpass filter the simulated performance has ($S_{11} = -10$ dB) a passband from 3.24 to 5.27 GHz, and the calculated fractional bandwidth for the microstrip bandpass filter is 47.7%, with a low insertion loss of -0.5 dB and overall reflection coefficient magnitude is better than -14 dB. Figure 3.5 shows the scattering parameters for the bandpass filter on the RT Duroid substrate ($h = 20$ mils).

As one of property of the substrate states, if the height of the substrate is increased then the bandwidth of the filter will be increased. Hence, the microstrip bandpass filter is studied on RT Duroid 5880 ($h = 60$ mil) substrate. The simulated bandpass filter design with RT Duroid 5880 substrate ($h = 60$ mil) increased the performance of the filter. The bandwidth of the filter is increased from 47.7% to 62% and the passband for the microstrip bandpass filter is from 2.9 GHz to 5.6 GHz. Figure 3.6 shows the scattering parameters for the microstrip bandpass filter on RT Duroid 5880 substrate ($h = 60$ mil). Here the insertion loss and reflection coefficient magnitude are better than -0.3 dB and -19 dB, respectively.
3.4 DESIGN OF THE MICROSTRIP BANDPASS FILTER ON FR4 SUBSTRATE

As RT Duroid 5880 substrate is expensive substrate, so the microstrip bandpass filter is modified for the design on the FR4 substrate (dielectric constant, $\varepsilon_r=4.4$, and loss tangent, $\tan\delta = 0.04$) that is lossy as compared to the substrate RT Duroid 5880. The size of the filter is scaled for the design on FR4 substrate. Figure 3.7 shows the circuitry design of three-pole edge coupled microstrip bandpass filter on the FR4 substrate. Keeping the fabrication in the mind, the size of the filter is increased by a factor of twice.

As one the properties states that with the increase in the size of the circuitry of the filter, the bandwidth of the filter shifts towards the lower frequencies. Thus, with the increase in the size of the filter circuitry by a factor of twice, an observation is made that the frequency band is shifted towards the lower frequencies. Table 3.2 shows the parametric values the microstrip bandpass filter on the FR4 substrate. The simulation results are shown in Figure 3.8, with a passband of 1.6 GHz to 3.05 GHz, and the fractional bandwidth of 62.36%. The insertion loss and reflection coefficient magnitude for the filter are verified as -2.2 dB and -17dB, respectively.
Figure 3.7. Three-pole edge coupled microstrip bandpass filter on FR4 substrate material.

Table 3.2. Parametric Values for Bandpass Filter Shown in Figure 3.7 for FR4 Substrate

<table>
<thead>
<tr>
<th>S.No.</th>
<th>Parameter</th>
<th>Value (in mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>$l_1$</td>
<td>18.75</td>
</tr>
<tr>
<td>2.</td>
<td>$l_2$</td>
<td>18.84</td>
</tr>
<tr>
<td>3.</td>
<td>$W_1$</td>
<td>0.3</td>
</tr>
<tr>
<td>4.</td>
<td>$W_2$</td>
<td>0.19</td>
</tr>
<tr>
<td>5.</td>
<td>$S_1$</td>
<td>0.11</td>
</tr>
<tr>
<td>6.</td>
<td>$S_2$</td>
<td>0.135</td>
</tr>
</tbody>
</table>

3.5 SUMMARY OF THE RESULTS

A three-pole parallel coupled-line microstrip bandpass filter is designed on RT Duroid 5880 substrate with height of 20 mils and 60 mils, with fractional bandwidth of 48% and 62%, respectively. With the help of scaling method, an equivalent bandpass filter is designed on FR4 substrate with height of 60 mils and scaled by a factor of twice for fabrication purposes. The fractional bandwidth of the bandpass filter is 62% (1.6 GHz - 3.05 GHz) is achieved parallel to the design of RT Duroid 5880. In the next chapter, different defected ground structures (DGS) will be studied and implemented to the microstrip bandpass filter to study their effect and filter’s response.
Figure 3.8. Scattering parameters for the bandpass filter FR4 substrate (h=60 mils).
CHAPTER 4
DEFECTED GROUND STRUCTURE AND ITS EFFECT ON THE FILTER

Defected Ground Structures (DGS) are the defects that are placed on the ground of planar transmission line that includes microstrip, coplanar and coplanar waveguide. They can be periodic or non-periodic, which disturbs the shield current distribution in the ground plane because of the defects. Due to this disturbance there will be change in the characteristics of a transmission line such as line capacitance and inductance. Therefore, any defect etched in any shape of the ground plane raises the effective capacitance and inductance of the circuit [15]. Also, the defected ground structures generally cause an increase in the effective permittivity, since they got their evolution from photonic band gap (PBG) structures [16].

