Wideband bandstop frequency selective structures

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To my family and those who wish to use the contents of this work for the goodness of mankind.

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Summary

Frequency selective surfaces brought several design challenges to light, and have been central to much research recently. Known limitations are: poor filtering characteristics; sensitivity to large oblique incident angles when used as both filters and polarization manipulators, which results in a shift frequency response and deteriorated in- and out-of-band transmission and reflection responses; and the inability to maintain a satisfactory angular performance over a very wide bandwidth. This thesis introduces three novel designs based on a new class of three-dimensional frequency selective structure to present promising wideband bandstop structures suitable for different applications, following simple design procedures.

To fulfill the wideband requirements in many applications, and since bandstop structures with very wide band and stable response have not been reported yet, ultra-wide bandstop structures with good angular stability are proposed in this thesis. Providing a wide band and an angular stable response is a challenging task as the response of the upper part of the band deteriorates due to relatively larger unit-cell sizes. Therefore, using a novel and common design procedure based on a parallel strip line (PSL) unit-cell, two structures are proposed using different concepts, exhibiting ultra wide bandstop responses and different unit-cell thicknesses and angular stability levels. The main advantage of the PSL is that the unit-cell size perceived by the incident E-field has no effect on the frequency response and bandwidth, giving us freedom to control its size for an improved angular stability. The first design exploits higher order harmonics to achieve a fractional bandwidth of 78%. It features a new design approach for this class of structures to excite and properly position the harmonic to widen the bandwidth. It also provides a wide harmonic-free out-of-band response and represents a good solution when better out-of-band responses are in demand. The second design involves a cascaded structure of stacked PSL unit-cells. The fractional bandwidth attained is 100% with good angular stability up to 60° over the entire band. Both structures are relatively superior to the state of the art alternatives in terms of bandwidth, out-of-band

performance, and angular stability.

Polarization rotation function has also been introduced to the bandstop structure to demonstrate a stable rotator with good cross-polarization isolation utilizing a parallel plate waveguide (PPW). It is based on the same design procedure proposed for the first two structures. An L-shaped slot is etched on one of the PPW's sides to trap and rotate the orthogonal incident electric field component. The realized rotator has an insertion loss of 1.3 dB at 10 GHz and exhibits a good angular performance that outperforms the available alternatives in angular stability. Due to the structure's stop band performance, it suppresses the co-polarized field component at the frequencies surrounding the center frequency of the rotated component, improving the cross-polarization isolation.

Finally, a few topics are suggested regarding related future works. They involve other ultra-wide bandstop solutions with improved bandwidth, as well as conformal and multi-band variations of frequency selective structures.

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List of Acronyms

Abbreviations	Full Expressions
2D	Two-dimensional
3D	Three-dimensional
ADS	Advanced design system
BW	Bandwidth
CST	Computer simulation technology
FSS	Frequency selective surface / frequency-selective structure
HFSS	High frequency structure simulator
PCB	Printed circuit board
PEC	Perfect electric conductor
PMC	Perfect magnetic conductor
PPW	Parallel plate waveguide
PSL	Parallel strip line
RCS	Radar cross section
SIW	Substrate integrated waveguide
TEM	Transverse electromagnetic

Chapter 1

Introduction

1.1 Background

Frequency selective surfaces (FSS) have been the focus of many researchers in recent years. FSSs are essentially periodic structures arranged in a two-dimensional (2D) infinite array [1]. They offer filtering (bandpass/stop) and polarization manipulation characteristics and can be deployed in many applications in association with antennas as well as radomes. They are used to reduce the side/back-lobe, radar cross section (RCS), electromagnetic interference and bore sight error in radome systems, and may also be deployed as mere polarization rotating/converting surfaces suitable for certain applications. Moreover, there are applications which demand increased bandwidths, such as shipboard radars, which suggest the need for high performance and higher bandwidth FSS designs. Furthermore, polarization manipulation surfaces are also required to compensate the de-polarization losses in radomes, which produce angle tracking errors in radar applications [2]. They are also used in satellite applications where circularly polarized waves are exploited in polarimetric imaging, radiometers, multi-path effect reduction, atmospheric absorption, and other civil/military applications.

Exciting an FSS is done using incident plane waves with no individual generator connected to each element. A portion of the energy at certain frequencies is transmitted forward upon striking the surface, and another portion reflects back. The geometries of an FSS are numerous, as will be seen in the following section. On the other hand, structures that are frequency selective can also take three-dimensional (3D) forms with significant thickness compared to their conventional counter part, and are called frequency selective structures. More details are explained in a following sections.

1.2 Two-dimensional Frequency Selective Surfaces

Two-dimensional FSS, often referred to as the conventional type, involves etching different geometrical shapes on planar printed circuit boards (PCBs) with insignificant conductor thickness with respect to free space wavelength, such that capacitive and inductive properties are realized to shape the response accordingly. The profile thickness of the final product is usually low when a single layer is considered. The design of this type is relatively simple. The forms of the geometrical elements etched on the substrates are classified into different groups: center connected or N-poles, loop, plate, and hybrid types [1], as shown in Figure 1.1. Many more designs such as gangbuster, fractal, miniaturized-element and others have been developed to obtain similar or better bandpass and bandstop responses. Conventional FSS can also be built using multiple layers of such thin structures.

In general, conventional FSSs with unit-cell sizes that are comparable to half of the operating wavelength suffer from reduced angular stability under large oblique incident angles, smooth filtering responses, and offer narrow bandwidths. Single conductor lay-

ers are found to be insufficient to realize wide bandwidths and sharp responses, even in bandpass FSSs, which is why multi-layers are employed. To tackle the unit-cell size, a class of FSS with miniaturized elements is introduced in [3] to reduce the unit-cell sizes, producing highly stable bandpass responses under oblique incidence. Using the same class, a bandstop design was introduced in [4], producing wide bandwidths (up to 5:1) exploiting simple filter theory. However, it features a limited out-of-band performance with wide bandwidths and has to deploy multi-layers to attain them. Other miniaturized designs could also tackle the angular stability issue in different ways in [5, 6, 7, 8]. Ultimately, a common conclusion can be drawn: sharp and angular stable wideband responses are difficult to achieve using two-dimensional surfaces. It is worth noting that this thesis focuses on bandstop structures.

1.3 Three-dimensional Frequency Selective Structures

3D frequency selective structures are periodic structures that feature significant conductor thicknesses (relative to λ_o) extending along the direction of propagation, compared to their conventional counterparts, offering more flexibility in the designs for potentially better performance. According to the literature, the ratio $\frac{L_c}{\lambda_o}$ is around 10^{-3} to 10^{-4} in conventional FSS and $\geq 10^{-1}$ in the 3D counterpart, where L_c is the conductor thickness. A few structures of different forms have been proposed. So far, multi-mode cavity resonators have shown good potential for realizing frequency selective structures, where the cavity could be an SIW structure, shielded microstrip line, or a conventional waveguide structure. The desired frequency response may be controlled by varying the the number of propagating modes in the resonators and their coupling with the air. Earliest designs of the 3D form involved deploying substrate integrated waveguide tech-

Figure 1.1: Classification of geometrical element shapes of conventional 2D FSS [1].

nology (SIW) in [9, 10], whose posts' lengths are equal to the substrate thickness. The second novel design and significant step made was presented in [11] where an array of shielded microstrip lines was deployed. Such structures can exhibit a pseudo-elliptic response and achieve sharp filtering characteristics and an angular stable performance. Only a few 3D structures that do not employ multi-mode resonators have been proposed, but either exhibited a smooth response [12], or even posed difficulties in realization [13].

