CHAPTER 3

ANALYSIS OF MICROSTRIP PATCH ANTENNA ARRAYS WITH EBG STRUCTURES AND FSS

3.1 INTRODUCTION

This chapter explains the methods to improve the directivity, and bandwidth of microstrip antenna array, using EBG structures and the FSS as a superstrate layer. This chapter also details the concept, principles of operation and the geometries of the unit cell of various EBG structures used, along with the microstrip antenna array configuration. Using the Multiconductor Transmission-Line (MTL) theory, the dispersion equations for the mushroom-like EBG and fork-like EBG, are obtained theoretically and verified. In addition to the equivalent circuit of the mushroom-like and fork-like EBG structures, the dispersion diagram is obtained theoretically from the dispersion equation. In this thesis, two approaches are proposed for improving the performance of the microstrip antenna array. In the first method, the EBG structures are placed on either side of the feedline of the microstrip antenna array. The optimal position of the EBG structures on both sides of the feedline is calculated using a parametric study. In the second method, the FSS is placed as a superstrate layer in the aperture coupled microstrip antenna array. The parametric study is done, for identifying the optimal height at which the FSS can be placed from the radiating patch. The parametric studies were performed using the CST microwave studio. In this research, the various EBG structures have been incorporated along with the
microstrip antenna arrays. The proposed microstrip antenna array is fabricated, using the dielectric material FR-4, having a dielectric constant of 4.4 and thickness of 1.6 mm.

The bandwidth and directivity of the microstrip antenna array can be increased, using the following techniques, which are available in the literature.

- Using thick and low permittivity substrates.
- Introducing closely spaced parasitic patches on the same layer of the fed patch (15% BW) (Shackelford et al 2003)
- Using a stacked parasitic patch (multilayer, bandwidth reaches 20%) (Shackelford et al 2003)
- Aperture coupling (10% BW, high back lobe radiation) (Sullivan and Schaubert 1986, Pozar 1986)
- L-probe coupling slot (Luk et al 1998)
- Using loop shaped FSS as the superstrate in an aperture coupled antenna (Prihadi et al 2012)

This research introduces electromagnetic bandgap structures and the frequency selective surface, for the performance improvement of microstrip antenna arrays. Compactness and antenna array area reduction can be achieved through suitable unitcells of the EBG structures. With this idea, the work has designed a set of new geometries of the EBG structures, and the placement of these EBG structures along with the antenna array configuration, for directivity and overall performance enhancement.
3.2 PRINCIPLE OF THE EBG STRUCTURES

Electromagnetic BandGap (EBG) structures are artificial periodic or aperiodic objects, that prevent or assist the propagation of electromagnetic waves in a specified band of frequency, for all angles of arrival and all polarization states. They are usually realized by the periodic arrangement of dielectric materials and metallic conductors. By incorporating a special texture on a conductive surface, it is possible to alter its radiofrequency surface properties. The period of the surface texture is much smaller than the wavelength, the structure can be described using an effective medium model, and its qualities can be summarized by a single parameter, called surface impedance. EBG structures often implemented using the mushroom-like geometry proposed by Sievenpiper et al (1999), utilizing the plain square structure, are still too large to be used in an antenna. The lattice period of the mushroom-like EBG is $0.1\lambda_o$, where $\lambda_o$ is the free space wavelength. A high-impedance surface, shown in cross section in Figure 3.1, consists of an array of metal patches on a flat conductive sheet. They are arranged in a two-dimensional lattice, and are usually formed as metal protrusions, connected to the continuous lower conductor by vertical posts. They can be visualized as mushroom-like EBGs protruding from the surface.

![Figure 3.1 Cross section of a high impedance surface](image)

If the protrusions are small compared to the wavelength, their electromagnetic properties can be described, using lumped circuit elements i.e., capacitors and inductors. These structures are inductive at low
frequencies, and hence support Transverse Magnetic (TM) waves, and are capacitive at high frequencies, which enable them to support Transverse Electric (TE) waves. Within the forbidden gap though, these structures suppress both TE and TM waves from propagating through the structure.

Figure 3.2 (a) and (b) 3x3 Two-D mushroom-like EBG lattice and equivalent circuit of mushroom-like EBG cell

The partial arrangement of a 2D mushroom-like EBG structure and its equivalent circuit are shown in Figure 3.2 (a) and (b). They behave as a parallel resonant lumped circuit (LC), which acts as electric filters to block the flow of currents along the sheet. The proximity of the neighbouring metal elements, provides the capacitance, and the long conducting path linking them together, provides the inductance. Thus, the approach to increase the inductance and capacitance will naturally result in the decrease of the bandgap position.

The central frequency of the bandgap is given in Equation (3.1)

$$f_0 = \frac{1}{(2\pi\sqrt{LC})}$$  \hspace{1cm} (3.1)

where \( L = \mu_0 h \)  \hspace{1cm} (3.2)
\[
C = a\varepsilon_0 \frac{(\varepsilon_r + 1)}{\pi} = \left[ \frac{a+g}{g} \right]
\]  

(3.3)

where \( h \) - thickness of the substrate in mm, \( L \) - inductance in Henry, \( C \) - capacitance in Farad, \( 'b' \) is the period of the EBG structure, i.e., \( b=a+g \), \( g \) is the gap between two adjacent EBG patches, \( a \) is the width of the EBG patch, \( h \) is the thickness of the substrate, \( L \) is the inductance, and \( C \) is the capacitance. \( \mu_0 \) and \( \varepsilon_0 \) are the permeability and permittivity of free space, respectively. It should be noted that the radius effect of the vias cannot be considered in (3.2) and (3.3), even though this seriously affects the bandgap properties. The bandwidth of the electromagnetic bandgap is given as

\[
BW = \frac{1}{\eta \sqrt{\frac{L}{C}}}
\]

(3.4)

where \( \eta \) is the free space wave impedance and \( BW \) is the bandwidth.

The research work focuses on the reflection phase feature of EBG surfaces: when plane waves normally illuminate an EBG surface, the phase of the reflected field changes continuously from +180° to -180° versus frequency. One main application of this feature is that one can replace a conventional Perfect Electric Conductor (PEC) conductive surface with an EBG surface for a microstrip antenna. For this antenna array design, within the operational frequency band of an EBG structure, a microstrip antenna radiates efficiently by having significant return loss, directivity and radiation patterns. The operational frequency band is the overlap of the input-match frequency band, and the surface wave frequency bandgap. It is revealed that the reflection phase can be used to identify the input-match frequency band, inside which a microstrip antenna exhibits good return loss. A dispersion diagram can be used to identify the bandgap within the band, in which the surface waves are suppressed: they help to improve the radiation patterns of
the microstrip antenna array. The bandgap is very close to its input-match frequency band, resulting in an effective operational frequency band.

An evaluation of the performance of a microstrip antenna and antenna array on a thin grounded slab has been discussed in chapter 2. In this case a high dielectric constant substrate has to be used, to achieve a quadratic reflection phase. Although the microstrip antenna shows improved return loss, the directivity and radiation patterns exhibit division in the broadside direction, and the directivity is due to strong surface waves. Thus, the input-match frequency band does not overlap the surface-wave frequency bandgap in the thin grounded dielectric slab case. As a result, there is no effective operational frequency band, so that the thin grounded slab cannot work efficiently as a conductive surface for the microstrip antenna configuration. The Perfect Electric Conductor (PEC) has a $180^\circ$ reflection phase for a normally incident plane wave, while the Perfect Magnetic Conductor (PMC), which does not exist in nature, has a reflection phase of $0^\circ$. Recent studies on EBG structures have revealed that they can satisfy the PMC-like condition in a certain frequency band (Sievenpiper et al. and Yang et al. 1999).

However, the EBG structures are more than a PMC surface, as emphasized in this thesis. The reflection phase of an EBG surface varies continuously from $+180^\circ$ to $-180^\circ$ versus frequency, not only at $180^\circ$ for a PEC surface or $0^\circ$ for a PMC surface. This reflection phase feature makes the EBG surface unique. One potentially main application of this surface is its usage as the conductive surface for a microstrip antenna configuration, which is desirable in many wireless communication systems.

EBG structures can be designed in various shapes, and every shape will have a different frequency bandgap. Something special about the EBG structure is that, it can be designed to be inductive or capacitive. Figure
3.3(a),(b),(c) and (d) shows the partial arrangements of different types of unit cells of various EBG structures available in the previous literature (Sievenpiper et al 1999, Yang et al 2005 and Abedin and Ali 2005).

![Figure 3.3 Various EBG structures and their arrangements](image)

(a) Mushroom-like (b) Fork-like (c) Uni-planar (d) Bent-strip planar

The period of the mushroom-like EBG lattice approaches 10% of the free space wavelength (Sievenpiper et al 1999). The period of the fork-like EBG lattice is reduced to 7% of the free space wavelength (Yang et al 2005). The Uni-planar Compact (UC) EBG is still reduced in size, but employs a high dielectric constant substrate, and the stop band frequency is also high (Yang et al 1999). The bent strip planar EBG has the same lattice period as the UC-EBG, but results in a low stop band frequency with a low dielectric constant, and a high cost substrate (Abedin et al 2005). So, realizing a more compact EBG structure, that uses relatively thin and less expensive substrates, is needed for low frequency antenna applications. In this thesis, in addition to the basic mushroom-like EBG structures, fork-like and compact EBG structures are designed and developed.
3.3 DISPERSION ANALYSIS OF THE EBG STRUCTURES USING MULTICONDUCTOR TRANSMISSION LINE (MTL) THEORY

Many of the EBG structures used in microwave and RF engineering are designed by using transmission line circuits. In this chapter, the mushroom-like and fork-like EBG structures are considered for modelling and analysis, using the transmission line network theory and Bloch-Floquet’s Theorem. Using these analytical procedures, the characteristics equation or dispersion equation can be obtained, which provides information about the excited modes and their propagation constants. An infinite periodic structure can be viewed as a repetitive array of one such unit cell in the x and y directions. The dispersion equation can be obtained by applying the Bloch-Floquet theorem in both directions, thereby relating the voltages and current at one port to the voltages and current at another port, along the same axis.