DGS have a wide number of applications in planar resonators, high characteristic impedance transmission lines, filters, dividers/combiners, oscillators, antenna, and power amplifiers [17, 18].

The general characteristics of DGS’s involve stop band, slow-wave effect and high impedance. DGS needs less circuit size of only a unit or a few periodic structures showing slow-wave effect [19]. DGS have better accuracy than regular defect structures and can be easily designed, and executed which makes it practically important for applications in microwave circuits.

For compact design there are extensive numbers of applications in both active and passive devices [20]. Every DGS structures have its individual characteristics depending on the geometry or their ability of filtering unwanted signals and tuning high-order harmonics. Thus without any increase in circuit complexity, distinctive characteristic can easily be achieved by placing essential DGS configurations for desired circuit operations like wider bandwidth or bandstop characteristics [21].

4.1 Parametric Studies

In the previous chapter, the design of the microstrip bandpass filter was discussed. In this chapter, the parametric study of DGS structures will be analyzed with the bandpass filter
to observe the filters responses like impedance matching and the bandwidth. In parametric studies, one parameter will be varied at a time and the other parameters will be kept constant. For example, in Figure 4.1 the parameter L (length of the structure) will be varied keeping the other parametric values W (width) and D (spacing) constant. Thus, the response of the filter will be determined with respect to the varying parameters length, width and spacing to obtain the optimal values for these parameters.

![Diagram of DGS structure](image)

**Figure 4.1. Parametric study of the proposed DGS structure.**

The proposed DGS structures are obtained by etching the numerous horizontal and vertical defective patterns, with same length, width and spacing between them. Due to the discontinuity of impedance in defective region, an electromagnetic resonance is attained and thus a band-gap is formed.

Types of defected ground structure studied are:
- Horizontal structures
- Vertical structures
- Angled vertical structures
- Combination of vertical and horizontal structures
- Spiral structures
- Symmetric structures
4.2 STUDY ON HORIZONTAL DEFECTED GROUND STRUCTURES

Figure 4.2 shows a microstrip bandpass filter with the horizontal placed defected ground structures. Here the proposed periodic horizontal DGS structures are studied on the ground plane of the filter and its effect are studied on the characteristics of the filter. In the horizontal structures, the length and width of the narrow slots and their spacing have been varied to obtain the superior filter characteristics.

![Figure 4.2. Microstrip bandpass filter with horizontal defected ground structures.](image)

The parametric values for the proposed optimized DGS structures are \(L= 5\, \text{mm}\), \(W = 1.58\, \text{mm}\) and \(S = 0.248\, \text{mm}\), respectively. The scattering parameters for the microstrip bandpass filter with the horizontal defected ground structures are shown in Figure 4.3. Here observations exhibited that the performance of the filter has increased with DGS as compared to filter without DGS. The frequency passband for the bandpass filter without DGS was 3.24 GHz - 5.27 GHz with the fractional bandwidth of 47.7%. After applying the horizontal defects on the ground plane, the frequency passband was 3.0 GHz - 5.5 GHz with the fractional bandwidth of 61% along with insertion loss (-0.5 dB) and reflection coefficient...
Figure 4.3. Scattering parameters for the bandpass filter with horizontal structures.

magnitude (-10.5 dB). So, it occurred that the passband is increased in both the lower and upper end of frequency band of the bandpass filter and these observations concluded that with this type of configuration the performance of the filter has enhanced without any increase in the design of the filter.