1.4 Applications

1.4.1 Filtering

Frequency selective surfaces/structures can be used to:

1. Reduce the out-of-band RCS of an antenna;
2. Realize multi-frequency reflectors in multi-band reflector antenna applications;
3. Improve the front-to-back ratio of antennas;
4. Selecting or shielding of certain frequencies in civil/military applications;
5. Protect communication systems from interferences by nearby or far existing transmitters, especially for military applications as in shipboard radars.

1.4.2 Polarization Manipulation

As polarization manipulators, frequency selective surfaces/structures can be used in applications, such as the following:

1. Satellite applications where circularly polarized waves are used to reduce the multipath effect and atmospheric absorption (linear-to-circular converters);
2. Compensate for de-polarization effects in radomes where unwanted angle tracking errors are generated in radar applications;
3. Applications requiring polarization isolation like polarimetric imaging radars and radio meters.

1.5 Motivation and Objectives

A wide variety of methods concerning the design of bandpass 3D structures are developed and proposed in the literature, but with little focus on bandstop structures. Designs that exhibit angular stable, broadband and sharp responses, are to be explored here. Moreover, broadband bandstop structures are demanded in applications where the higher-order harmonics generated from a certain source are to be rejected.

An example of such applications is ship-born communication systems where the frequency of operation of their radar installations and most other systems are relatively low (typically below 3 GHz). Such systems could generate higher-order harmonics which can interfere with the X-band systems co-located on the ship. They may also generate grating lobes, producing high RCS levels when the antennas are covered with slotted radomes, increasing the chance of detection by other X-band radars. Thus, equipping these systems with radomes having broadband bandstop frequency selective surfaces/structures is a safe practice.

Furthermore, the few polarization rotation structures that have been reported in the literature heavily rely on multi-layer technology. The alternative among the recently reported polarization rotators has a relatively large unit-cell size and can only handle relatively small oblique incident angles. Thus, a better polarization rotator design could be proposed, which exhibits an improved angular performance by exploiting the pre-existing high performance multi-mode resonator structure, as it was the case in [14] when an earlier SIW structure was adopted and modified to design a polarization rotating.

1.5.1 Limitations of Bandstop Frequency Selective Surfaces/Structures

1.5.1.1 Two-dimensional Surfaces

It is a challenging task to design conventional bandstop FSSs when angular stable sharp responses are targeted. Typically, the performance achieved due to the common design geometries whose sizes are comparable to the operating wavelength share one or more of the following:

1. Smooth filtering [8, 15, 16, 17], following the transmission curve of a first order resonator;
2. Relatively large unit-cell sizes comparable to the wavelength of operation, thereby deteriorated responses under oblique incidence without miniaturization techniques;
3. Increased complexity and fabrication cost when sharper filtering (higher order) is required. Multilayer structures are typically employed;
4. Relatively narrow bandwidths for applications similar to the one discussed in Section 1.5.

1.5.1.2 Three-dimensional Structures

Bandstop frequency selective structures are not free of design challenges either. One or more of the following common issues are shared by such structures:

1. Due to certain geometries [12, 18, 19, 20, 21, 22, 23] they may pose fabrication difficulties or require relatively complex assemblies compared to their conventional counterparts. This can also degrade their mechanical robustness and ultimately the performance stability;

2. In the case of the structure in [11], sensitivity to polarization is a limitation since the design is made in such a way that the incident E-field should have a certain polarization;
3. Smooth filtering responses, as it appears in some structures [12, 18, 19, 20, 21, 22];
4. The highest fractional bandwidths with sharp selectivity reported so far is 50% [11] which may not cater for the applications mentioned in Section 1.5.

1.5.1.3 Polarization Manipulators

Polarization manipulation frequency selective surfaces/structures pose the following challenges during the design process:

1. Increased physical complexity of the structures reported in the literature [24, 25, 26, 27], which exploit multiple cascaded layers of meander-lines, wire grid and strip grid structures to achieve rotation, leading to large thicknesses;
2. Sensitivity to oblique incidences in the design based on periodic structures (substrate integrated waveguide (SIW)) in [14], whose unit-cell size is comparable to the wavelength.

1.5.2 Objectives

The objectives of this thesis are: 1) Propose two structures using two different design concepts, to achieve wide bandstop responses with good angular performance and sharp filtering features, based on one basic geometrical structure proposed in this thesis. 2) Present a new class of polarization rotators using another variation of the basic

geometrical structure proposed earlier, exhibiting an angular stable response with good cross-polarization isolation. Simplified theory and design details for each structure are established, with simple circuit modeling procedures (for the bandstop structures), and ultimately verifying the theoretical computations with measurements.

1.6 Organization of the Thesis

Chapter 1 introduces the FSS and frequency selective structures, their types, and the functions they offer. The motivations and objectives behind our work are included.

Chapter 2 provides a comprehensive literature survey on 3D frequency selective structure and the solutions found so far to enhance the response, angular performance, and bandwidth, are summarized for reference in our proposed designs in this thesis. The other part of the survey concerns the reported polarization rotator designs and their strengths and weaknesses for comparison with our proposed polarization rotator.

Chapter 3 shows a proposed design that exhibits a wide bandstop response using a simple structure, achieving 78% bandwidth with good out-of-band performance, with a relatively stable response under oblique incidence given its bandwidth. The necessary parametric studies, the theory of operation, design details, and measured results are included.

Chapter 4 presents another structure which is specifically designed to cope with relatively large oblique incidences while featuring a wide bandwidth of 100% using a cascaded configuration. Simulation and measurement results are also shown.

Chapter 5 describes the proposed polarization rotator, its simple design procedure and theory behind its operation, with the necessary parametric studies, including final

results under normal and oblique incidences.

Chapter 6 draws a conclusion for this thesis and introduces ideas for future work.

Chapter 2

Literature Review

2.1 Introduction

Frequency selective structures and frequency selective surfaces (FSS) have both received extensive attention recently. Applications that employ them are both civil and military oriented. They fulfill bandpass and bandstop filtering requirements, which can be deployed for antenna sidelobe suppression, radar cross section (RCS) reduction, polarization manipulation (rotation and conversion), and more beside the filtering functions they offer. Mainly, they fall into two general categories, bandpass and bandstop types. Our focus will be on the bandstop type.

Conventional bandstop FSS designs feature a few limitations, as shown previously, including performance degradation under high oblique incident angles, limited bandwidth, and often poor selectivity. Numerous works have been carried out in an effort to come up with three-dimensional bandstop structures that offer wider bandwidths, angular stable as well as sharp filtering responses. Significant highlights from both conventional and 3D bandstop designs are reviewed here. These works could partially tackle the above men-

tioned problems. Structures have been proposed to either cover a wide bandwidth with smooth filtering or angular dependent response, or cover relatively narrow bandwidths with stable and sharp filtering response. Similarly, polarization manipulation structures based on period structures proposed feature relatively large unit-cell sizes and thus a more angle dependent response. This chapter also explores the potential to tackle the issues for better solutions in terms of filtering bandwidth, angular stability, sharp response, aiming toward more established and simplistic design.

2.2 Frequency Selective Structures/Surfaces as Filters

Some highlights from conventional bandstop FSSs are reviewed first, revealing their geometries and performance features.