3.3.1 Analysis of Mushroom-like EBG Structures

The unit cell of the 2-D periodic structure to be analyzed, is depicted in Figure 3.4. It consists of three distinct planes of metallization, the ground plane, a centre (square) mushroom-like EBG patch plane (conductor 2), and an upper shielding metal plane (conductor 1). Additionally, the centre of each mushroom-like EBG patch is connected to the ground with a via. As there are two planes of metallization above the ground, it is anticipated that the lowest order modes are quasi TEM, and may be captured by the MTL analysis.
‘p’ is the length and width of conductor 1 and the ground plane, m is the distance between conductor 1 and 2, n is the distance between conductor 2 and the ground plane, d is the length and width of the mushroom-like EBG structure, i.e., conductor 2.

An equivalent transmission line circuit that represents the unitcell as a cascade of passive elements, is introduced in Figure 3.5, with the following transfer matrix of the unit cell:

$$T_{\text{unitcell}} = T_{2C} T_{\text{MTL}(1)} T_L T_{\text{MTL}(1)} T_{2C}$$ \hspace{1cm} (3.5)

The matrix, $T_{2C}$, represents the equivalent series gap capacitance due to the fringing fields between patches. The value of the gap capacitance is doubled ($2C$) to account for the fact, that the overall structure is a cascade of unitcells. The matrix, $T_L$, represents the equivalent shunt inductance due to the via. The matrix, $T_{\text{MTL}(1)}$, represents the transfer matrix for a uniform section of the MTL (Faria 1993 and Paul 1994), of length, $l = p/2$. 
Figure 3.5 Equivalent MTL circuit for the mushroom-like EBG structure

The dispersion curves were calculated by computing the eigenvalues of $T_{\text{unitcell}}$. The unitcell of the structure to be analyzed, with the relevant dimensions and loading parameters are shown as

\begin{align*}
p &= 10 \text{mm} \\
m &= 12 \text{ mm} \\
d &= 6.7 \text{ mm} \\
n &= 1.67 \text{ mm} \\
\mu_0 &= 4\pi \times 10^{-7} \text{ H/m} \\
\varepsilon_{r1} &= 1 \\
\varepsilon_{r2} &= 4.4 \\
C &= 5.5 \times 10^{-12} \text{ F} \\
L &= 20 \times 10^{-9} \text{ H}
\end{align*}

The transfer matrices of the MTL sections are completely characterized by the per unit-length capacitance and inductance matrices, $C_{\text{MTL}}$, $L_{\text{MTL}}$ respectively. The $C_{\text{MTL}}$ and $L_{\text{MTL}}$ matrices are $2 \times 2$, as there are two conductors above the ground.

\begin{equation}
C_U = \frac{\varepsilon_r d}{m} \tag{3.6}
\end{equation}
\[ C_L = \frac{\varepsilon_n d}{n} \]  
(3.7)

\[ L_L = \frac{n \mu_0}{d} \]  
(3.8)

\[ L_H = \frac{m \mu_0}{d} \]  
(3.9)

\( C_{MTL} \) and \( L_{MTL} \) matrices are related to the per unit length shunt admittance and series impedance matrices, as shown in Equations (3.10) and (3.11)

\[ C_{MTL} = \begin{bmatrix} C_U & -C_U \\ -C_U & C_U + C_L \end{bmatrix} \]  
(3.10)

\[ L_{MTL} = \begin{bmatrix} L_U & -L_U \\ -L_U & L_U + L_L \end{bmatrix} \]  
(3.11)

where \( C_U, C_L, L_U, L_L \) are the components of the \( C_{MTL} \) and \( L_{MTL} \) matrices, consisting of the upper two conductors and lower two conductors on their own.

The \( Z \) and \( Y \) matrices used in the analysis of the MTLs are the same as those used in the coupled-mode analysis of the coupled transmission lines (Gupta et al 1996).

\[ Z = j\omega L_{MTL} \]  
(3.12)

\[ Y = j\omega C_{MTL} \]  
(3.13)

From Figure 3.5, it can be seen that the patch layer conductors are loaded with a series capacitor of the value \( 2C \) at the edges of the unitcell, and
a shunt inductor of the value $L$ on the ground at the centre of the unitcell.
Approximating this load as being continuous, the $Y$ and $Z$ matrices are
modified, as shown in Equations (3.14) and (3.15).

\[
Z_{\text{loaded}} = Z + \begin{bmatrix} 0 & 0 \\ 0 & 1/j\omega CP \end{bmatrix} \tag{3.14}
\]

\[
Y_{\text{loaded}} = Y + \begin{bmatrix} 0 & 0 \\ 0 & 1/j\omega CP \end{bmatrix} \tag{3.15}
\]

On substituting Equations (3.12) and (3.13) in Equations (3.14) and
(3.15), Equation (3.16) will be obtained.

\[
Z_{\text{loaded}} Y_{\text{loaded}} = \begin{bmatrix} L_U C_U & -L_U C_U + L_L \left( C_L - \frac{1}{j\omega^2 LP} \right) \\ C_U/C_{LP} & -C_U/C_{LP} + \left( C_L - \frac{1}{j\omega^2 LP} \right) \left( L_L - \frac{1}{j\omega^2 CP} \right) \end{bmatrix} \tag{3.16}
\]

On comparing Equations (3.16) and (3.17)

\[
Z_{\text{loaded}} Y_{\text{loaded}} = \begin{bmatrix} a_1 & -b_1 \\ b_2 & -a_2 \end{bmatrix} \tag{3.17}
\]

where $a_1 = \omega^2 (L_U C_U)$ \hfill (3.18)

\[
b_1 = \omega^2 \left( -L_U C_U + C_L - \frac{1}{j\omega^2 LP} \right) L_L \tag{3.19}
\]

\[
a_2 = -\omega^2 \left( -\frac{C_U}{j\omega^2 CP} + \left( C_L - \frac{1}{j\omega^2 LP} \right) \left( L_L - \frac{1}{j\omega 2CP} \right) \right) \tag{3.20}
\]

LP-Linear Polarization, CP-Cross Polarization
\[ b_2 = -\omega^2 \left( -\frac{C_v}{\omega^2 CP} \right) \] (3.21)

According to Bloch Floquet’s theorem,

\[
\begin{bmatrix}
C_n^- \\
\mathbf{c}_n^-
\end{bmatrix} = \begin{bmatrix}
a_1 & b_1 \\
b_2 & a_2
\end{bmatrix} \begin{bmatrix}
C_{n+1}^- \\
\mathbf{c}_{n+1}^-
\end{bmatrix}
\] (3.22)

\[
\begin{bmatrix}
C_{n+1}^+ \\
\mathbf{c}_{n+1}^+
\end{bmatrix} = e^{-\gamma \rho} \begin{bmatrix}
C_n^+ \\
\mathbf{c}_n^+
\end{bmatrix}
\] (3.23)

\[ a_1a_2 - b_1b_2 + e^{2\gamma \rho} - e^{-\gamma \rho} (a_1 + a_2) = 0 \] (3.24)

\[ a_1a_2 - b_1b_2 + \gamma^2 + \gamma^2 (-a_1 - a_2) = 0 \] (3.25)

\[ \gamma^4 + \gamma^2 (-a_1 - a_2) + a_1a_2 - b_1b_2 \] (3.26)

Equation (3.26) is the dispersion equation or characteristic equation of the geometry shown in Figure 3.5. In order to find the critical point frequencies, a slope has to be taken. The slope of the dispersion curve is zero at critical points. On differentiating Equation (3.26) with respect to \(\omega\), the Equation (3.27) will be obtained.

\[
\frac{d\gamma}{d\omega} = \frac{\gamma_2 \frac{d}{d\omega} (a_1 + a_2) - \frac{d}{d\omega} (a_1a_2 - b_1b_2)}{2\gamma \left[ 2\gamma^2 - (a_1 + a_2) \right]} \] (3.27)

the denominator of Equation (3.27) will become zero; when

\[ 2\gamma \left[ 2\gamma^2 - (a_1 + a_2) \right] = 0 \] (3.28)

\[ \gamma_2 = \frac{a_1 + a_2}{2} \] (3.29)
On substituting equation (3.29) into Equation (3.26), Equation (3.30) will be obtained.

\[
\text{Discriminant} = (a_1 + a_2)^2 - 4(a_1a_2 - b_1b_2) = 0
\]  

(3.30)

G and H are the roots of the equation (3.30)

The frequencies corresponding to Equation (3.30) will describe the critical points G and H. At this point, it will be illuminating if the roots of the dispersion Equation (3.26) are written. Equation (3.26) is quadratic in \(\gamma^2\) with the solutions given by

\[
\gamma^2 = \left(\frac{a_1 + a_2}{2}\pm\sqrt{(a_1 + a_2)^2 - 4(a_1a_2 - b_1b_2)}\right)
\]  

(3.31)

\[
\gamma^2 = \alpha + j\beta
\]  

(3.32)

where

\(\gamma\) is the propagation constant

\(\alpha\) is the attenuation constant expressed in nepers

\(\beta\) is the phase constant expressed in radians

On comparing Equation (3.31) and Equation (3.32), the following equations will be obtained:

\[
\alpha = \frac{(a_1 + a_2)}{2}
\]  

(3.33)

\[
\beta = \frac{\sqrt{(a_1 + a_2)^2 - 4(a_1a_2 - b_1b_2)}}{2}
\]  

(3.34)

It is observed that the frequencies for which G and H are satisfied, correspond to the zeroes of the discriminant of the dispersion Equation (3.26). After some algebraic simplifications, it may be written as Equation (3.35)
Discriminant = \( \omega_4 \left[ Q_1 + \frac{R_1}{\omega^2} + \frac{S_1}{\omega^3} + \frac{T_1}{\omega^4} - \frac{U_1}{\omega^5} \right] \) \hspace{1cm} (3.35)

where \( Q_1 = L_U C_U - L_L C_L \) \hspace{1cm} (3.36)

\( R_1 = (L_U C_U - L_L C_L) \left[ \frac{2L_U}{LP} - \frac{2}{LP} (C_U - C_L) \right] \) \hspace{1cm} (3.37)

\( S_1 = \frac{C_U^2}{(CP)^2} + \frac{2C_U C_L}{(CP)^2} + \frac{C_L^2}{(CP)^2} - \frac{2C_U L_L}{(CP)(LP)} + \frac{2C_L L_L}{(CP)(LP)} + \frac{4C_L L_L}{(CP)(LP)} + \frac{L_L^2}{(LP)^2} \) \hspace{1cm} (3.38)

\( T_1 = -\frac{2}{(CP)(CL)} \left[ \frac{C_U}{(CP)} + \frac{C_L}{(CP)} + \frac{L_L}{(LP)} \right] \) \hspace{1cm} (3.39)

\( U_1 = \frac{1}{(LP)^2(CP)^2} \) \hspace{1cm} (3.40)

The zeroes of Equation (3.35) will reduce to the solution of a quadratic equation in \( \omega^2 \), but when the condition \( \varepsilon_1 = \varepsilon_2 \) is satisfied it can be shown from the expressions in Equations (3.10) and (3.11), that \( L_U C_U - L_L C_L = 0 \). Thus \( Q_1 = R_1 = 0 \) and hence the Equation (3.35) becomes quadratic in \( \omega^2 \), as shown in Equation (3.41) with roots given in Equations (3.42) and (3.43).