4.3 STUDY ON THE VERTICAL DEFECTED GROUND STRUCTURES

Figure 4.4 shows the microstrip bandpass filter using the vertical defected ground structures. Now, the proposed periodic vertical DGS structures are studied on the ground plane of the bandpass filter. The parametric values for the proposed optimized DGS structures are $L = 2.75$ mm, $W = 4.75$ mm and $S = 4$ mm, respectively.

The scattering parameters for the microstrip bandpass filter with the vertical defected ground structures are shown in Figure 4.5. Here, it was observed that the performance of the filter has significantly boosted by applying vertical defects. For this case, the frequency passband is 2.75 GHz - 5.27 GHz with the fractional bandwidth of the filter is 63%, while it was 47.7% for the bandpass filter without DGS. Here the insertion loss and reflection
Figure 4.4. Microstrip bandpass filter with vertical defected ground structures.

Figure 4.5. Scattering parameters for the bandpass filter with vertical structures.
coefficient magnitude are better than -0.3 dB and -12 dB, respectively. Furthermore the passband has improved in lower frequency band of the bandpass filter.

**4.4 STUDY ON THE ANGLED VERTICAL DEFECTED GROUND STRUCTURES**

Figure 4.6 shows a microstrip bandpass filter with the angled vertical defected ground structures. Here the implied periodic vertical DGS structures are studied on the ground plane of the filter with an angle, to observe the effects on the bandpass filter. The angle of the structure is varied with an angle in the multiples of 15 and realized an optimal value of 30 degrees.

![Figure 4.6. Microstrip bandpass filter with angled vertical defected ground structures.](image)

The parametric values for Figure 4.6 are similar as the values for the vertical DGS with an angle of 30 degrees for each DGS structure. The scattering parameters for the angled vertical DGS are shown in Figure 4.7. Here the frequency passband is 2.75 GHz - 5.27 GHz and the fractional bandwidth for the filter is 63% with improved impedance matching to the previous case. The insertion loss and reflection coefficient magnitude are -0.3 dB and -15 dB, respectively. So far, it has been observed that the vertical structures have a better fractional bandwidth with low losses than the horizontal structures.
4.5 Study on the Combination of Vertical and Horizontal Defected Ground Structures

In this section, the microstrip bandpass filter is studied with the arrangement of horizontal and vertical DGS on the ground plane shown in Figure 4.8. Here, the best parametric values of both horizontal and vertical structures are selected. Both structures are joined on the 50 Ω transmission line and then characterized with the arrangement.

The scattering parameters for the combined DGS structures are shown in Figure 4.9. The frequency passband is 2.89 GHz - 5.15 GHz with the fractional bandwidth of 56%. The insertion loss and reflection coefficient magnitude are better than -0.5 dB and -10 dB, respectively.

4.6 Study on the Spiral Defected Ground Structures

In this section, spiral DGS structures are studied to see their effects on the bandpass filter. Figure 4.10 shows the microstrip bandpass filter with spiral defected ground structure on the ground plane. The parametric values for the proposed optimized DGS structures are L = 5 mm, W = 1 mm and S = 1.75 mm, respectively. The simulated results displays multiple frequency bands 2.75 GHz - 3.55 GHz, 3.72 GHz - 3.9 GHz, 4.35 GHz - 4.5 GHz and 4.62
Figure 4.8. Microstrip bandpass filter with combination of vertical and horizontal defected ground structures.

Figure 4.9. Scattering parameters for bandpass filter with combination of vertical and horizontal structures.
GHz - 5.03 GHz. Figure 4.11, shows the scattering parameters for the spiral DGS with an insertion loss and reflection coefficient magnitude better than -0.8 dB and -10dB

4.7 STUDY ON THE SYMMETRICAL DEFECTED GROUND STRUCTURES

In this section, the microstrip bandpass filter is examined with symmetrical square shaped DGS on the ground plane shown in Figure 4.12.

The parametric values for the proposed optimized DGS structures are \( L = 1.5 \text{ mm}, \ W = 1.5 \text{ mm}, \text{ and } S = 0.5 \text{ mm} \), respectively. The scattering parameters for this symmetrical DGS structures are shown in Figure 4.13. Here, the frequency passband is 3.02 GHz - 5.3 GHz with a fractional bandwidth of 54.8%. The insertion loss and reflection coefficient magnitude are better than -0.3 dB and -11 dB, respectively.