Numerous conventional bandstop designs proposed in the literature using different techniques to extend the bandwidth. A wideband bandstop FSS was proposed in [28] by combining two individual bands through cascading two bandstop structures, each covering a narrow band. A -10 dB bandwidth of 19% was achieved using a thin PCB structure. Other designs could exhibit ultra wideband responses in [16], [29], and [30] where a double metallic layer of identical spiral strips, garland structures, and tripole loops were used, respectively. More were introduced in [31] where two substrate layers of stacked FSS reflector were employed to achieve a 122% BW, and in [32] as well, attaining a BW of 74%, both of which have structures of different shapes on each metal layer. Other conventional structures were built with miniaturized unit-cell elements to tackle the angular stability limitations in [8, 29], where spiral loops and swatiska unit-cell elements were used, respectively, such that the resonance elements are folded to minimize

the unit-cell's physical size.

These conventional designs feature one or more of the following: Unclear design procedures and/or theory of operation, wide bandwidths with smooth responses, sharp responses with an out-of-band performance showing spurious resonances or high insertion loss, high dependence on angle of incidence for those offering very wide bandwidths. It hints that one or two of the issues could be tackled at a time with conventional band-stop designs so far.

Since 3D frequency selective structures are a recent development and only a few have been reported compared to their conventional alternatives, they are briefly reviewed here, highlighting their strengths and weaknesses. Also, it is worth mentioning that fewer studies have reported bandstop structures compared to their bandpass counterparts. Since our focus is mainly on the bandstop structures, it is a good chance to review the bandstop related literature works here. The purpose is to focus on the concepts behind their operation, which were developed to tackle the limitation issues explained in Chapter 1.

Multiple 3D structures have been proposed in [18] by extruding conventional cross and Jerusalem cross shapes into 3D structures (pyramidal and sawtooth) which resulted in enhanced angular performance, lower frequencies for the 3D variations, and a relatively better fractional bandwidth. Figure 2.1 shows some of the extruded variations of the structure. Bandwidths were narrow in general and responses were smooth. Since it was a study oriented work, the fabrication can be difficult, and not suitable for practical applications. The highest fractional bandwidth achieved was using a 3D square loop

geometry, reaching 50% with a thickness of $0.024\lambda_o$, with acceptable angular performance and a unit-cell size of $0.195\lambda_o$. Investigations were done to produce a variant that exhibits 41% bandwidth at -10 dB and very good angular stability with a thickness of $0.1025\lambda_o$ and a relatively small unit-cell size of $0.133\lambda_o$. Moreover, the out-of-band response worsens as the oblique incident angles increase, producing additional zeros adjacent to the fundamental zero.

Figure 2.1: 3D view of various extruded conventional FSS elements [18].

Other 3D bandstop structures were proposed in [12, 19] where a 3D cylinder is used. The transmission zero is determined by the diameter of the rings and width of the conducting strip. For lower values of the cylinder unit-cell lengths, the structure exhibits a bandstop response, while the response changes significantly when high values of cylinder length are used, showing a clear bandpass characteristic. Though the lengths of the structure can be significantly reduced by loading the cylinders with a dielectric material, the response is smooth rather than sharp, and the structure is bulky. However, the maximum fractional bandwidth achieved for the stop band reached 112%, yet with an out-of-band response featuring a high insertion loss. To operate as a bandstop structure, the thickness of a unit-cell needs to be around $0.5\lambda_o$ and the unit-cell size is close to 79% when no dielectric filling is used. This implies that the performance may be angle dependent since the unit-cell size as seen by the incident electric field is relatively large.

The same author later proposed two other novel 3D structures. The first is an angular stable design in [20], which was realized by enlarging the aperture of an array of square cross-section cylinder resonators. The structure was tested under oblique incident angles up to 45° and have demonstrated good stability. The fractional bandwidth achieved is 18% showing a smooth response, with a unit-cell size of $0.88\lambda_o$ and a thickness of $0.2\lambda_o$. Using the aperture enlarging technique, a very good angular stable response was achieved though the unit-cell size is relatively large. The near out-of-band response featured a high insertion loss.

Later, the author presented another novel design of a reconfigurable frequency selective structure using spring resonator structures in [21], as shown in Figure 2.2. Mechanical tuning can be achieved by applying pressure on the spring elements that are fixed on a platform. Its circuit model is established with a simple design approach well backed by equations. The structure's response switches between bandpass and bandstop when varying the height h of the spring elements. As a bandstop structure, the fractional bandwidth exhibited is relatively high (28% at -20 dB) with a smooth response identical to that obtained using the designs proposed in [12, 19], and a high insertion loss in the out-of-band region. The minimum unit-cell size and thickness are $0.4\lambda_o$ and $0.16\lambda_o$, respectively. While the angular performance has not been tested, the unit-cell size hints

Figure 2.2: 3D view of the spring resonator frequency selective structure [21].

to an angle dependent response.

A miniaturized 3D bandstop structure was proposed in [33] providing an angular stable performance. The design employs multi-layer (4 layers) PCB to layout a four-legged loop structure, as shown in Figure 2.3. The unit-cell size used is 0.2λ , which yielded a very stable response under oblique incident angles up to 60° . The structure was realized using an inexpensive FR4 board with a relatively high tangent loss of $\delta = 0.025$. However, the filtering response is relatively smooth, featuring a relative bandwidth of ($\sim 10\%$).

The performance features exhibited by the 3D bandstop frequency selective structures designs reported in the literature so far are summarized in Table 2.1.

Since the 3D bandstop structures have been reported less than their bandpass counterparts, some bandpass designs presented in the literature shall also be highlighted here to review the concepts to achieve the attractive performance features we are aiming for in our bandstop designs: wideband, angular stable, and sharp selectivity.

Figure 2.3: 3D view of the four-legged loop unit-cell [33].

Table 2.1: Summary of proposed 3D bandstop structures.

Feature	Extruded [18]	Cylindrical [12, 19]	Square cylinder [20]	Spring [21]	4-legged loop [33]
<i>Unit-cell size</i> / λ_o	19.5%	79%	88%	40%	20%
<i>Thickness</i> / λ_o	2.4%	50%	20%	16%	4.8%
<i>%BW/-10 dB</i>	50%	112%	18%	28% @ -20 dB	10%
<i>Angular performance</i>	Very good up to 60°	-	Good up to 45°	Angle dependent	Very good up to 60°
<i>Selectivity</i>	Smooth	Sharp	Smooth	Smooth	Smooth
<i>Near out-of-band</i>	Spurious stop bands	High insertion loss	High insertion loss	High insertion loss	Good

One of the earliest forms of 3D structures was fabricated using SIW cavities to achieve bandpass filtering in [9]. Sharp filtering characteristics were obtained, producing resonances as a result of the periodicity of the grid slot array as well as the cavity, yielding a more angular stable solution compared to most conventional FSS designs that use single layer PCBs. Yet, the insertion loss becomes significant at large oblique incidences.

The same authors later demonstrated in [34] that a cascaded structure could be realized [9], sharpening the roll-offs significantly, producing a highly selective filter with similar tolerance to oblique incidence. Moreover, in [10], the first quasi-elliptical bandpass structure was designed using the SIW technology by exciting two different resonances in the same SIW cavity by generating cross coupling between two cascaded SIW cavities to realize multi-mode filters, which greatly reduced the volume of the product as well. However, the fractional bandwidth is narrow (1.67%) with high sensitivity to oblique incidence. Dual band and two-sharp-band variations were developed later by the same authors in [10, 35], presenting sharper responses and relative stability under a maximum of 30° incidence.

A quasi-elliptic bandpass response was obtained with a broad bandwidth of about 30% in [22]. The structure consists of a 2D periodic array of apertured cavities that

exploit six circular slots on each side of the cavity classified into primary and secondary resonators to achieve the overall response. A sharp filtering response is achieved only on the upper skirt of the band.