Discriminant = \( S_1 + \frac{T_1}{\omega^2} + \frac{U_1}{\omega^4} \) \hspace{1cm} (3.41)

\( \omega_4^2 = \left( \frac{1}{(LP)(CP)} \right) \left( \frac{1}{\left( \frac{C_U}{(CP)} + \frac{C_L}{(CP)} + \frac{L_L}{(LP)} + \frac{2 \sqrt{C_U} \sqrt{L_L}}{\sqrt{2} \sqrt{LP} \sqrt{CP}} \right)} \right) \) \hspace{1cm} (3.42)
\[
\omega_H^2 = \frac{1}{(LP)(CP)} \left( \frac{1}{\frac{C_U}{(CP)} + \frac{C_L}{(CP)} + \frac{L_d}{(LP)} + \frac{\sqrt{C_U L_L}}{\sqrt{L_P CP}}} \right) 
\]

(3.43)

\(\omega_G\) and \(\omega_H\) are the frequencies corresponding to the critical points of G and H respectively. The dispersion diagram is obtained from the dispersion equation as shown in Figure 3.6. The critical points are labelled as G and H. It is easy to show that the denominator of Equation (3.42) is positive and hence \(\omega_G^2\) represents the larger of the two real positive, frequencies representing the solutions of Equations (3.42) and (3.43). It is observed that the width of the complex mode band may be controlled, for a given top and bottom conductor plane spacing, by varying the loading parameters, L and C. Conversely, if the load values L and C are fixed, one may vary the plate spacing to control the bandwidth. These expressions can be used as a simple aid in designing a structure with a desired band response. From Figure 3.6, it is clear that no eigen mode exists in the frequency range of 4.4 GHz to 5.2 GHz. Thus, this frequency region is defined as a surface wave bandgap. No surface waves can propagate in the EBG structure inside this frequency bandgap.
The MTL analysis is carried out for the mushroom-like EBG unitcell. From Figure 3.7 the band gap is noted as 2.2 GHz to 5.8 GHz. From Figure 3.7 it is shown, that no eigen mode exists in the frequency range of 4.6 GHz to 5.15 GHz. Thus, frequency region is defined as a surface wave bandgap. Surface waves cannot propagate in this region, when a fork-like EBG structure is inside. From the diagrams it is clear that the bandgap region of the mushroom-like EBG structure is larger than that of the fork-like EBG structure.

The simulated dispersion characteristic for the mushroom-like EBG, using CST microwave studio is shown in Figure 3.7. For mushroom-like EBG structures, the stop band between mode 1 and mode 2 is noted to be 3.7 GHz to 5.2 GHz. Therefore, the surface wave bandgap obtained from the dispersion diagram is from 3.7 GHz to 5.2 GHz.
Compactness is always of importance in wireless communications. In this chapter, a novel compact fork-like EBG design is investigated. It is known that slots on the microstrip antenna help to reduce to the antenna size, because they change the current distribution on the patch antenna (Zhang and Yang 1998). This idea is also applied to the fork-like EBG design. An extension of Sievenpiper’s EBG structure is the fork-like EBG. The structure of the fork-like metal patch is comparatively easier than other shapes of EBGs, and it is more compact. It significantly increases the capacitance between the neighbouring elements, because of the slots and stretches present in it. The lattice period of the fork-like EBG is $0.04 \lambda_0$. Size of the fork-like EBG is reduced to 5% of the free space wavelength. The patch is etched in such a way, that it has a fork-like shape. There is only one microstrip line at the bottom of the etched patch, which is placed at the centre. Increasing the length of the microstrip will increase the capacitance. It will also work at a lower frequency than Sievenpiper’s EBG structure, which is usually larger.

Figure 3.7 Dispersion diagram of mushroom-like EBG

3.3.2 Analysis of Fork-like EBG Structures
The unit cell of the 2-D periodic structure to be analyzed is depicted in Figure 3.8. It consists of three distinct planes of metallization, the ground plane, a centre (square) fork-like EBG patch plane (conductor 2), and an upper shielding metal plane (conductor 1). Additionally, the centre of each fork-like EBG patch is connected to the ground with a via. As there are two planes of metallization above the ground, it is anticipated that the lower order modes are quasi TEM, and may be captured by the MTL analysis.

![Figure 3.8 Two-D unitcell of the fork-like EBG structure](image)

‘p’ is the length and width of the conductor 1 and ground plane, m is the distance between conductors 1 and 2, n is the distance between the conductor 2 and the ground plane, d is the length and width of the fork-like EBG structure, i.e., conductor 2.

The dispersion curves were calculated by computing the eigen values of $T_{\text{unitcell}}$. The unitcell of the structure to be analyzed, with the relevant dimensions and loading parameters are shown as

\[ p = 8 \text{ mm} \]
m = 12 mm

\[ d = 4.8 \text{ mm} \]

n = 1.67 mm

\[ \mu_0 = 4\pi \times 10^{-7} \text{ H/m} \]

\[ \varepsilon_{\varepsilon_1} = 1 \]

\[ \varepsilon_{\varepsilon_2} = 4.4 \]

C = 8.2 \times 10^{-12} \text{ F}

L = 14 \times 10^{-9} \text{ H}

Figure 3.9 (a) shows the details of a partial arrangement of the unitcell to form a fork-like EBG structure, and Figure 3.9 (b) one unit of the fork-like EBG lattice respectively. The square patch size (a) is 4.8 mm x 4.8 mm, the gap between neighboring elements (g) is 0.5 mm, the width of the microstrip line stretched (s) is 1.25 mm, and the slot etched at 1 mm each. The radius of a via (r) is 0.4 mm.

(a) Partial arrangement of Fork-like EBG structures

(b) Dimension of one unit cell of the Fork-like EBG structure
The operative mechanism of the EBG structure can be explained by an LC filter array. From Equation (3.1), in order to achieve an even more compact EBG structure, the equivalent capacitance $C$ and inductance $L$ should be increased. From Equation (3.2), if the dielectric material and its thickness have been chosen properly, the inductance $L$ cannot be altered. Therefore, only the capacitance $C$ can be increased. The equivalent LC circuit acts as a two dimensional electric filter in this frequency range, to block the propagation of the surface waves. The inductor $L$ results from the current flowing through the via, and the capacitor $C$, due to the gap effect between the adjacent patches. Thus, the approach to increase the inductance or capacitance will naturally result in the decrease of the band-gap position.

![Image](image_url)

**Figure 3.10** $\alpha$-$\beta$ diagram of the Fork-like EBG from MTL theory

The simulated $\alpha$-$\beta$ characteristic for the fork-like EBG, using a CST microwave studio is shown in Figure 3.10. The simulated dispersion diagram of the fork-like EBG is shown in Figure 3.11. From Figure 3.11 it is seen that the band gap exists from 2 GHz to 5.2 GHz. For fork-like EBG structures, the stop band between modes 1 and 2 is noted as from 4.2 GHz to 5.2 GHz. Therefore, the surface wave bandgap obtained from the dispersion diagram is 4.2 GHz to 5.2 GHz.
At low frequencies, the TE surface waves are cutoff, since the surface impedance of the periodic structure is inductive, due to the coils on the grounded dielectric substrate, and therefore, the TM surface wave is supported. This supported wave exists predominantly above the surface of the periodic structure, and has a propagation constant very close to that of free space. Therefore, as the frequency is increased, the propagation constant of the TM surface wave follows the light line as shown in Figures 3.7 and 3.11. Also, the surface impedance becomes capacitive, and the TE surface modes are supported. Since the E-field has no vertical component for the TE modes, this mode remains largely unaffected by the presence of the vertical coils. Therefore, the fundamental TE mode is essentially the TE mode supported by the structure without the vertical conducting coils. When the backward wave intersects the light line, contra directional coupling occurs between the backward wave and the forward travelling surface waves. This contra directional coupling causes the backward and forward TM modes to form a stop band. Above and below the stopband, the backward wave and the TM surface wave join to form continuous bands.

Figure 3.11 Dispersion diagram of the fork-like EBG structure
3.3.3 Analysis of Compact Dual Band EBG Structures

Compactness and miniaturization are always of importance in wireless communications. In this chapter, a novel compact dual band EBG design is investigated. Figure 3.12 shows the compact dual band EBG structure with unit cell dimensions. It is well known, that slots on the microstrip antenna help to reduce to the antenna size, because they change the current distribution on the patch antenna (Zhang and Yang 1998). The same idea is also applied to the compact dual band EBG design. An attempt is made here to design and implement a compact dual band EBG structure. Instead of operating EBG structures for a single band of operation, an extension of the dual band operation is made here. Even though the structure of the fork-like metal patch is comparatively easier than other shapes of EBGs, and it is more compact, it significantly increases the capacitance between the neighbouring element, because of the slots and stretches present in it. The fork-like EBG can operate for a single band. The lattice period of the fork-like EBG is $0.144 \lambda_0$. The size is reduced to 3 % of the free space wavelength. The patch is etched in such a way that it has two slots, which are opposite to each other. Increasing the length of the slot will increase the capacitance. It will also work at a lower frequency than Sievenpiper’s EBG structure, which is usually larger. The band gap characteristics are obtained using the suspended strip line method. It is obtained as 3 GHz to 3.36 GHz and 4.5 GHz to 5.1 GHz.
3.4 GEOMETRY OF THE INSET FEED MICROSTRIP PATCH ANTENNA AND ANTENNA ARRAY INCORPORATED WITH EBG STRUCTURES