Table 4.1 shows the comparison of the different DGS structures, and conveys the response of the filter after the different DGS structures are applied to it.

4.8 SUMMARY OF CHAPTER

The performance of the filter is improved with the help of DGS without any expansion in the size of substrate. Numerous different defected ground structures are
Figure 4.11. Scattering parameters for bandpass filter with spiral structures.

Figure 4.12. Microstrip bandpass filter with symmetrical defected ground structures.
Figure 4.13. Scattering parameters for bandpass filter with symmetrical structures.

Table 4.1. Comparison Between the Different DGS Structures

<table>
<thead>
<tr>
<th>S.No.</th>
<th>Type of Structure</th>
<th>Frequency Passband (GHz)</th>
<th>Bandwidth</th>
<th>Behavior of filter with DGS</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Microstrip Bandpass Filter without DGS</td>
<td>3.24 - 5.27</td>
<td>47.7%</td>
<td>Original filter</td>
</tr>
<tr>
<td>2.</td>
<td>BPF with Horizontal DGS</td>
<td>3.0 - 5.5</td>
<td>61%</td>
<td>Increased band in the upper and lower frequency</td>
</tr>
<tr>
<td>3.</td>
<td>BPF with Vertical DGS</td>
<td>2.75 - 5.27</td>
<td>63%</td>
<td>Increased band in the lower frequency</td>
</tr>
<tr>
<td>4.</td>
<td>BPF with Angled Vertical DGS</td>
<td>2.75 - 5.27</td>
<td>63%</td>
<td>Better impedance achieved in case of vertical DGS structure</td>
</tr>
<tr>
<td>5.</td>
<td>BPF with combination of Horizontal and Vertical DGS</td>
<td>2.89 - 5.15</td>
<td>56%</td>
<td>Shift in frequency band towards the lower frequency</td>
</tr>
<tr>
<td>6.</td>
<td>BPF with Spiral DGS</td>
<td>2.75 - 3.55, 3.72 - 3.9, 4.35 - 4.5, and 4.62 - 5.03</td>
<td>…</td>
<td>Multiple Bands</td>
</tr>
<tr>
<td>7.</td>
<td>Symmetrical DGS</td>
<td>3.02 – 5.3 GHz</td>
<td>55%</td>
<td>Increased band in lower frequency only</td>
</tr>
</tbody>
</table>
reviewed with microstrip bandpass filter and their individual response is observed to the filter. Through various DGS structures, the vertical structures are preferred for further analysis and fabrication because of their better impedance matching and bandwidth with filter. In the next chapter, the design of the microstrip bandpass filter will be examined with the optimized vertical DGS on RT Duroid 5880 and FR4 substrate. The fabrication and measurement results will be discussed in Chapter 7.
CHAPTER 5

MICROSTRIP BANDPASS FILTER USING DGS

5.1 DESIGN OF BANDPASS FILTER WITH DGS

In this chapter the design of the defected ground structures will be implemented to the microstrip bandpass filter. Numerous defected ground structures have been examined in the Chapter 4. It has been observed that the vertical defected ground structure has the better performance as compared to the other arrangements. Therefore for the final design, the microstrip bandpass filter with vertical DGS structure is preferred and optimized for the substrate RT Duroid 5880 ($\varepsilon_r=2.2$, $h=60$ mil).

Figure 5.1 shows the circuit geometry for the proposed coupled line microstrip bandpass filter using DGS structure in Ansoft Corporations High Frequency Structure Simulator. Further, the parametric values are studied ($h = 60$ mil), to accomplish a greater bandwidth and improved impedance matching. Figure 5.2 shows the microstrip bandpass filter with the defected ground structure upon RT Duroid 5880 substrate in Ansoft Designer [13]. The optimized parametric values for the proposed DGS structures are $L = 5.41$ mm, $W = 3.195$ mm and $S = 1$ mm, respectively.

The scattering parameters for the microstrip bandpass filter with DGS arrangement on RT Duroid substrate ($\varepsilon_r=2.2$, $h=60$ mil) is shown in Figure 5.3. The frequency passband is 2.65 GHz - 5.75 GHz with a fractional bandwidth of 74%. Here, the insertion loss and reflection coefficient magnitude are better than -0.3 dB and -15 dB, respectively.