Using a waveguide based structure, a quasi-elliptic bandpass response was also obtained using the design proposed in [23, 36]. The design is based on a waveguide and two cross-shaped elements connected by a transmission line. The length of the transmission lines control the frequency of the response. Sharp roll-offs were obtained with a wide 3 dB fractional bandwidth of $\sim 70\%$.

A broad pass band was obtained by multiple resonances generated by the structure proposed in [13]. The design is meant for shielding by employing a metallic cube having its corner cut and edge chamfered, which poses practical difficulties during fabrication. The 3 dB bandwidth achieved is 13% and the unit-cell size is $1\lambda_o$, which may not qualify the structure to perform well under oblique incidence.

In conclusion, only a few wideband frequency selective structures are addressed in the literature, and it is noticed that obtaining wideband, sharp and angular stable responses is still a challenging task. However, there is room for improvement and there is a clear trend followed to tackle sharpness and wide bandwidths. Multi-mode resonators are employed in unit-cells, as it can be seen in the 3D bandpass designs reviewed here. However, having both sharp filtering and broadband characteristics in one structure appears difficult here, especially when good angular performance is in demand.

Recently, a novel 3D structure was proposed in [11], deploying an array of vertically

Figure 2.4: Geometry of the structure and unit-cell of the stacked microstrip frequency selective structure [11].

stacked shielded microstrip lines. The array and its unit-cell are shown in different views in Figure 2.4. The concept involves exciting two TEM modes when the incident plane wave shine in a perpendicular manner to the microstrip conductors. Two regions will form for the E-field to pass through. One is the substrate region, and the other is the air region. Both regions will allow the structure to generate reflection zeros at certain points in the band. Having the structure designed carefully with a high dielectric constant can allow the two regions to generate reflection zeros such that higher order resonant transmission zeros are placed between the two reflection zeros, forming the desired stop band. This structure will form the base of our studies in this thesis, and will further be explored here to gain an insight into its operating principles.

The design involves calculating and optimizing a few parameters, including the dielectric constant of the substrate ϵ_r , ratio of t/b , ratio of d/h , and length of the unit cell L . The structure exhibits a pseudo-elliptic bandstop response, with sharp roll-offs, maximum bandwidth of 52.5%, and a stable performance under oblique incidences, when carefully designed. Thicknesses of $0.25\lambda_o$ and less are achievable with very small unit-cell sizes smaller than $0.15\lambda_o$. It also exploits the well known PCB wet etching fabrication technique, which makes it relatively easy to realize and imposes no restrictions in fabrication as many 3D designs may do.

The structure was adopted several times to produce multiple designs with custom bandpass and bandstop responses and improved physical characteristics as well in [11, 37, 38, 39, 40, 41, 42, 43, 44]. The structures have shown sharp filtering and little angle dependence, leading to angular stable responses under a large variation of oblique incidence, and provided wide bandwidths, room for tunability features, and multi-band designs, with clear theory and design procedures for most of them. Since the structure has proven promising characteristics, it is further studied in this thesis for further development and simplification.

The general equivalent circuit model of a multi-mode resonator for this type of cavity resonance based structure is shown in Fig. 2.5. Any 3D frequency selective structure based on multi-mode cavity resonators may be expressed as a cascade of three elements: an air-to-resonator discontinuity, multi-mode resonator, and a second air-to-resonator discontinuity. The air-to-resonator discontinuities could be expressed as K-inverters (impedance inverters) sandwiched between two transmission line sections [46]. The electrical lengths of the transmission line sections of the discontinuities are expressed

Figure 2.5: Equivalent circuit of a multi-mode cavity resonator with external coupling structures [45].

by $\Phi_0, \Phi_1, \dots$, and Φ_N while the lengths of the multi-mode resonators are expressed by $\Psi_0, \Psi_1, \dots$, and Ψ_N , each is represented at an individual propagation mode. N refers to the number of propagating modes, Z_o and Z_n refer to the port and transmission line characteristic impedances, respectively. K_n is the impedance of the nth K-inverter.

The reflection coefficient can be represented in a closed-form expression after calculating the ABCD matrix of the given equivalent circuit and later calculating the S_{11} parameter. The ABCD matrix corresponding to the equivalent circuit of the multi-mode cavity resonator becomes:

$$(21)$$

$$\begin{bmatrix} 0 & jK \\ \frac{1}{jK} & 0 \end{bmatrix} \cdot \begin{bmatrix} \cos\Psi_n & jZ_o\sin\Psi_n \\ \frac{j\sin\Psi_n}{Z_o} & \cos\Psi_n \end{bmatrix} \cdot \begin{bmatrix} 0 & jK \\ \frac{1}{jK} & 0 \end{bmatrix} = \begin{bmatrix} -\cos\Psi_n & \frac{-jK^2\sin\Psi_n}{Z_o} \\ \frac{-jZ_o\sin\Psi_n}{K^2} & -\cos\Psi_n \end{bmatrix}$$

where the outer portions of the transmission lines Φ_o of the discontinuities on both sides have been considered as part of the source/load feed.

Since the S_{11} conversion matrix is given by:

$$S_{11} = \frac{A + B/Z_o - CZ_o - D}{A + B/Z_o + CZ_o + D} \quad (2.1)$$

For a given propagation mode, the reflection coefficient is found using the following equation:

$$S_{11}^{(n)} = \frac{j(\bar{k}_n^4 - 1)\tan\Psi_n}{2\bar{k}_n^2 + j(\bar{k}_n^4 - 1)\tan\Psi_n} \quad (2.2)$$

where

$$\bar{K}_n = \frac{K_n}{\sqrt{Z_o Z_n}}, \quad \varphi_n = 2\phi_n + \Psi_n$$

Thus, when:

$$\left| \sum_{n=1}^N S_{11}^{(n)} \right| = \begin{cases} 1 & \text{transmission zero occurs} \\ 0 & \text{transmission pole occurs (reflection zero)} \end{cases} \quad (2.3)$$

The values controlling the reflection response are Ψ_n and $\bar{K}_n$. The location of $S_{11} =$

0 is found when $\Psi_n = 180^\circ$ and 360° , representing the fundamental and harmonic resonant frequencies, or when $\overline{K}_n = 1$. The higher order modes exploited in these cavities can yield similar responses achieved in [47, 48, 49], where the same principle is used to achieve ultra wide bandwidths.

2.3 Frequency Selective Structures/Surfaces as Polarization Manipulators

Many polarization manipulation structures have been reported. They are generally classified into three groups: polarization selectors, linear polarization rotators that can introduce rotation to the incident plane waves, and polarization converters which mainly convert linearly polarized waves into circularly polarized waves. Polarization rotators will be reviewed briefly since they are relatively less reported in the literature, and have their own limitations, as it will be seen later.

One of the earliest forms of polarization rotators exploited multi-layer structures [24]. It consists of three or more parallel equi-spaced wire grids whose wires are oriented in different angles. By properly choosing the number of wire grid layers used, their spacing, and their angle of orientation, powerful rotation performance can be obtained. An example design demonstrated a seven layer 90° rotator achieving 42% relative bandwidth with very good angular stability (up to 60° incidence), and very low insertion loss of about 0.1 dB.

Another design based on the same multi-layer concept and its general principles and

design features were proposed in [25], in which a 45° rotation performance is demonstrated using only three layers. The axial ratio is less than 2 dB over the operating band, with less than 0.5 dB of insertion loss.