From the previous analysis of the EBG structures, the size and shape are characterized. The designed and developed EBG structures are incorporated with the microstrip antenna and antenna array configurations. The compactness in the EBG structures as well antenna structures are obtained by placing the EBG structures near the inset feedline of the microstrip antenna. For the optimal placement of the EBG structures near the feedline, a parametric study has been done. This is checked for the microstrip patch antenna before incorporating it in the microstrip patch antenna array corporate feed. The intent of this thesis is to improve the performance of the antenna array with two different feed techniques, by reducing surface wave propagation, spurious feed radiation, and feed network loss. Initially, the microstrip antenna array is constructed using the inset feed method, where a simple microstrip line is given as a feed line. This is a simple structure and it can be fabricated along with the patch on the same layer of the board. In the proposed antenna designs, the results are compared with the placement of the EBG structures, around the antenna and near the feedline of the antenna. Also, the antenna array is designed to operate at dual band of frequencies.
3.4.1 Parametric Study on Placing the EBG Structures near the Feedline

The parametric study is done mainly to determine the optimal place at which the EBG structures can be placed near the inset feedline of the microstrip antenna. The mushroom-like EBG structures are innovatively positioned on either sides of the inset feedline of the microstrip antenna. The band gap characteristics of the EBG structures are to forbid the propagation of surface waves in certain range frequency bands. In this design, the EBG band gap is 4.4 GHz to 5.2 GHz, and allows the band in this range to radiate effectively at its own frequencies with full power. When the EBG structures are placed near the feedline, the spurious radiations from the feed line due to impedance mismatch can be absorbed, and controlled radiation takes places at the operating frequency. For placing the EBG structures near the feedline, a parametric study is done. Figure 3.11 depicts the $S_{11}$ (dB) versus frequency (GHz) obtained from different placements of the EBG structure at distances of 1 mm, 1.5 mm, 2 mm and 2.5 mm from the inset feedline.

![Figure 3.13](image)

**Figure 3.13** Parametric study of placing the EBG structures near the feedline
Table 3.1 shows the $S_{11}$ (dB) of the single patch antenna with mushroom-like EBG structures on both sides of the inset feedline at various distances (d mm). It is observed from Figure 3.13 that, i) when the EBG structures are placed at 1 mm ($0.016 \lambda_o$) distance from the feedline, it results in a return loss value of 16.7 dB, ii) when the EBG structures are placed at a distance of 1.5 mm ($0.025 \lambda_o$) from the feedline, it results in a return loss value of 28 dB, iii) when the EBG structures are placed at 2 mm ($0.032 \lambda_o$) from the feedline, it results in a return loss value of 13.56 dB, and iv) when the EBG structures are placed at 3 mm ($0.048 \lambda_o$) from the feedline, it results in a return loss value of 10.41 dB. It is obvious from the above results that the optimal distance of the EBG structures from the feedline is 1.5 mm. This is because of the non interaction of the coupling of the TM modes from the coaxial probe and the EBG structures. Hence, the band gap characteristics of the EBG structures enhance the antenna’s performance.

When the EBG structures are at a distance of less than 1.5 mm from the feedline, the surface waves and feedline impedance matching, affect the performance of the antenna. In this, the EBG structures interact mainly with the cylindrical TM surface waves excited by the coaxial probe feed. This degrades the antenna’s performance as well. When the EBG structures are placed at a distance of more than 1.5 mm from the feedline, it suppresses the surface waves, but the feedline radiation is not controlled. This reduces the return loss and affects the antenna radiation patterns. The feed radiation also leads to undesired cross polarized radiation.

With respect to the patch element the uppermost EBG structures should be placed at a minimum of 0.048 $\lambda_o$ (3 mm) from the lower edge of the patch. This is established to be the most suitable distance, as it avoids the fringing field effects, and does not affect the radiation from the patch.
Table 3.1 Parametric values of $|S_{11}|$ (dB) at different positions of EBG structures from the feedline

| Sl. No. | EBG distance from feed line ($\lambda_o$) ‘$d_{mm}$’ | $|S_{11}|$ (dB) | Resonant Frequency (GHz) |
|---------|---------------------------------|----------------|-------------------------|
| 1       | 0.016                           | -16.7          | 4.635                   |
| 2       | 0.024                           | -28            | 4.78                    |
| 3       | 0.032                           | -13.56         | 4.68                    |
| 4       | 0.048                           | -10.41         | 4.63                    |

The next analysis is done for comparing the performance of the microstrip antenna surrounded by the EBG structures, and the microstrip antenna with the EBG structures near the inset feedline. The comparison is carried out for the mushroom-like EBG structures and Fork-like EBG structures. Figure 3.14 shows the graph between $S_{11}$ (dB) versus frequency (GHz) of two different EBG structures, namely, the mushroom-like EBG and the fork-like EBG. From the graph, the $S_{11}$ values of the microstrip antenna with feedline EBG structures are 1) mushroom-like EBG is 28 dB and 2) fork-like EBG is 37.2 dB. The $S_{11}$ value of the microstrip antenna surrounded by the EBG structures are 1) mushroom-like EBG structures is 16 dB and 2) fork-like EBG structure is 20 dB. It is also noticed that the bandwidth is significantly improved when the EBG structures are placed at the feedline of the antenna. From these results, the conclusion made in this thesis, is to place the EBG structures near the inset feedline of the antenna.
Figure 3.14 Comparison of the results with the EBG structures near the feedline and around the patch

3.4.2 Microstrip Patch Antenna with Mushroom-like EBG Structures

Chapter 2 detailed the rectangular microstrip patch antenna configuration; it has been selected here, to perform the comparison between the antenna, with and without EBG structures at the feedline. Figure 3.15 (a) and (b) shows the microstrip antenna with the conventional placing of the mushroom-like EBG structures, and the feedline placement of the mushroom-like EBG structures. The usefulness of placing the EBG structures on both sides of the feedline, is considered with a rectangular microstrip antenna, with a dimension of 19 mm x 14.4 mm. The antenna is designed to operate at 4.8 GHz. Initially, the approach is verified by placing the mushroom-like EBG structure and fork-like EBG structure. Even though the fork-like EBG shows a better result, the mushroom-like EBG has a wider bandwidth compared to the fork-like EBG structure. Hence, mushroom-like EBG structures are placed and analyzed for their performance. A prototype of the single microstrip antenna operating at single frequency was implemented and fabricated on an FR-4 substrate, of a dielectric constant of 4.4 and thickness
of 1.6 mm. A column of 1 x 3 mushroom-like EBG structures and 1 x 4 fork-like EBGs are placed on both sides of the inset feedline of the microstrip antenna, and are investigated separately. The microstrip antenna is fed via a coaxial line probe, and the excitation is given at the microstrip line, which is connected with the patch design. The dimension of the unit cell mushroom-like EBG and fork-like EBG are shown in Figures 3.9 and 3.10. The size of the ground plane and the substrate is 30 mm x 30 mm.

3.4.3 Two Element Microstrip Patch Antenna Array with Mushroom-like EBG Structures

While the previous chapter discussed the configuration of the microstrip patch antenna array, the design of the microstrip patch antenna array with mushroom-like EBG structures near the corporate feedline, is now presented in Figure 3.16. The ground plane lies at the bottom side of the dielectric substrate, and its size is 120 mm x 120 mm. The dielectric constant
of the substrate is 4.4 and it has a thickness of 1.6 mm. In this antenna array configuration, the mushroom-like EBG structures are placed in an aperiodic arrangement. The main focus here is to reduce the surface wave propagation, spurious feed radiation, and feed network loss. The spurious radiation that occurs in the feed network, is not very crucial if the antenna array size is small. This radiation becomes significant if the antenna array becomes larger. Hence, this is an important factor to be considered while constructing the antenna array.

**Figure 3.16** Geometry of a two element microstrip patch antenna array with mushroom-like EBG structures at the corporate feed

### 3.4.4 Four Element Microstrip Patch Antenna Array with Mushroom-like EBG Structures

While the previous chapter discussed the configuration of the microstrip patch antenna array without EBG structures, the design of a four element microstrip patch antenna array with mushroom-like EBG structures near the corporate feedline, is now presented in Figure 3.17. The ground plane lies at the bottom side of the dielectric substrate and its size is 160 mm x 160 mm. The dielectric constant of the substrate is 4.4 and it has a thickness of 1.6 mm. In this antenna array configuration, the mushroom-like
EBG structures are placed in an aperiodic arrangement. The main focus here is to reduce the surface wave propagation, spurious feed radiation, and feed network loss. The spurious radiation that occurs in the feed network is not very crucial if the antenna array size is small. This radiation becomes significant if the antenna array becomes larger. Hence, this is an important factor to be considered while constructing the antenna array.

Figure 3.17 Geometry of a four element microstrip patch antenna array with mushroom-like EBG structures at the corporate feed

3.4.5 Dual Frequency-Dual Polarized Microstrip Patch Antenna Array with Mushroom-like EBG Structures

In chapter 2, the design of a dual frequency-dual polarized microstrip antenna array was detailed. In this chapter, the dual frequency-dual polarized microstrip patch antenna array with mushroom-like EBG structures for improved performance, is discussed. The dual band microstrip patch antenna array with mushroom-like EBG structures is shown in Figure 3.18. The antenna is designed to operate at 3.12 GHz and 4.8 GHz. The radiating patches are tilted individually at +45° and −45°, for achieving dual polarization. To enhance the performance in these two bands, mushroom-like
EBG structures of two different sizes are placed. To suppress the surface waves at 3.12 GHz, mushroom-like EBG of 12 mm x 12 mm dimension and a via radius of 0.6 mm are incorporated. In order to suppress the surface waves at 4.8 GHz, a mushroom-like EBG 6.7 mm x 6.7 mm, and a via radius of 0.4 mm is implemented. The gap between the EBG cells is 0.5 mm. The ground plane size is 60 mm x 60 mm. Both the EBG structures effectively reduce the surface waves, and increase the co-polarization effects by reducing cross-polarization. But, placing more EBG cells is not possible, because it occupies more space.