As FR4 substrate is very low-cost and easily available in market compared to RT Duroid 5880 substrate, so the bandpass filter is fabricated on the FR4 substrate ($\varepsilon_r = 4.5$, $\tan \delta = 0.04$, $h = 60$ mil). As the dielectric constant increases, the size of the filter decreases. So the bandpass filter and DGS is scaled w.r.t the original filter design for FR4 substrate, but taking the fabrication in consideration the following filter is scaled by a factor of twice. As the size of the filter is amplified, it was observed that there is a
Figure 5.1. Circuit geometry for the proposed coupled line microstrip bandpass filter using DGS.

Figure 5.2. Microstrip bandpass filter with the proposed DGS on RT Duroid 5880 substrate.

shift in the frequency band towards the lower frequency range keeping the similar bandwidth.

Figure 5.4 shows a microstrip bandpass filter with the proposed defected ground structure on FR4 substrate with the scaled parameters. The parametric values for the proposed optimized DGS structures are $L = 7.6$ mm, $W = 4.5$ mm and $S = 1.41$ mm, respectively.
Figure 5.3. Scattering parameters for the microstrip bandpass filter with DGS structure on RT Duroid 5880 substrate.

Figure 5.4. Microstrip bandpass filter with the proposed DGS on FR4 substrate.

The scattering parameters for the microstrip bandpass filter with DGS are shown in Figure 5.5, the frequency passband of the filter is from 1.45 - 3.05 GHz with a fractional bandwidth of 71%. Here, the insertion loss and reflection coefficient magnitude are better than -2 dB and -12 dB, respectively.
Figure 5.5. Scattering parameters for the microstrip bandpass filter with DGS structure on FR4 substrate.

### 5.2 Summary of Chapter

In this chapter, proposed optimized DGS structure is successfully implemented to the bandpass filter. The fractional bandwidth for the bandpass filter using DGS structures for RT Duroid 5880 and FR4 substrate are 74% and 71% with the insertion loss better than -0.3 dB and -2 dB and reflection coefficient magnitude better than -15 dB and -12 dB. In Chapter 7, the fabrication and measurements of the bandpass filter using DGS on FR4 are discussed to validate results.
CHAPTER 6

EFFECTS OF EBG STRUCTURES ON DGS
MICROSTRIP FILTER

6.1 EBG STRUCTURES

Electro Magnetic Bandgap (EBG) materials have numerous numbers of applications in RF and microwave engineering including microwave and optical cavities, filters, waveguides, and smart artificial surfaces etc. Usually, periodic dielectric or metallic structures were used to attain bandgap behavior with periodicity value comparable to the wavelength. It generally requires 4–5 periods to provide suitable bandgap characteristics (high isolation) along with large physical space, which is essential for combining the EBG into an arrangement [22].

The mushroom type EBG materials (most popular EBG Structure), which can control the electromagnetic wave propagation at certain frequency band for certain arrival angles, is formed from dielectric structures in 1 dimension or 2 dimensions. Its bandgap features are shown in two ways: The suppression of surface wave propagation, and the in-phase reflection coefficient [23]. When a plane wave is exposed on an EBG substrate, it causes change in phase against frequency from +180° to –180°. Then the reflection phase of a surface wave, which is dependent upon the phase of reflected electric field normalized to the phase of incident electric field at the reflecting surface, predicts the angles of incident plane wave. So for normally incident plane wave the perfect electric conductor (PEC) has a reflection phase of 180°, while the perfect magnetic conductor (PMC) has a reflection phase of 0°, which generally does not exist in nature [24, 25].

6.2 STUDY OF EBG STRUCTURES WITH MICROSTRIP BANDPASS FILTER AND DGS

In this section, the microstrip bandpass filter using DGS is studied with EBG structures. In the design, the two layers of RT Duroid 5800 substrate are realized. The top
layer has the microstrip bandpass filter, the lower layer has EBG structures and the defects are positioned on the ground plane.