A meander line based multi-layer structure was later reported in [26], featuring arbitrary rotation of a linearly polarized wave, and another multi-layer design featuring periodic arrays of wires presented in [27]. Many layers are needed to achieve a satisfactory performance, which renders the final structure thick and heavy.

All the previous designs are primarily based on multi-layer technology where many layers have to be employed in order to perform the arbitrary rotation functions they were proposed to achieve. However, recently, a polarization rotating structure was proposed in [14] based on an SIW cavity unit-cell. Two slots are etched on both the top and bottom conductor layers of a unit-cell, and in a perpendicular fashion, as shown in Figure 2.6. The incident wave of a certain polarization will be selected by the slot which is orthogonal to it such that the energy of the wave is coupled to the slot. The slot generates a field in the cavity, which is carefully designed to resonate at that particular frequency point, which ultimately excites a perpendicular slot at the other side to radiate the energy. The 3 dB bandwidth achieved is 9.1% at the center frequency 35 GHz, with an insertion loss of 0.2 dB in the pass band. The unit-cell size is $0.67\lambda_o$, which is relatively large for a stable response to be achieved under a highly oblique incidence, which is the reason why the angular stability was only tested under 10° and 20° . In an effort to test the structure's stability, it was re-modeled and its performance was found to be highly angle dependent.

Later, a group demonstrated a reflective surface in [50] exhibiting polarization rota-

Figure 2.6: Geometry of the proposed polarization rotation structure based on SIW cavity [14].

tion of 90° using exactly the same structure used in [14], except that both of the slots are etched on the top layer of the structure. The lower layer is a full PEC, which blocks the transmission and helps to fully reflect the field. The bandwidth performance achieved is about 9.5% with 90° rotation and a co-polarization reflection coefficient lower than -10 dB. Oblique incidence tests were done up to 25° revealing its angular dependence.

2.4 Summary

In conclusion, wideband bandstop frequency selective structures have seldom been reported in the literature. Also, many steps were taken to improve the filtering response and performance under oblique incidence, which involved: (i) using SIW technology for improving the filtering response since they share some of the properties of rectangular waveguides; (ii) using multiple resonators to enhance the sharpness; (iii) generating cross coupling between two cascaded SIW cavities; (iv) cascading structures; (v) extruding conventional 2D structures; (vi) using multiple slots (rings, lines, etc) on different sides of a single 3D unit-cell to excite more resonances; (vii) and using miniaturized

loop structures to reduce the unit-cell size. In general, these solutions tackled one or two of the three main issues mentioned before (bandwidth, stability and filtering response), leaving the other(s) unsolved, which require novel structures in which all the three qualities meet. Moreover, some applications have wider bandwidth demands (as mentioned in Chapter 1), while at least maintaining the angular performance over these bands. Lastly, to avoid complexities seen in many works, the novel structures in this thesis have to be simple and include clear design steps taken to facilitate the design process (modeling, computation and fabrication).

With regards to polarization rotators, it is noticeable that most of the proposed structures are heavily based on multi-layer technology, which might not qualify them for some practical applications as they may reserve a large space. Also, it is obvious that structures that employ SIW technology tend to possess large unit-cell sizes, which introduces angle dependence when an angular stable response is in demand. However, the structure succeeded in performing the function using only one substrate layer, unlike the other solutions presented in the literature.

Thus, following this literature review and the motivations mentioned, there will be a room for us to carry out our research toward the desired improvements, which are angular stable performance for both filters and rotators, and wide band and sharp filtering responses for filters. Given that, simple and reliable design steps are to be developed with the necessary theoretical analysis required to elaborate the mechanism behind each structure's operation.

Chapter 3

Array of Vertically Stacked Parallel Strip Lines

3.1 Introduction

As investigated in Chapter 2, many of the proposed wide band frequency selective structure/surface designs exhibit one or more of the following performance limitations: Smooth filtering characteristics; sensitivity to oblique incident angles which causes shifted and/or deteriorated in-band and out-of-band transmission and reflection responses; and bandwidth limitations. Thus, two ultra wide bandstop structures are proposed in this thesis, based on an array of stacked parallel strip lines (PSLs) as well as a cascaded array of stacked PSLs. Also, the mechanical properties are to be tackled such that the cascaded design maintains a robust build for a stable performance.

A new structure is introduced in [11], showing a quasi-elliptic bandstop response using an array of vertically stacked shielded microstrip lines. Each shielded microstrip line

Figure 3.1: Geometry of the vertically stacked PSL of the proposed frequency selective structure.

is considered as a unit-cell, representing a dual-mode resonator. The structure has been thoroughly investigated and is found to provide a rejection bandwidth of around 52.5% at best, with high selectivity and stable performance under large oblique incident angles. However, since higher bandwidths are in demand, techniques are adopted to enhance the bandwidth, retaining the stable angular performance. After conducting careful investigations on alternative potential design variations of the unit-cell and the behavior of the two parallel resonators individually, it was established that using a PSL unit-cell, simpler modeling and design steps can be deduced. Moreover, it is found that an extra transmission zero of a higher order resonance could be introduced and placed adjacent to the two fundamental zeros already generated by the dual mode resonator. This could significantly widen the bandwidth, achieving 78% at -10 dB. This resulted in a struc-

ture with good out-of-band performance since a harmonic was shifted down, leaving a wide upper out-of-band region with low insertion loss and free of spurious resonances. However, since harmonics are more sensitive to oblique incidence, the performance could deteriorate under large incident angles with angle dependence compared to the original structure, yielding a more sensitive alternative, yet with a significant increase in bandwidth.

The performance exhibited by this structure is superior to that of the state of the art designs available to date in terms of bandwidth, angular stability given the wide fractional bandwidth, out-of-band response as well as the simple design steps.

3.2 Array of Vertically Stacked Parallel Strip Lines

3.2.1 Design and Analysis

The first proposed structure is shown in Figure 3.1. It is desirable to have a structure that could be implemented using simple design equations and guidelines. This is made possible using an array of vertically stacked PSLs separated by a very thin gap s in the x-direction such that each unit-cell resembles two PSLs stacked on top of each other with an air gap filling one of them, and a substrate with a relatively high dielectric constant ϵ_r filling the other, as shown in Figure 3.1(c). The substrate has a thickness of h and a conductor thickness of 0.035 mm. The period length h extends toward both sides of the y-direction. The front and back views are perfectly identical.

This and all the following structures proposed in this thesis operate under a linear

and single polarized wave whose E-field lines are parallel to the yz-plane, having $\phi = 90^\circ$. Oblique incident angles are represented by θ only, keeping $\phi = 0^\circ$, indicating the xz-plane as the only oblique incidence plane.

Each unit-cell is simulated with perfect magnetic conductor (PMC) sidewalls. A PSL represents a dual-mode resonator, which generates two reflection zeros separated by a relatively wide distance across the frequency spectrum. The modes excited are due to two different regions in the resonator. One is the air region, which is filled with air, and the other is the substrate region, which is filled with a dielectric.

The PSL unit-cell design process involves critically designed parameters such that harmonic transmission zeros (higher order resonances) are excited within the band, which is formed by two widely separated reflection zeros, and properly positioned adjacent to the fundamental transmission zeros, producing a wider band. The higher order resonance is obtained by using a relatively higher dielectric constant compared to the air ($\epsilon_r = 1$) for the substrate to increase the coupling, which allows including more transmission zeros within the stop band as will be shown in the next section, as well as translating the whole band (transmission and reflection zeros) to a lower region. An appropriate value of ϵ_r could be found such that the translation of the whole band can properly place the newly excited harmonic within the stop band to widen the bandwidth.