![Figure 3.18 Geometry of a dual frequency-dual polarized microstrip patch antenna array with mushroom-like EBG structures](image)

**Figure 3.18** Geometry of a dual frequency-dual polarized microstrip patch antenna array with mushroom-like EBG structures

### 3.4.6 Dual Frequency-Dual Polarized Microstrip Patch Antenna Array with Compact Dual Band EBG Structures

The previous chapter detailed the design and configuration of a dual frequency-dual polarized two element microstrip patch antenna array. The effect of tilting separately the radiating patches + 45° and − 45° in microstrip antenna array structures, yields drastic effects in polarization. Each radiating patch is given dual resonance in order to achieve dual polarization. From this design, linear polarization and orthogonal polarization can be achieved. The dual band-dual polarized microstrip patch antenna
array, with and without dual band compact EBG structures, is shown in Figure 3.19. The microstrip antenna array is designed to operate at 3.12 GHz and 4.8 GHz. The dual band EBG structure has the ability to suppress the surface waves at both the frequencies. This is because of the slits present in the EBG structure. The two element microstrip antenna array is designed tilted at $+45^\circ$ and $-45^\circ$ separately, in order to achieve dual polarisation. The EBG structures are placed near the corporate feedline, thereby compactness in the antenna array design is achieved. The ground plane size is 60 mm x 60 mm. Both the antenna array designs are simulated, using a dielectric substrate having a dielectric constant of 4.4 and a thickness of 1.6 mm. The excitation of feed is given at the edge of the board, using a coaxial probe SMA connector.

![Figure 3.19 Geometry of a dual frequency-dual polarized microstrip patch antenna array with compact dual band EBG structures](image)

### 3.5 APERTURE COUPLED MICROSTRIP PATCH ANTENNA ARRAY WITH FREQUENCY SELECTIVE SURFACE AS A SUPERSTRATE

The detailed analysis and design procedure of an aperture coupled antenna and antenna array were discussed in chapter 2. The performance and directivity of the aperture coupled microstrip antenna array can be improved, by placing the FSS as a superstrate. The FSS acts like a partial reflector. Near its resonance frequency where the reflection coefficient of the surface is
unity, the radiating source and the FSS superstrate layer produce significant improvement in its performance. The aperture coupled microstrip antenna operates at 3.12 GHz and 4.8 GHz. The analysis and placing of the FSS as a superstrate is explained here. The placement of the FSS and its dimensions are obtained through the parametric analysis. The FSS is placed as a superstrate layer and verified for its performance in the aperture coupled microstrip antenna of i) a single element, ii) two element and iii) four element arrays. By using the FSS, the radiation characteristics such as back radiation and side lobe level can be controlled effectively, for the specified band of frequencies at which the antenna array is designed to operate. One of the most important challenges in the design of the microstrip antenna array is achieving greater directivity and better radiation characteristics.

3.5.1 Parametric Analysis of the FSS as a Superstrate

The effects of the dimensions and height of the FSS from the radiating element, are studied in the parametric analysis. By varying the height of the FSS layer from the radiating patch, the band gap of the FSS is analyzed. The FSS layer is designed to eliminate the surface waves at the operating frequency of 4.8 GHz. Also, it eliminates the surface waves at the lower harmonics obtained at 3.12 GHz. By adjusting the height of the FSS layer different return loss values are obtained. For the same behavior, the dimensions of the FSS metal patches are modified. The square patches are designed to resonate at 3.12 GHz and 4.8 GHz. Hereby, the dimension of the FSS layer is studied by varying the number of cells. Increasing the dimensions of the FSS layer eventually increases the radiating aperture of the antenna design. The effects of the height of the FSS layer on the return loss of the antenna are shown for different heights.
The dimensions of the FSS layer, such as the height and unit cell dimensions are obtained using the parametric results. By varying the distance (d) between the array and the FSS layer, the resonance frequency of the antenna is noted. Figure 3.20 shows the arrangement of the FSS layer as a superstrate. The elements of the FSS are modified according to the return loss value of the patch and FSS layer. The dimensions of the metal patches in the FSS are varied, based on the resonance value. The appropriate dimension of the metal patches and the height at which the FSS can be placed are obtained. Figure 3.21 depicts the effects of the height of the FSS layer on the return loss vs frequency graph. The antenna is simulated with optimal FSS layer height for better results. The optimum height of 17.5 mm from the radiating patch is measured from the parametric studies. The dimension of the metal patch is 13 mm x 11 mm. Therefore, the dimension of the FSS layer must be chosen such that, the reflected wave from the edges of the FSS can be neglected. Table 3.2 shows the parametric study results for different heights of the FSS from the radiating patch.
Table 3.2 Return loss in dB for various heights and dimensions of the FSS

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>d (mm)</th>
<th>Return Loss(dB)</th>
<th>$f_1$(3.12GHz)</th>
<th>$f_2$(4.8GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>14</td>
<td>-6</td>
<td>-5.4</td>
<td></td>
</tr>
<tr>
<td>2.</td>
<td>15</td>
<td>-11</td>
<td>-12.2</td>
<td></td>
</tr>
<tr>
<td>3.</td>
<td>16</td>
<td>-13</td>
<td>-13.5</td>
<td></td>
</tr>
<tr>
<td>4.</td>
<td>17.5</td>
<td>-18.8</td>
<td>-21.6</td>
<td></td>
</tr>
<tr>
<td>5.</td>
<td>18.5</td>
<td>-15.4</td>
<td>-14</td>
<td></td>
</tr>
<tr>
<td>6.</td>
<td>19.5</td>
<td>-18.4</td>
<td>-13.1</td>
<td></td>
</tr>
</tbody>
</table>

3.5.2 Aperture Coupled Microstrip Patch Antenna with FSS

The geometry of the aperture coupled microstrip patch antenna without the FSS was discussed in chapter 2. The implementation of the FSS layer as a superstrate in the aperture coupled microstrip patch antenna configuration is shown in Figure 3.22. The bottom most layer is the feedline, above which a dielectric material with a dielectric constant of 4.4 is placed. An U- shaped aperture coupled slot is placed above this dielectric layer. This
aperture coupled slot is designed to resonate at dual frequencies. A dielectric substrate is placed in between the radiating patch and the aperture slot with a dielectric constant of 4.4. The top most layer in the antenna design consists of a metal FSS layer, which is placed at a height of 17.5 mm from the radiating patch. The radiating patch has a dimension of 19 mm x 14 mm. Both the dielectric materials have a dielectric constant of 4.4 and thickness of 1.6 mm. The aperture coupled antenna designed with the FSS operates at 3.12 GHz and 4.8 GHz. The geometry of the arrangement of all the antenna layers is shown in Figure 3.22.

![Figure 3.22 Geometry of the aperture coupled microstrip patch antenna with FSS](image)

### 3.5.3 Two Element Aperture Coupled Microstrip Patch Antenna Array with FSS

The previous chapter have defined the configuration of the aperture coupled two element microstrip patch antenna array without the FSS. The geometry of the aperture coupled two element microstrip patch antenna array with the FSS is shown in Figure 3.23. The feed network is very important in an antenna array configuration. Corporate feed is used to cascade the array element, with appropriate quarter wave transformers. The feedline is given in the bottom most layer. A dielectric substrate is sandwiched between the
feedline and an U-shaped aperture slot. Another dielectric substrate is placed on top of the aperture slot, and the radiating patches are kept above this dielectric material. The FSS layer is placed at a height of 17.5 mm from the radiating patches. The dimension of the radiating patch is 19 mm x 14 mm. The dielectric material having a dielectric constant of 4.4 and a thickness of 1.6 mm is used. The geometry of the aperture coupled microstrip antenna array with the FSS is shown in Figure 3.23.

![Figure 3.23 Geometry of the two element aperture coupled microstrip patch antenna array with FSS](image)

3.5.4 Four Element Aperture Coupled Microstrip Patch Antenna Array with FSS

The configuration of the aperture coupled four element microstrip patch antenna array without the FSS was discussed in chapter 2. The geometry of the aperture coupled four element microstrip antenna array with the FSS is shown in Figure 3.24. The feed network is very important in the antenna array configuration. A corporate feed is used to cascade the array element, with appropriate quarter wave transformers. The corporate feed network is given in the bottom most layer. A dielectric substrate is
sandwiched between the feedline and an U-shaped aperture slot. Another
dielectric substrate is placed on top of the aperture slot, and the radiating
patches are kept above this dielectric material. FSS layer is placed at a height
of 17.5 mm from the radiating patches. The dimension of all the four radiating
patches is 19 mm x 14 mm each. Dielectric material having a dielectric
constant of 4.4 and thickness of 1.6 mm is used. The geometry of the aperture
coupled microstrip antenna array with FSS is shown in Figure 3.24.

![Figure 3.24 Geometry of the four element aperture coupled microstrip
patch antenna array with FSS](image)

3.6 RESULTS AND DISCUSSIONS

In order to validate the simulated results, prototypes of the
microstrip antennas with EBG and FSS were fabricated. The return loss was
measured using an Agilent 8510B Vector Network Analyzer (VNA) and the
radiation patterns were obtained in the anechoic chamber which is 8m x 4m x
4m ferrite chamber constructed by Siepel has the frequency range from 800
MHz to 18 GHz and this anechoic chamber is validated by the CTIA (Cellular
Telecommunication and Internet Association).
3.6.1 Single Frequency Microstrip Patch Antenna Incorporated with EBG Structures in the Conventional Placing and at the Inset Feedline

The main focus of this thesis is to find the optimal placing of the EBG structure in such a way, that it gives performance enhancement and compactness. Placing more number of EBG structures also causes parasitic effects and losses. Hence it is essential to maintain with minimal the number of EBG cells. Initially, a parametric study is done by placing the EBG structures around the antenna and at the inset feedline of the antenna. The Simulated $S_{11}$ of the rectangular single element microstrip antenna is shown in Figure 2.13 in chapter 2. The simulated $S_{11}$ of the microstrip antenna surrounded by the mushroom-like EBG and fork-like EBG, the placement of the mushroom-like EBG and fork-like EBG at the feedline are shown in Figure 3.25. The resonant frequencies of the mushroom-like and fork-like EBGs around the antenna are located at 4.62 GHz and 4.7 GHz respectively, and their corresponding return losses are -16 dB and -20 dB respectively. The resonant frequencies of the mushroom-like and fork-like EBGs at the inset feedline are located at about 4.76 GHz and 4.68 GHz respectively, and their corresponding return losses are -28.17 dB and -37 dB respectively. The bandwidth of the mushroom-like EBG structure is greater than that of the fork-like EBG structure. The $S_{11}$ of the microstrip antenna with EBG structures near the feedline shows considerable improvement. Therefore, EBG structures when placed near the feedline, will lead to greater suppression of the surface waves, by increasing the antenna efficiency. Also, the bandgap characteristics of the mushroom-like EBG are wider, which will lead to a significant reduction of the surface waves in this bandgap, compared to the fork-like EBG. Therefore, the mushroom-like EBG structures to be placed near the feedline, are proposed for enhanced antenna performance and size reduction.
Figure 3.25 Simulated $S_{11}$ (dB) of the microstrip patch antenna surrounded by mushroom-like and fork-like EBG structures placed at the feedline