After experimenting numerous designs, the proposed mushroom type EBG is studied with microstrip bandpass filter using DGS as presented in Figure 6.1. Here the effect of periodicity of the EBG structures is studied on the microstrip bandpass filter, which is discussed in the following cases.

![Figure 6.1. The proposed mushroom type EBG with microstrip bandpass using DGS.](image)

### 6.2.1 Case 1

In this section, the (2 x 7) mushroom type EBG structures are studied with the microstrip bandpass filter and DGS. With the help of parametric analyses, the value of radius and spacing between the structures are determined. The radius for the respective mushroom structure is 2.83 mm and has a spacing of 8.5 mm from the center of the filter. The scattering parameters for the microstrip bandpass filter using DGS with mushroom EBG structures are shown in Figure 6.2. It is observed that with the help of EBG structures the performance of the filter has notably increased as compared to the prior cases in previous chapters. The frequency passband for the filter with EBG structures is 2.50 GHz - 5.80 GHz with a fractional bandwidth of 79.5%. The insertion loss and reflection coefficient magnitude are better than -0.3 dB and -14.5 dB, respectively.

### 6.2.2 Case 2

In this section, the microstrip bandpass filter using DGS is studied with two layers of (4 x 7) mushroom type EBG structure on each side as shown in Figure 6.3, to observe its
effects on the performance of the filter. The parametric values for the EBG structure are similar as case 1, radius of 2.83 mm and a spacing of 8.5 mm.

The scattering parameters for the 2–layer EBG structure with the bandpass filter using DGS are shown in Figure 6.4. An enhanced impedance matching with the reflection coefficient magnitude below than -18 dB and insertion loss better than -0.3 dB, is observed.
as compared to the case 1. The frequency passband for the filter with EBG structures is 2.50 GHz - 5.80 GHz with a fractional bandwidth of 79.5%.

6.2.3 Case 3

In this section, the (6 x 7) 3-layer EBG structures are studied alongside microstrip bandpass filter and DGS to realize its results on the filter’s performance presented in Figure 6.5. Here it occurred that the relative phase, bandwidth and the impedance matching is same as the 2-layer EBG structures with bandpass filter using DGS. The parametric values of the EBG structure are 2.83 mm (radius) and 8.5 mm (spacing). Figure 6.6 illustrations the scattering parameters for the 3-layer EBG with microstrip bandpass filter using DGS. The frequency passband for the filter with EBG structures is 2.50 GHz - 5.80 GHz with a fractional bandwidth of 79.5%. The insertion loss and reflection coefficient magnitude are better than -0.5 dB and -18 dB, respectively.

Figure 6.7 shows the comparison of reflection coefficient magnitude vs frequency of all the cases studied earlier. Figure 6.8 shows the graph for reflection coefficient phase vs. frequency. The phases of all the cases are compared BPF with DGS and no EBG, BPF with EBG and no DGS, and a 50-Ω line with EBG design w.r.t frequency. Here it was detected that a narrowband phase (3.2 GHz – 4.2 GHz) is obtained for bandpass filter using DGS alongside EBG w.r.t - 90 to 90 degrees.
Figure 6.5. Configuration of the 3–layer EBG structure with the bandpass filter using DGS.

Figure 6.6. Scattering parameters for the 3-layer EBG with bandpass filter using DGS.
Figure 6.7. Reflection Coefficient Magnitude vs. frequency for all cases.
Figure 6.8. Reflection coefficient phase vs. frequency for all cases.

**6.3 SUMMARY OF THE CHAPTER**

In this chapter, the effects of EBG structures are studied with the bandpass filter using DGS. From results, it occurred that the performance of the bandpass filter is increased to almost 80% with the insertion loss and reflection coefficient magnitude better than -0.5 dB and -18 dB, respectively. The fractional bandwidth for bandpass filter with DGS was 74%. So, with the analysis of EBG structures it was concluded that the performance of the filter is improved with the assistance of EBG structures. A narrowband phase is also noted here. In the next chapter the fabrication and measurement of the bandpass filter using DGS will be discussed.
CHAPTER 7

FABRICATION, MEASUREMENTS AND RESULTS

The design of the microstrip bandpass filter with defected ground structures (DGS) was discussed in chapter 5. In this chapter, the design of the filter with DGS is fabricated on the FR4 substrate ($\varepsilon_r = 4.5$, $\tan \delta = 0.04$, $h = 60$ mil) is discussed. The filter is fabricated with the help of the CAD Machine at SDSU Engineering Lab. The measurement results are attained with the help of the Anritsu 37269D vector network analyzer from 1 GHz to 4 GHz, which can measure scattering parameters with a frequency range of 40 MHz - 40 GHz. The measurement setup for the bandpass filter with DGS is shown in Figure 7.1.