According to [51], for a PPW transmission line of our case, it resembles a rectangular waveguide which is bifurcated by a PEC layer of zero thickness orthogonal to the direction of the incident E-field lines. The waveguide bifurcation case is shown in Figure 3.2 with cross-sectional and side views. In our case, there is no extension waveguide beyond

the bifurcation point ($d = 0$) since our dual PPW configuration is directly exposed to free space (at T_1). The circuit model of this bifurcation case is shown in Fig. 3.3, which is eventually characterized by a discontinuity capacitor at each of the air-to-air and air-to-substrate interfaces that are formed between each of the unit-cell's two parallel plates. Our final circuit model parameters will involve the aforementioned capacitors to yield more realistic results, and as close as possible to those evaluated using the full-wave model. Therefore, the simplified circuit model of a PSL unit-cell is shown in Figure 3.4. Considering it as a dual-mode resonator, two parallel resonators are considered to account for all the reflection zeros generated by the structure. Discontinuity capacitors can be seen at the air/unit-cell interfaces.

Figure 3.2: E-plane bifurcation of a PPW with equivalent circuit model [51].

Figure 3.3: Equivalent circuit of PPW with E-plane bifurcation [51].

The two parallel transmission lines modeled in the circuit have an electrical length

$$(31)$$

Figure 3.4: Circuit model of a PSL/PPW unit-cell.

of $\theta = 180$ at frequencies where reflection zeros are generated. Since the PSL exhibits a very close response to that of a PPW, the impedance values are calculated using simple PPW equations. Alternatively, they can be evaluated using full-wave simulation by checking the port impedances calculated at each mode generated by the PSL unit-cell.

As for the full-wave model, the modes generated are shown in Figures 3.5 and 3.6. It shows the top view of a unit-cell where the top, bottom and middle lines are perfect electric conductors (PECs) and the side walls are perfect magnetic conductors (PMCs). Two modes can be seen, and as mentioned before, one in the substrate region and the other in the air region. In Figure 3.6, it shows that though the PSL's width W is not as wide as the unit-cell because of the small introduced gap s , almost all of the field distribution exists only within the two regions. Therefore, a PSL with a small gap s can be approximated as a PPW in this case.

The dispersion diagram is also shown here to relate the phase change that happen in radians/m versus frequency when using a dielectric constant ϵ_r of 4.3, comparing both the PPW and PSL cases against each other. Figure 3.7 shows that two fundamental TEM modes appear (Mode 1 and Mode 2) with almost identical trends when PPW and

Figure 3.5: Two modes generated in a PPW (top view). Red: strong. Blue: weak or none.

Figure 3.6: Two modes generation in a PSL (top view). Orange: strong. Blue: weak or none.

PSL cases are compared. The two lines correspond to the two modes excited in the air and the substrate regions shown before. Comparing the computed impedance values of the two regions of the PPW against those of the PSL using full-wave simulation shows a good agreement between the two.

3.2.2 Design Example

3.2.2.1 Theory and Procedures

The design process of the generic bandstop structure unit-cell has been made simpler and more systematic than that presented in [11] since the unit-cell is now considered as a PPW in our case. Thus, it allows us to exploit PPW equations in [52] to easily

Figure 3.7: Dispersion diagram of PPW and PSL compared using a substrate with $\epsilon_r = 4.3$, $h = 1.22$ mm, $p = 1.85$ mm, $W = 2$ mm, and $s = 0.3$ mm.

calculate the impedance without resorting to full-wave solutions as well as the discontinuity capacitance values using closed-form equations. Moreover, the curves that are meant to facilitate the design of the structure's length are provided since the unit-cell's length is not simple to deduce. Therefore, the structure's initial parameters can easily be calculated for our case.

To model the circuit parameters, impedances of the resonators are calculated using the following set of equations:

$$Z_{sub} = \frac{\eta}{\sqrt{\epsilon_r}} \frac{h}{W}, \quad Z_{air} = \eta \frac{p-h}{W}, \quad Z_o = \eta \frac{p}{W} \quad (3.1)$$

where Z_{sub} is the impedance of the resonator formed by the PPW filled with substrate material, Z_{air} is the impedance of the other resonator formed by the PPW which is filled with air, Z_{total} is the total impedance of the port (source), $\eta = 120\pi$ ohms, p = period length of unit-cell in mm, h = substrate thickness in mm, W = unit-cell width in mm and ϵ_r = dielectric constant of the substrate.

It is worth noting that the expression for calculating the port's characteristic impedance Z_o is obtained after considering the fact that for one unit-cell, it extends across both the air and the substrate regions vertically, making it equal to the period (p), and across the unit-cell's width horizontally (W). It is placed at a distance from the parallel strips and in free space where ϵ_r is unity. To calculate the initial electrical lengths (in terms of frequency) of the resonators in the circuit model, the following equations are used:

$$F_{air@180} = \frac{c}{2L}, \quad F_{sub@180} = \frac{c}{2L\sqrt{\epsilon_r}} \quad (3.2)$$

where c = speed of light in m/sec, L = substrate/strip length in mm and $F_{x@180}$ = frequency at which a transmission line exhibits a fundamental resonance while being 180° in electrical length.

At the reference plane T_1 shown in Figure 3.2, the corresponding circuit parameters are expressed according to the circuit model shown in Fig. 3.3 using the following equations [51]:

$$\frac{X_1}{Z_o} = \frac{p-h}{p} \cot\left(\frac{2\pi d}{\lambda_o}\right), \quad \text{and} \quad \frac{X_2}{Z_o} = \frac{h}{p} \cot\left(\frac{2\pi d}{\lambda_g}\right) \quad (3.3)$$

where

$$\frac{2\pi d}{\lambda} = \frac{2p}{\lambda} \left[\frac{p-h}{p} \ln \frac{p}{p-h} + \frac{h}{p} \ln \frac{p}{h} \right] + S_1 \left(\frac{2p}{\lambda}; 0, 0 \right) - S_1 \left(\frac{2(p-h)}{\lambda}; 0, 0 \right) - S_1 \left(\frac{2h}{\lambda}; 0, 0 \right) \quad (3.4)$$

$$S_1(x; 0, 0) = \sum_{n=1}^{\infty} \left(\sin^{-1} \frac{x}{n} - \frac{x}{n} \right) \quad (3.5)$$

The equivalent circuit shown in Figure 3.3 only applies when $2p < \lambda_g < \infty$. The value λ_g represents the guided wavelength in the substrate region and λ_o represents the free space wavelength in the air region.

Since X_1 and X_2 represent the imaginary impedances used to obtain the discontinuity capacitance values at the reference plane T_1 , using equations (3.3), (3.4) and (3.5), one can obtain the capacitances of both the air and substrate interfaces using the two simple closed form expressions:

$$C_{air} = \frac{\tan \left(\frac{2\pi d}{\lambda_o} \right) p}{2\pi 10^9 f_c (p-h) Z_o}, \quad \text{and} \quad C_{subs} = \frac{\tan \left(\frac{2\pi d}{\lambda_g} \right) p}{2\pi 10^9 f_c h Z_o} \quad (3.6)$$

where

$$X_c = \frac{1}{\omega C}$$

where f_c = center frequency and C_x is the discontinuity capacitance.