The simulated $S_{11}$ of the mushroom-like and fork-like EBGs placed near the inset feedline of the microstrip patch antenna are shown Figure 3.26. The simulated return loss of the microstrip patch antenna with the mushroom-like EBG is obtained as -28.17 dB. The measured return loss of the microstrip patch antenna with the mushroom-like EBG is obtained as -25.4 dB. The corresponding resonant frequencies are 4.76 GHz and 4.81 GHz. In simulation, the antenna is matched to ($|S_{11}| < -10$ dB) with an impedance bandwidth of 160 MHz from the cut off frequencies 4.66 GHz to 4.82 GHz, which represents the fractional bandwidth of 3.36 %. In measurement, the antenna is matched to ($|S_{11}| < -10$ dB) with an impedance bandwidth of 120 MHz obtained from the cut off frequencies 4.74 GHz to 4.86 GHz, which represents the fractional bandwidth of 2.5 %. The simulated return loss value of the antenna with the fork-like EBG is obtained as 37 dB, and its corresponding resonant frequency is 4.68 GHz. The antenna is matched to ($|S_{11}| < -10$ dB) with an impedance bandwidth of 110 MHz obtained from the cut off frequencies of 4.62 GHz to 4.73 GHz, which represents the fractional
bandwidth of 2.36%. The impedance bandwidth is significantly greater in the mushroom-like EBG than in the fork-like EBG. Compared with the simulated result, the measured result shows good agreement within the operating band. The measured result shows a larger bandwidth and the resonant frequency shifts to a higher frequency. This is by minor differences between the practical and simulated models for the FR-4 substrate. In addition, the dielectric constant and dissipation factor are not stable when the frequency changes. In general, good agreement is observed between the measured and simulated results. Figure 3.27 shows the photograph of the fabricated microstrip patch antenna with mushroom-like EBG structures at the feedline, and the measurement setup of the microstrip patch antenna, using the Vector Network Analyzer. The measured and simulated resonant frequency, return loss, bandwidth, VSWR and directivity are given in Table 3.3.

![Simulated and measured S11 (dB) of the mushroom-like and fork-like EBG structures](image)

**Figure 3.26** Simulated and measured $S_{11}$ (dB) of the mushroom-like and fork-like EBG structures
3.6.1.1 Radiation pattern

The simulated and measured radiation patterns of the single frequency microstrip antenna, with and without mushroom-like EBG, at the inset feedline are indicated in Figure 3.28 (a) and (b). The agreement between the results is quite good.

Figure 3.28 (a) and (b) Simulated and measured radiation pattern of the single microstrip patch antenna without and with mushroom-like EBG structures at the feedline
The simulated and measured directivity of the single frequency microstrip patch antenna without mushroom-like EBG is 7.1 dBi and 6.3 dBi. The simulated and measured directivity of the single microstrip patch antenna with mushroom-like EBG is 9 dBi and 7.1 dBi. It is clearly shown from the figures that the directivity is improved when the EBG structures are incorporated. The simulated and measured results are in good agreement with each other. The VSWR of the microstrip antenna with mushroom-like EBG is 1.08 and 1.10. The microstrip antenna with mushroom-like EBG satisfies the requirements of indoor wireless applications. From the VSWR values, it is understood that the increased diversity gain also counter-acts the channel fading caused by multipath propagation. The simulated and measured return loss, VSWR and directivity of the microstrip antenna with mushroom-like EBG at the feedline, are given in Table 3.3.

Table 3.3 Simulated and measured results of the microstrip patch antenna with EBG structures

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna Parameters</th>
<th>With mushroom-like EBG</th>
<th>With Fork-like EBG</th>
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<td></td>
<td></td>
<td>Simulated</td>
<td>Measured</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequency(GHz)</td>
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<td>4.81</td>
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<td>2.</td>
<td>Return loss(dB)</td>
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<td>-26.56</td>
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<td>3.</td>
<td>Bandwidth (MHz)</td>
<td>120.00</td>
<td>125</td>
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<td>4.</td>
<td>VSWR</td>
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<td>1.10</td>
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<tr>
<td>5.</td>
<td>Directivity(dBi)</td>
<td>9</td>
<td>7.1</td>
</tr>
</tbody>
</table>
3.6.2 Two Element Microstrip Patch Antenna Array with Mushroom-like EBG Structures

The simulated and measured $S_{11}$ of the two element microstrip patch antenna array was discussed in chapter 2. Figure 3.29 shows the simulated and measured results of two element microstrip patch antenna arrays, with the mushroom-like EBG structure near the corporate feedline. The antenna array with the EBG structure modifies the impedance matching of the design. The measured return loss results of the antenna array with mushroom-like EBG match the ($|S_{11}| < -10$) dB with an impedance bandwidth of 190 MHz, obtained from the cut off frequencies of 4.65 GHz to 4.84 GHz, which represents a fractional bandwidth of 4.02 % at the resonant frequency of 4.72 GHz. The simulated return loss result matched the ($|S_{11}| < -10$) dB with the impedance bandwidth of 140 MHz obtained from the cut off frequencies 4.61 GHz to 4.75 GHz, which represents a fractional bandwidth of 3 % at the resonant frequency of 4.67 GHz. From the simulated and measured results, the maximum return loss value is obtained as -19 dB and -24.7 dB. Also, the simulated return loss value of the fork-like EBG matched the ($|S_{11}| < -10$ dB) with the impedance bandwidth of 120 MHz from the cut off frequencies of 4.6 GHz and 4.72 GHz, which represents a fractional bandwidth of 2.6 % at the resonant frequency of 4.65 GHz. The peak return loss values are obtained as 17.6 dB. From the results shown in Figure 3.29, it can be observed, that there is a slight shift in the resonant frequency between the simulated and measured results. This is due to fabrication errors. Figure 3.30 shows the photograph of the fabricated two element microstrip patch antenna array with mushroom-like EBG at the corporate feed network, and the measurement setup of the antenna using the Vector Network Analyzer.
3.6.2.1 Radiation pattern

The simulated and measured radiation pattern of the two element microstrip antenna array with mushroom-like EBG structures placed near the corporate feedline at 4.8 GHz are indicated in Figure 3.31 (a) and (b). There
are some distortions in the measured radiation pattern, which are attributed to the feed cable and connector. The simulated and measured radiation pattern for the antenna array without EBG structures is obtained as 13.03 dBi and 12.11 dBi. The simulated and measured radiation pattern for the antenna array with the EBG structures is obtained as 16.19 dBi and 14.95 dBi. In Figure 3.31(a) the simulated radiation pattern has a grating lobe and side lobes, which reduce the efficiency of the antenna array. In Figure 3.31(b) the simulated radiation pattern shows only the main lobe and a small back lobe. This is due to the placement of the mushroom-like EBG structures at the corporate feedline. The simulated and experimented results are in good concurrence with each other. The measured and simulated resonant frequency, return loss, VSWR and directivity are given in Table-3.4.

Figure 3.31  (a) and (b) Simulated and measured radiation pattern of the two element microstrip patch antenna array with mushroom-like EBG structures
Table 3.4  Simulated and measured results of the two element microstrip patch antenna array with EBG structures

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna parameters</th>
<th>With Mushroom-like EBG</th>
<th>With Fork-like EBG</th>
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<tr>
<td></td>
<td>Simulated</td>
<td>Measured</td>
<td>Simulated</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequency(GHz)</td>
<td>4.67</td>
<td>4.72</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss(dB)</td>
<td>-18.61</td>
<td>-24.33</td>
</tr>
<tr>
<td>3.</td>
<td>Bandwidth (MHz)</td>
<td>150.00</td>
<td>164.21</td>
</tr>
<tr>
<td>4.</td>
<td>VSWR</td>
<td>1.27</td>
<td>1.13</td>
</tr>
<tr>
<td>5.</td>
<td>Directivity(dBi)</td>
<td>16.19</td>
<td>14.95</td>
</tr>
</tbody>
</table>

3.6.3  Four Element Microstrip Patch Antenna Array with Mushroom-like EBG Structures

The simulated $S_{11}(dB)$ of the four element microstrip patch antenna array with mushroom-like EBG structure at the corporate feed network is shown in Figure 3.32. The simulated $S_{11}$ of the four element microstrip patch antenna array with the mushroom-like EBG matched the $|S_{11}| < -10$ dB with the impedance bandwidth of 130 MHz from the cut off frequencies of 4.68 GHz to 4.81 GHz, which represents a fractional bandwidth of 2.8%. The resonant frequency of the antenna array is 4.65 GHz, with a peak return loss value of -21 dB.
Figure 3.32  **Simulated $S_{11}$ (dB) of the four element microstrip patch antenna array with mushroom-like EBG structures at the corporate feed**

### 3.6.3.1 Radiation pattern

The simulated radiation pattern of the four element microstrip patch antenna array with EBG structures is shown in Figure 3.33 (a) and (b). In simulation, the peak directivity of the antenna array with and without mushroom-like EBG structures, is obtained as 16.25 dBi and 18.21 dBi at the resonant frequency of 4.8 GHz. The major disadvantage in the antenna array construction is the side lobe levels and back radiation. Figure 3.33 (a) shows the simulated four element antenna array without EBG structures, in which the side lobe levels are more. From Figure 3.33 (b) it is clear that the antenna array with EBG structures, has a significant reduction in the side lobe levels and back radiation. The comparison of different antenna designs, with and without mushroom-like EBG structures is shown in Table 3.5.
Figure 3.33 (a) and (b) Simulated radiation pattern of the four element microstrip patch antenna array without and with mushroom-like EBG structures

Table 3.5 Comparison of the different antenna designs of with and without mushroom-like EBG structures