![Figure 7.1. Measurement setup for the bandpass filter.](image_url)

A 3-pole parallel coupled-line bandpass filter is fabricated with and without defected ground structures. Their measured results are compared to investigate the effects of the DGS on the performance of the filter. As RT Duroid 5880 is a very pricey substrate, so the project is fabricated on FR4 substrate, which is very inexpensive and easily accessible in the market.
Figure 7.2 displays the fabricated microstrip bandpass filter using DGS. Table 7.1 shows the parametric values for the fabricated bandpass filter. The dimensions for the fabricated filter are (90 mm x 15 mm).

Figure 7.2. (a) Top view of the microstrip bandpass filter with DGS, (b) bottom view of the microstrip bandpass filter with the DGS, (c) microstrip coupled lines of the fabricated filter.
Table 7.1. Optimized Design Parameters for the Microstrip Bandpass Filter Using DGS

<table>
<thead>
<tr>
<th>Design Parameters (mm)</th>
<th>FR4L</th>
<th>FR4W</th>
<th>DGS_L</th>
<th>DGS_W</th>
<th>DGS_d</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>90</td>
<td>15</td>
<td>7.6</td>
<td>4.5</td>
<td>1.41</td>
</tr>
</tbody>
</table>

### 7.1 Measurement and Results of Microstrip Bandpass Filter

The measured results of the scattering parameters for the microstrip bandpass filter on FR4 substrate \((\varepsilon_r = 4.5, \tan \delta = 0.04, h = 60 \text{ mil})\) from the network analyzer are shown in Figure 7.3. The measurement results have a frequency passband of 1.9 GHz - 3.05 GHz, with a fractional bandwidth of 44%. The insertion loss and reflection coefficient magnitude are better than -3 dB and -10 dB, respectively. Since FR4 substrate is very lossy material as compared to the RT Duroid 5880, so it was observed that the losses are very high in the higher frequency band as the material has a high loss tangent of 0.04 dB.

![Figure 7.3. Scattering parameters for the microstrip bandpass filter on FR4 substrate.](image-url)
7.2 MEASUREMENT AND RESULTS OF MICROSTRIP BANDPASS FILTER WITH DGS STRUCTURES

The measured results of the scattering parameters for the microstrip bandpass filter with DGS on FR4 substrate \((\varepsilon_r = 4.5, \tan \delta = 0.04, h = 60 \text{ mil})\) are displayed in Figure 7.4. The frequency passband measured for the fabricated filter is 1.62 GHz - 3.18 GHz, with a fractional bandwidth for the filter is 65%. Here, the insertion loss and reflection coefficient magnitude are better than -2 dB and -10 dB, respectively.

![Figure 7.4. Scattering parameters for the microstrip bandpass filter with DGS structures on FR4 substrate.](image)

7.3 SIMULATED VS. MEASUREMENT RESULTS (BANDPASS FILTER WITH DGS)

In this section, the scattering parameters of the measured results are compared with the simulated results of microstrip bandpass filter with DGS that were discussed in Chapter 5. Here, it was noted that the simulated and measured results agrees reasonably. From Figure 7.5, it occurred that the reflection coefficient magnitude agree a lot and the measured results overlaps the simulated results except in the lower frequency band. Better impedance matching was seen in the case of the measured results, though for fabrication purposes each DGS slot on sides is removed. There are some residual errors among simulation and measurements possibly due to the following reasons. First is the incorrect PCB engraving which can be the main reason, second is the embedded parasitic effect of
Figure 7.5. Comparison of simulated and measured result for the microstrip bandpass filter with DGS structures on FR4 substrate (a) insertion loss (b) Reflection Coefficient Magnitude.
SMA connectors, third is the error in fabrication process and the last problem may come from low quality of FR4. The first, second and third problem can be overcome by considering other precise realization process. The parasitic effects of SMA connector are another embedded problem that can be de-embedded by using wafer-measurement system. Figure 7.5 (a) & (b) shows the comparison of simulated and measured results of insertion loss and reflection coefficient magnitude of the bandpass filter with DGS structures on FR4 substrate.