The principle of operation of this structure lies behind the analysis of each of the air and substrate regions in a unit-cell individually and as shown in Fig. 3.8. Since the

Figure 3.8: Full-wave simulated reflection response of the air and substrate regions individually compared with that of the two regions combined (unit-cell) using $L = 12$ mm, $W = 2$ mm, $s = 0.3$ mm, $h = 1.22$ mm, $p = 1.85$ mm and $\epsilon_r = 4.5$.

length of the unit-cell is directly related to the reflection response S_{11} , each of the air and substrate regions is simulated and plotted to graphically analyze the positions of the zeros. Following the fact that having the two resonators resonate at the same frequency point in-phase generates a reflection zero, and out-of-phase to generate a transmission zero [38], the new proposed design is deduced. It is worth highlighting that the air and substrate regions show resonance points every 180° , meaning that the phase inverts at every subsequent resonant point across the band.

By choosing a proper ϵ_r , suitable positioning of the reflection zeros of the two resonators can be arranged such that the combined response resembles that shown in Fig. 3.8. The first reflection zero generated by the air region meets the first higher order zero generated by the substrate region at the same frequency point and out-of-phase,

resulting in a transmission zero. This promises an uninterrupted rejection band which spans from the fundamental resonant point to the second higher order harmonic, both of which are generated by the substrate. Eventually, this method exploits the fundamental resonance generated by the air region to widen the in-band rejection response. On the other hand, the first higher order resonance generated by the air region is placed at the point where the third higher order resonance generated by the substrate region exists. The two resonant points are in-phase, yielding a reflection zero, keeping a wide out-of-band region beyond the rejection band. The same cycle of combinations continues and a second image of the rejection band forms by higher order resonances, with increasing angular dependence.

The three curves plotted in Fig. 3.8 are obtainable using the equivalent circuit model as well after recalculating the circuit parameters when ϵ_r is changed. It is done by individually simulating the air and substrate regions and producing two responses such that their combined response, when the resonators are connected in parallel, yields to a three transmission zero stop band.

The transmission response against different values of ϵ_r is compared in Fig. 3.9 to show the direct effect on the transmission response, showing that values of ϵ_r near 4.5 are appropriate for our design to obtain a wide rejection band. The reflection curve for the unit-cell shown in Fig. 3.8 is related to the transmission curve when $\epsilon_r = 4.5$ here. It is worth noting that values close to 4.5 for ϵ_r can regenerate the same transmission response using different lengths of the unit-cell, which means that they are suitable for a wide range of frequencies. Also, low ϵ_r values produce widely separated transmission zeros across the spectrum of operation. As ϵ_r goes higher, the general response

Figure 3.9: Parametric study demonstrating the effect of ϵ_r on the transmission response

will shift downwards while strengthening the higher order resonances to a point where three transmission zeros are generated within the band in our case, which are strong enough to generate a wider stop band below -10 dB. Ultimately, the very common and commercially available ϵ_r value of 4.3 is found to be suitable for our design.

As for the physical length L , Figure 3.10 shows the curve which helps in choosing an initial value for the unit-cell's length for different frequency points extending from 5 GHz to 26 GHz, given that the fractional bandwidth provided is fixed at 78%. Every frequency point on the curve represents the -10 dB starting frequency point of the fundamental resonance on the transmission curves so that the designer may choose the suitable initial length based on this rejection level at the starting point of the band of interest. A cheaper laminate is eventually used compared to that deployed in the original design proposed in [11].

Figure 3.10: Unit-cell length design curve with most optimal values of ε_r shown

The ratio $\frac{h}{p}$ is chosen as 0.65, initially. It controls the in-band rejection and out-of-band reflection magnitudes. Increasing the ratio would mean decreasing the air gap given a fixed dielectric substrate thickness h , which would lower the in-band transmission further but raise the out-of-band reflection ears. Therefore, tuning the ratio in the optimization stage might be needed till the best balance is obtained between reflection and transmission responses, maintaining them within the -10 dB limit. The curve can be regenerated for any value of ε_r .

After the initial value of the length is chosen, optimization may be performed to fit the lowest, center, or highest frequencies properly within the desired band. Later, the physical model could be designed and simulated as a unit-cell using a full-wave simulation package, in which the results should closely match those obtained using the circuit model. The circuit model helps in designing and simulating using simpler tools (circuit

simulators), and could help to save much time compared to the full-wave. For the example design, the center frequency is set at 10 GHz, which provides an acceptable frequency range to avoid accuracy issues during the fabrication process and substrate losses that rise while frequencies rise. It also corresponds to a reasonable thickness for the structure.

3.2.2.2 Simulated Results

The simulation results are obtained using ADS package for circuit simulation, and CST package for full-wave simulation. Fig. 3.11 shows the model used in CST, with the following simulation setup parameters:

1. Gaussian input pulse that extends from 3 to 25 GHz
2. TE polarization is used, in which the E-field is orthogonal to the parallel strips
3. Frequency domain solver is used in CST to solve the problem, using Finite Element Method
4. Average number of tetrahedral mesh cells (tetrahedrons) that have been adaptively meshed through multiple passes is 17,000

Figure 3.11: 3-D model designed in CST simulation package.

Table 3.1: Circuit and physical parameters of the stacked PSL array bandstop structure.

Circuit parameters: C/pF, Z/Ω, θ/GHz						
C_a	C_s	Z_a	Z_s	Z_o	θ_a	θ_s
0.009	0.0099	139.71	130.47	410.255	10	4.822
Physical Dimensions / mm						
ϵ_r	s	h	p	L	W	
4.3	0.3	1.22	1.85	15	2	

5. S-parameter error threshold value = 0.01, with 10,000 sample points for each simulation
6. Average time taken for each simulation is 3 minutes and 30 seconds
7. Peak memory used on average in each simulation is 344 MB

The computer used to obtain the average simulation speed given above uses a 4th generation Intel i7 CPU with four cores, each operating at 2.4 GHz with 8 GB of system memory.

A bandstop structure is to be designed with an arbitrary starting frequency of 5.85 GHz and a stop point at 13.6 GHz (since the relative bandwidth is fixed at 78%), both at -10 dB on the transmission curve. Table 3.1 lists the physical and circuit parameters that can be obtained after using the design equations and the curve in Fig. 3.10 to calculate the physical length L . By drawing a horizontal line for the target -10 dB starting frequency of the fundamental stop band, the intersection point with the curves will decide the approximate length to use and reveals the stop frequency as well.

The results obtained using the circuit model have been compared to those produced using full-wave simulation in Figure 3.12. It is noticeable that the results from both methods are in a good agreement, especially for the lower part of the band. Slight fre-

Figure 3.12: Circuit model vs. full-wave simulation results for the stacked PSL array bandstop structure under normal incidence.

quency shifting appears in the upper part because: the capacitors in the circuit model are considered fixed and only calculated for the center frequency; gap s is not modeled in the circuit since the unit-cell is considered a perfect PPW; the tangent loss of the substrate changes with frequency and not involved in circuit simulation. Also, the full-wave model is done with zero conductor thickness.

Since the third transmission zero of the lower band is generated by a higher order resonance, it yields a more sensitive transmission zero under oblique incidence, but with the advantage of freeing a wider out-of-band region with low insertion loss (0.67 dB at worst). The rejection level could be further improved by simply reducing the air gap (increasing the $\frac{h}{p}$ ratio). However, this might slightly raise the reflection ears on both sides of the stop band, which is why the value may need optimization. If the out-of-band reflection response is of low priority, the air gap could be reduced to improve the

rejection region and even boost the bandwidth slightly.

Figure 3.13: Sensitivity of transmission response near 11 GHz with different unit-cell types: PPW, PSL, and Microstrip.