<table>
<thead>
<tr>
<th>Antenna design</th>
<th>$S_{11}$(dB)</th>
<th>$S_{21}$(dB)</th>
<th>Antenna design</th>
<th>$S_{11}$(dB)</th>
<th>$S_{21}$(dB)</th>
<th>Gain</th>
<th>Side lobe(dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Patch antenna array with EBG</td>
<td>-9</td>
<td>-18</td>
<td>Patch antenna with EBG</td>
<td>-19</td>
<td>-</td>
<td>10</td>
<td>18</td>
</tr>
<tr>
<td>Patch antenna</td>
<td>-16</td>
<td>-30</td>
<td>Dual band antenna with EBG</td>
<td>-17.5</td>
<td>-18.5</td>
<td>13.5</td>
<td>10</td>
</tr>
<tr>
<td>Proposed design</td>
<td>-22</td>
<td>-33</td>
<td>Proposed design</td>
<td>-22</td>
<td>-33</td>
<td>13.71</td>
<td>7.5</td>
</tr>
</tbody>
</table>

3.6.4 Dual Frequency-Dual Polarised Microstrip Patch Antenna Array with Mushroom-like EBG Structures

The dual frequency-dual polarised two element microstrip patch antenna without mushroom-like EBG structures was discussed in chapter 2.
In this section, the simulated and measured $|S_{11}|$ (dB) of the dual band two element microstrip antenna array with mushroom-like EBG structures, are explained in Figure 3.34. In simulation, the $|S_{11}|$ (dB) graph shows the dual resonant frequencies, which are located at about 3.2 GHz and 4.72 GHz. The antenna array matched the ($|S_{11}| < -10$ dB) with the impedance bandwidth of 100 MHz and 110 MHz, from the cut off frequencies of 3.14 GHz to 3.24 GHz and 4.69 GHz to 4.78 GHz respectively, which represents the fractional bandwidth of 3.12 % and 2 % respectively. Two different dimensions of the mushroom-like EBG structures are incorporated, for the suppression of the surface waves at both the frequencies. In measurement, $|S_{11}|$ (dB) the resonant frequencies are located at about 3.1 GHz and 4.73 GHz and have the -10 dB impedance bandwidth of 70 MHz and 100 MHz, from the cut off frequencies 3.11 GHz to 3.18 GHz and 4.7 GHz to 4.8 GHz respectively, which represents the fractional bandwidth of 2.2 % and 2.1 %. The $|S_{11}|$ (dB) of the simulated value is greater than that of the measured value. From the graph result, it is shown that the placing of mushroom-like EBG structures near the inset feedline of the microstrip patch antenna array, considerably reduces the surface waves, which in turn increases the gain of the antenna array.

![Figure 3.34](image.png)

**Figure 3.34** Simulated and measured $S_{11}$ (dB) of dual frequency-dual polarized microstrip patch antenna array with mushroom-like EBG structures
3.6.4.1 Radiation pattern

The simulated and measured radiation patterns of the dual band two element microstrip antenna array with mushroom-like EBG structures are indicated in Figure 3.35 (a) and (b). In the measured radiation pattern there are some ripples, because of the connector and cable feed. In simulation, the peak directivity of the antenna array without EBG structures at 4.8 GHz is obtained as 13.2 dBi. In measurement, it is obtained as 9 dBi. The agreement between the simulated and measurement results is quite good. The simulated and measured results of the resonant frequency, return loss, VSWR and directivity are shown in Table-3.5.

![Simulated and measured radiation pattern of the dual frequency-dual polarized microstrip patch antenna array without and with mushroom-like EBG structures](image)

(a) Simulated directivity, measured co polarisation and cross polarisation.

(b) Simulated directivity, measured co polarisation and cross polarisation.

Figure 3.35 (a) and (b) Simulated and measured radiation pattern of the dual frequency-dual polarized microstrip patch antenna array without and with mushroom-like EBG structures.
Table 3.6 Simulated and measured results of the dual frequency-dual polarized microstrip patch antenna array with mushroom-like EBG structures

<table>
<thead>
<tr>
<th>Sl.No</th>
<th>Antenna parameters</th>
<th>Simulated</th>
<th>Measured</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>$f_1$</td>
<td>$f_2$</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequency(GHz)</td>
<td>3.01</td>
<td>4.73</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss(dB)</td>
<td>-13.8</td>
<td>-21.4</td>
</tr>
<tr>
<td>3.</td>
<td>VSWR</td>
<td>1.7</td>
<td>1</td>
</tr>
<tr>
<td>4.</td>
<td>Directivity(dBi)</td>
<td>9.5</td>
<td>13.2</td>
</tr>
<tr>
<td>5.</td>
<td>Bandwidth(MHz)</td>
<td>51</td>
<td>68.2</td>
</tr>
</tbody>
</table>

3.6.5 Dual Frequency-Dual Polarized Microstrip Patch Antenna Array with Compact Dual Band EBG Structures

An effort is made in simulation, for the two element microstrip patch antenna array to achieve dual polarization by tilting each radiating patch at $+45^\circ$ and $-45^\circ$ individually. In the same arrangement, the dual band is also obtained by giving dual feed to the radiating patches. Also an attempt is made for enhanced performance and compactness in EBG structures. Dual band EBG structures are designed and developed for suppressing the surface waves at both the band of frequencies. Figure 3.36 depicts the simulated $S_{11}$ of the dual band microstrip antenna array, with and without compact dual band EBG structures. The simulated $S_{11}$(dB) results of the dual band-dual polarized two element microstrip patch antenna array without compact dual band EBG structures, are 10.2 dB and 10.3 dB at the resonant frequencies of 3.01GHz and 4.6 GHz respectively. The simulated $S_{11}$(dB) results of the dual band-dual polarized two element microstrip patch antenna array with compact dual band EBG structures are -15.2 dB and -18 dB at the resonant frequencies of
3.22 GHz and 4.81 GHz respectively. The microstrip patch antenna array with EBG structures matched the \(|S_{11}| < -10 \text{ dB}\) with the impedance bandwidth of 150 MHz and 110 MHz from the cut off frequencies of 3.11 GHz to 3.26 GHz and 4.73 GHz to 4.84 GHz respectively, which represents the fractional bandwidth of 4.7 % and 2.3 % respectively. The maximum return loss values obtained at the resonant frequencies are -16.4 dB and -18 dB respectively.

![Two element patch antenna array](image)

Figure 3.36 Simulated \(S_{11}(\text{dB})\) of the dual frequency-dual polarized microstrip patch antenna array with compact dual band EBG structures

### 3.6.5.1 Radiation pattern

The simulated radiation patterns of dual band-dual polarized microstrip patch antenna array, with and without compact EBG structures, are shown in Figure 3.37 (a) and (b). The peak directivity in simulation for both the antenna arrays is obtained as 13.2 dBi and 16.32 dBi.
Figure 3.37 (a) and (b) Simulated radiation pattern of dual frequency-dual polarized microstrip patch antenna array without and with compact dual band EBG structures

Table 3.7 Simulated results of the dual frequency-dual polarized microstrip patch antenna array with compact dual band EBG structures

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna parameters</th>
<th>Simulated</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>$f_1$</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequency(GHz)</td>
<td>3.22</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss(dB)</td>
<td>-15.2</td>
</tr>
<tr>
<td>3.</td>
<td>VSWR</td>
<td>1.3</td>
</tr>
<tr>
<td>4.</td>
<td>Directivity(dBi)</td>
<td>10</td>
</tr>
<tr>
<td>5.</td>
<td>Bandwidth(MHz)</td>
<td>150</td>
</tr>
</tbody>
</table>

3.6.6 Aperture Coupled Microstrip Patch Antenna with FSS

The simulated and measured results of the aperture coupled patch antenna were discussed in chapter 2. Figure 3.38 shows the $S_{11}$ simulated and measured results of the aperture coupled microstrip patch antenna with FSS. From the Figure it is seen that the antenna has dual frequencies. The
simulated return loss results of the aperture coupled antenna with FSS matched the ($|S_{11}| < -10$) with the impedance bandwidth of 230 MHz and 170 MHz obtained from both the cut off frequencies of 3.21 GHz to 3.44 GHz and 4.72 GHz and 4.89 GHz, which represents a fractional bandwidth of 6% and 4.5%. The dual resonant frequencies $f_1$ and $f_2$ are 3.27 GHz and 4.82 GHz. The measured return loss results of the aperture coupled antenna with FSS matched the ($|S_{11}| < -10$) with the impedance bandwidth of 180 MHz and 230 MHz from both the cut off frequencies of 3.16 GHz to 3.34 GHz and 4.56 GHz and 4.79 GHz, which represents a fractional bandwidth of 5.6% and 5%. The dual resonant frequencies $f_1$ and $f_2$ are 3.13 GHz and 4.69 GHz. The simulated return loss values of the aperture coupled microstrip patch antenna with FSS are obtained as -17.6 dB and -17.8 dB. The measured return loss values of the aperture coupled microstrip patch antenna with FSS are obtained as -14.16 dB and -14.01 dB. Compared with the simulated result, the measured result shows good agreement within the operational frequencies. The measured result displays lesser bandwidth, and the resonant frequencies are shifted to the lower frequencies. This is caused by minor differences between the practical and simulated models of the FR-4 substrate, and the dielectric constants, are varied for the theoretical and practical samples, which caused the frequency decreases. The measured and simulated resonant frequency, return loss, VSWR, bandwidth and directivity are shown in Table-3.8.

![Figure 3.38 Simulated and measured S_{11}(dB) of the aperture coupled microstrip patch antenna with FSS](image)

Figure 3.38 Simulated and measured $S_{11}$(dB) of the aperture coupled microstrip patch antenna with FSS
3.6.6.1 Radiation pattern

The simulated and measured radiation patterns of the aperture coupled microstrip patch antenna, with and without FSS, are indicated in Figure 3.39 (a) and (b). There are some ripples and distortions due to the sandwich of the two dielectric substrates to form a multilayer structure, feed cable and connector. The simulated peak directivity of the aperture coupled microstrip antenna, without and with FSS, is obtained as 6 dBi and 8 dBi at the resonant frequency of 4.8 GHz. In measurement, the radiation characteristics of co polarization and cross polarization are obtained. The aperture coupled microstrip antenna with FSS suppresses the surface waves and increases the directivity.