7.4 COMPARISON OF MICROSTRIP BANDPASS FILTER WITH AND WITHOUT DGS

The two designs are fabricated, one is microstrip bandpass filter without any defects on the ground plane and the other one is microstrip bandpass filter with defected ground structures and then their measured results were compared. The comparison of the measured results (insertion loss and reflection coefficient magnitude) of the designs are shown in Figure 7.6 (a) & (b). The frequency band of the microstrip bandpass filter without DGS is 1.95 GHz - 3.03 GHz with a fractional bandwidth of 44% compared to the frequency band for microstrip bandpass filter with DGS is 1.64 GHz - 3.16 GHz with a fractional bandwidth of 64%. So it was spotted that the performance of the filter has increased by using the defects on the ground plane of the microstrip bandpass filter as the bandwidth of the filter is increased from 44% to 64%.

7.5 SUMMARY OF THE CHAPTER

The microstrip bandpass filter using DGS on FR4 substrate is fabricated and measured with the network analyzer with the frequency passband of 1.64 GHz to 3.16 GHz. The fractional bandwidth of the filter with DGS is 64% with an insertion loss and reflection coefficient magnitude better than -3 dB and -10 dB, respectively. The results were then compared with the measured results of the microstrip bandpass filter without DGS, which has a frequency passband of 1.95 GHz to 3.03 GHz, fractional bandwidth of 44%. The conclusion was that there is a significant increase in the performance of filter with the help of DGS without any circuitry expansion.
Figure 7.6. Comparison of microstrip bandpass filter with and without DGS on FR4 substrate (a) Insertion loss (b) Reflection Coefficient Magnitude.
CHAPTER 8

CONCLUSION STUDY

In this thesis, a new way to improve the design and performance of the bandpass filter by using defected ground structures is introduced. Investigations are presented on the rectangular shaped vertical DGS structures and successfully cascaded with the bandpass filter. The fractional bandwidth for the bandpass filter on RT Duroid 5880 substrate (height=60 mil) was 64% is attained with the frequency passband of 2.9 GHz - 5.6 GHz with insertion loss better than -0.3 dB. After employing the defected ground structures the performance of the filter was improved to a fractional bandwidth of 74% without any increase in the circuit. In this case the frequency passband for the filter is 2.65 GHz - 5.75 GHz with insertion loss and reflection coefficient magnitude better than -0.3 dB and -15 dB, respectively. But as RT Duroid 5880 substrate is very expensive and taking fabrication into consideration, the design is scaled for FR4 substrate, which is lossy. For bandpass filter with DGS on FR4, a fractional bandwidth of 71% is attained with the passband frequency of 1.45 GHz - 3.05 GHz. For FR4 substrate, a bandpass filter with and without DGS is fabricated to find out the effect of DGS on the bandpass filter and were measured with the network analyzer to validate results. For microstrip bandpass filter using defected ground structures, a fractional bandwidth of 64% is achieved covering 1.64GHz to 3.16GHz with insertion loss better than -3dB. Compared to this, a fractional bandwidth of 44% with an insertion loss of -3dB was achieved by the microstrip bandpass filter without any DGS structures. There are some residual errors between simulation and measurement maybe due to the reasons of incorrect PCB engraving, the embedded parasitic effect of SMA connectors, error in fabrication process and the low quality of FR4. Therefore, it was concluded that the performance of the filter as well as a better impedance matching can be achieved by applying defects on the ground and also circuit size reduction is seen here. The effect of EBG on the bandpass filter displays the improvement in the performance of the filter which has a fractional bandwidth of about 80%, with the insertion loss better than -0.3 dB. The filter has
an application in Wireless-Local-Area-Network (WLAN) systems that operate in ISM band at 2.45 GHz.
REFERENCES