![Figure 3.39](image)

**Figure 3.39** (a) and (b) Simulated and measured radiation pattern of the aperture coupled microstrip patch antenna without and with FSS
Table 3.8 Simulated and measured results of the aperture coupled microstrip patch antenna with FSS at frequencies $f_1$ and $f_2$ (3.12 GHz and 4.8 GHz)

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna parameters</th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.</td>
<td>Resonant frequency(GHz)</td>
<td>$f_1$ 3.27 $f_2$ 4.82</td>
<td>$f_1$ 3.13 $f_2$ 4.69</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss(dB)</td>
<td>-17.6 -17.8</td>
<td>-14.16 -14.01</td>
</tr>
<tr>
<td>3.</td>
<td>VSWR</td>
<td>1 1</td>
<td>1.45 1</td>
</tr>
<tr>
<td>4.</td>
<td>Directivity(dBi)</td>
<td>6 7.8</td>
<td>5 6</td>
</tr>
<tr>
<td>5.</td>
<td>Bandwidth(MHz)</td>
<td>230 170</td>
<td>180 230</td>
</tr>
</tbody>
</table>

3.6.7 Two Element Aperture Coupled Microstrip Patch Antenna Array with FSS

The simulated and measured $S_{11}$ of the aperture coupled two element microstrip patch antenna array without FSS was shown in Figure 2.23 in chapter 2. The simulated and measured $S_{11}$ of the aperture coupled two element microstrip patch antenna array with FSS is shown in Figure 3.40. The simulated dual resonant frequencies $f_1$ and $f_2$ are located at 3.24 GHz and 4.8 GHz respectively. The antenna array matched the ($|S_{11}|< -10$ dB) with the impedance bandwidth of 180 MHz and 180 MHz, obtained from the cut off frequencies of 3.12 GHz to 3.3 GHz and 4.69 GHz to 4.87 GHz, which represents the fractional bandwidth of 5.5 % and 3.6 % respectively. The measured dual resonant frequencies $f_1$ and $f_2$ are located at are 3.3 GHz and 4.68 GHz respectively. The antenna array matched the ($|S_{11}|< -10$ dB) with the impedance bandwidth of 130 MHz and 110 MHz, obtained from the cut off frequencies of 3.22 GHz to 3.35 GHz and 4.6 GHz to 4.71 GHz, which represents the fractional bandwidth of 4 % and 2.4 % respectively. The
simulated maximum return loss values of microstrip antenna array are -17.2 dB and -18.1 dB. The measured maximum return loss values of the microstrip antenna array are -14.7 dB and -15.6 dB. The bandwidth of the simulated antenna array with FSS is greater than that of the measured antenna array with FSS. This is because the dielectric constant and loss tangent of the dielectric material are not stable in the simulation and practical models. The measured and simulated results show good concurrence within the operating range.

Figure 3.40  Simulated and measured $S_{11}(\text{dB})$ of the two element aperture coupled microstrip patch antenna array with FSS

3.6.7.1  Radiation pattern

The simulated and measured radiation patterns of the aperture coupled two element microstrip antenna array with FSS are indicated in Figure 3.41 (a) and (b) at the resonant frequencies of 3.12 GHz and 4.8 GHz. Some ripples and distortions are seen in the measured radiation pattern, due to the joining of the two dielectric substrates with a minor air gap and connector. The peak directivity of the antenna array with FSS in simulation and measurement, is obtained as 13.9 dBi and 11.2 dBi. It is observed from the results, that the directivity enhancement of the antenna array is achieved by
placing the FSS as a superstrate. The measured and simulated results are in good agreement with each other.

Figure 3.41 (a) and (b) Simulated and measured radiation pattern of the two element aperture coupled microstrip patch antenna array without and with FSS

Table 3.9 Simulated and measured results of the two element aperture coupled microstrip patch antenna array with FSS at frequencies \( f_1 \) and \( f_2 \) (3.12GHz and 4.8 GHz)

<table>
<thead>
<tr>
<th>Sl.No</th>
<th>Antenna parameters</th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>( f_1 )</td>
<td>( f_2 )</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequencies(GHz)</td>
<td>3.24</td>
<td>4.8</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss(dB)</td>
<td>-17.2</td>
<td>-18.1</td>
</tr>
<tr>
<td>3.</td>
<td>VSWR</td>
<td>1.10</td>
<td>1</td>
</tr>
<tr>
<td>4.</td>
<td>Directivity(dBi)</td>
<td>9.8</td>
<td>13.9</td>
</tr>
<tr>
<td>5.</td>
<td>Bandwidth(MHz)</td>
<td>180</td>
<td>180</td>
</tr>
</tbody>
</table>
3.6.8 Four Element Aperture Coupled Microstrip Patch Antenna Array with FSS

The design and construction of the four element aperture coupled microstrip patch antenna array without FSS has been shown in Figure 2.25 in chapter 2. The simulated and measured results $S_{11}$ of the aperture coupled four element microstrip patch antenna array with FSS is shown in Figure 3.42. From the diagram it is seen, that the antenna array operated at dual band of frequencies. The simulated dual resonant frequencies $f_1$ and $f_2$ are located at about 3.2 GHz and 4.67 GHz respectively. The antenna array matched the ($|S_{11}| < -10$ dB) with the impedance bandwidth of 130 MHz and 140 MHz, obtained from the cut off frequencies of 3.12 GHz to 3.25 GHz and 4.59 GHz to 4.73 GHz, which represents the fractional bandwidth of 4.1 % and 3 % respectively. In measurement, the dual resonant frequencies $f_1$ and $f_2$ are located at about 3.18 GHz and 4.7 GHz respectively, and their maximum return losses are 22.18 dB and 21.09 dB respectively. The measured return loss of the aperture coupled four element microstrip antenna array with FSS matched ($|S_{11}| < -10$ dB) with the impedance bandwidth of 170 MHz and 160 MHz, is obtained from the cut off frequencies of 3.11 GHz to 3.28 GHz and 4.73 GHz to 4.89 GHz respectively, which represents a fractional bandwidth of 5.3 % and 3.4 % respectively. The measured result is compared with the simulated result, and shows good agreement within the operating frequencies. There are minor differences between the measured and simulated results. This is because of the minor difference between the practical and simulated models of the FR-4 material. Also, the dielectric constant is not constant through the substrate, which causes the frequency shift. The measured and simulated resonant frequency, return loss, bandwidth, VSWR and directivity are shown in Table-3.7.
Figure 3.42 Simulated and measured $S_{11}(\text{dB})$ of the four element aperture coupled microstrip patch antenna array with FSS

3.6.8.1 Radiation pattern

The simulated and measured radiation patterns of the four element aperture coupled microstrip antenna array with FSS that operates at 3.12 GHz and 4.8 GHz, are indicated in Figure 3.43 (a) and (b). There are some ripples and distortions in the measured radiation pattern, due to the air gap between the two dielectric materials used to form the multilayer structure of the antenna array, feed cable and connectors. The simulated and measured peak directivity of the aperture coupled microstrip antenna array without FSS is obtained as 14 dBi and 11dBi. The simulated and measured peak directivity of the aperture coupled microstrip antenna array with FSS is obtained as 17.2 dBi and 15.1 dBi. The simulated and measured results are in good agreement with each other.
Figure 3.43 (a) and (b) Simulated and measured radiation pattern of the four element aperture coupled microstrip patch antenna array without and with FSS

Table 3.10 Simulated and measured results of the four element aperture coupled microstrip patch antenna array with FSS at frequencies $f_1$ and $f_2$ (3.12GHz and 4.8 GHz)

<table>
<thead>
<tr>
<th>Sl. No</th>
<th>Antenna parameters</th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>$f_1$</td>
<td>$f_2$</td>
</tr>
<tr>
<td>1.</td>
<td>Resonant frequency (GHz)</td>
<td>3.2</td>
<td>4.67</td>
</tr>
<tr>
<td>2.</td>
<td>Return loss (dB)</td>
<td>-17.4</td>
<td>-22.3</td>
</tr>
<tr>
<td>3.</td>
<td>VSWR</td>
<td>1.1</td>
<td>1</td>
</tr>
<tr>
<td>4.</td>
<td>Directivity (dBi)</td>
<td>13</td>
<td>17.2</td>
</tr>
<tr>
<td>5.</td>
<td>Bandwidth (MHz)</td>
<td>130</td>
<td>140</td>
</tr>
</tbody>
</table>

3.7 SUMMARY

Three different EBG structures (mushroom-like, fork-like and compact dual band) are designed and developed. To improve the performances, such as return loss, bandwidth and directivity of the inset feed
microstrip antenna array, Electromagnetic Bandgap structures (EBG) are used. The fork-like EBG is compact but has a narrow bandwidth. The mushroom-like EBG is simple and has a wider bandwidth. The compact dual band EBG has the ability to operate in dual frequencies for the suppression of surface waves. There is a need to design a low cost, compact EBG structure that operates in WLAN applications. The designed EBG structures are placed along with the microstrip antenna array, for enhanced performance. It was demonstrated that, not only was the input impedance matching possible, but also the frequency bandwidth could be wider. The aperture coupled microstrip antenna array is designed for minimized feed network loss and spurious radiations. The performance of this antenna array configuration is further improved by placing the Frequency Selective Surface (FSS) as a superstrate. To validate the proposed concepts, the experimental results have been presented, and they helped in making the following conclusions, showing that:

i) The directivity and bandwidth improvement of 1.2 dBi and 25 % respectively is achieved, when a single microstrip antenna with mushroom-like EBG structures is placed at the inset feedline. Also, the antenna size reduction of 18 % is achieved compared to that of the conventional placing of EBG structures.

ii) The directivity and bandwidth enhancement of 3 dBi and 58 % respectively is achieved, with a two element inset feed microstrip antenna array with mushroom-like EBG structures. Also, a size reduction of 32 % is achieved, by placing the EBG structures near the corporate feedline.

iii) Compared to the dual band-dual polarized microstrip patch antenna array, the dual band-dual polarized microstrip patch
antenna array, in which each radiating patch is given with dual feed, has improved return loss and effective dual polarisations at each radiating patch tilted at $+45^\circ$ and $-45^\circ$ respectively. Compactness is also achieved when the dual feed dual band microstrip antenna array with compact dual band EBG is placed at the feedline.

iv) The directivity and bandwidth enhancement of 3 dBi and 74 % respectively, is achieved, when an aperture coupled microstrip antenna array is incorporated with FSS as a superstrate layer.