Since the third transmission zero at 12.85 GHz is a higher order resonant point, a study has been done on the weakest point of the transmission performance (~ 11 GHz here) to observe the effect of oblique incidence on the rejection performance at that point since transmission response is the priority in this design. The parametric study is shown in Figure 3.13. It compares using a PPW (with no gap s) with a microstrip (with gap s) only on one side of the substrate, and a PSL with gap s available on both sides of the substrate. It shows that while frequency increases, both the transmission and reflection responses deteriorate quickly in the case of using PPW. While the other two cases demonstrated significant improvement and alleviated much of the deterioration. However, the PSL case is adopted here to allow us to exploit the PPW equations in the design process. In the next chapter where the second structure is proposed, it will be shown that PSL has another advantage over the shielded microstrip that is used in [11]

in that it achieves better angular stability at high frequencies.

Figure 3.14: Full-wave simulated response of the bandstop structure under oblique incidence.

Figure 3.14 shows the performance of the structure under oblique incidence. Testing was done only for a relatively low variation of incidence angles since a higher order resonance is involved. It can be shown that the in-band transmission response formed by the third transmission zero at 11 GHz deteriorates quickly, reaching -8.5 and -4.8 dB at 20° and 40° , respectively. The lower and higher out-of-band transmission response stay in a better shape compared to those of the in-band response under oblique incidence. The highest out-of-band insertion loss observed under normal incidence is -0.64 dB at 17 GHz and degrades to about -1.4 dB at the same frequency under 40° .

(a) A strip of adjacent unit-cells

(b) Backbone comb structure

(c) Assembled structure

Figure 3.15: Fabricated prototype showing individual parts and the final assembly

3.3 Fabrication and Measurement

The fabrication process is implemented using normal wet etching technology and involves two types of structures in general. The first is the main strip which has all the copper traces that constitute many long strips of adjacently placed PSLs as shown in Figure 3.15(a), with a total length of $3 \lambda_o$. The second part is the backbone comb structure shown in Figure 5.10(b). It is an FR-4 laminate with no conductor layers that is proposed to attain mechanical stability and keep the air gap uniform along the strips to obtain a better performance. The backbone can provide nearly accurate air gap values that are not easily obtainable using dielectric spacers as it has been tradi-

Figure 3.16: Setup for measuring transmission performance.

tionally done, especially that the values correspond to non-standard thicknesses that are not obtainable commercially. The total number of strips needed for this stacked structure is 90, which amounts to $3 \lambda_o$ in total. Note that the structure is square shaped here.

The transmission response measurement setup is shown in Figure 3.16. It involves using an identical pair of horn antennas to see through the structure which is fixed in a window to block the unwanted and fringing radiation from passing to the receiving horn. The response measured with the structure (partial pass) is normalized relative to the full pass response (without FSS) after removing the indirect components from both the full pass and the partial pass cases before normalization. This is done in the data processing stage and has the benefit of removing the unwanted field components that arrive through reflections (ceiling, ground, and sidewalls) as well as diffraction at the measurement screen. These indirect components are estimated by blocking the FSS opening in the measurement screen with a PEC material such that no direct transmission is possible. Only the indirect fields' data can then be recorded to be eliminated later. Such components, if significant, may add up to the transmission magnitude, yielding

inaccurate results. The window, which is placed exactly in the middle, is rotated when oblique incident tests are performed.

In addition, a technique is used to tackle the unwanted reflections that appear in the time domain signal by properly time gating the pulse of the measured signal during the calibration process instead of processing the frequency domain signal directly. This helps to alleviate the adverse effects of the unwanted reflections that can distort the result but on the expense of trimming insignificant details related to the time domain pulse of interest. A properly chosen cosine filter is later used to eliminate the high frequency components that can arise after the trimming process if necessary.

Figure 3.17: Full-wave simulated vs measured transmission response of the bandstop structure with three transmission zeros under normal incidence.

The measured transmission response under normal incidence is shown in Figure 3.17. A good match is observed, with transmission zeros at the same positions and an accept-

Figure 3.18: Full-wave simulated vs measured transmission response of the bandstop structure with three transmission zeros under 20° incidence.

Figure 3.19: Full-wave simulated vs measured transmission response of the bandstop structure with three transmission zeros under 40° incidence.

able agreement of the envelop within the stop band. The oblique incidence responses are shown in Figures 3.18 and 3.19.

3.4 Summary

This proposed design has shown that a bandwidth of 78% is attainable using a modified structure, and properly chosen ϵ_r to control the harmonics and exploit them in a constructive manner to widen the rejection band. This structure is characterized by its simplicity in terms of both the analysis and the assembly processes. However, it is sensitive to the variation in ϵ_r , which is why a substrate of good quality needs to be used.

This structure can tolerate a small variation of oblique incident angles, and is designed to allow a wide range of frequencies to pass efficiently below and above the band of interest, which surpasses the latest designs available in the literature in terms of the out-of-band performance. It is also superior to most of the designs proposed to date for its sharp filtering characteristic, similar to that exhibited in [11]. An almost equal ripple is observed in both the pass band and stop band, and the distances between the transmission poles and zeros are short enough to produce sharp leading and trailing edges at the beginning and the end of the stop band. This produces a response which is very similar to that of an elliptical filter, and agrees with the conclusion made in [11] where periodically arranged PCBs make it possible to obtain an FSS with pseudo-elliptic response. The unit-cell has a size of $0.07\lambda_o$, and a thickness of $0.5\lambda_o$ at the center frequency.

Chapter 4

Array of Cascaded Vertically Stacked Parallel Strip Lines

4.1 Introduction

In the previous chapter, a wide-band bandstop structure was proposed based on exciting higher order resonances that are adjacently placed near the fundamental resonances to widen the band by choosing a proper value of the dielectric constant. The response was stable under only a small variation of oblique incident angles and was designed out of a cheaper laminate to achieve the wide band goal. It is also characterized by good lower and higher out-of-band responses.

In this chapter, an alternative ultra wide bandstop structure is proposed using a cascaded configuration. It involves designing two individual unit-cells to cover two adjacent bands, as well as the matching section which connects the two unit-cells with minimal loss after cascading. It was found that using a shielded microstrip unit-cell, the structure

fails to yield an angular stable response as the performance at the higher portion of the band is significantly deteriorated. As a result, the ground conductor of the microstrip structure was modified such that it forms a parallel strip line (PSL) structure, which helps to dramatically improve the performance under oblique incident angles given the same unit-cell width used in [11]. The unit-cell width can be made even smaller to reduce the size perceived by the incident field for better angular performance. The -10 dB fractional bandwidth achieved is 100%, producing a wider bandwidth and a more stable alternative than the previous structure. Also, the PSL structure, once again, helps us to exploit the simple PPW equations to establish the basic circuit model.

The performance exhibited is superior to that of the state of the art designs available to date, according to the latest literature, in terms of better bandwidth, stability under large oblique incident angles over the operating band, as well as the out-of-band response.

4.1.1 Design and Analysis

The design steps involve realizing two PPWs with a certain strip length for each to produce two wide rejection bands, then cascade them to combine the two responses, with a small gap g separating between them. Figure 4.1 shows the structure's geometry. The unit-cell design is carried out twice, once for each individual PSL, using a similar procedure to that presented in the previous section. The resulting structure can operate under single polarization and in a similar fashion to the previous structure.

The two PSLs are to be etched on the same substrate. This helps to mechanically stabilize the structure, and thereby the response. The reason spacing s is introduced

Figure 4.1: Geometry of the cascaded structure.

between adjacent unit-cells is to obtain a more stable response at high frequencies by limiting the unit-cell size. As shown before, the filtering performance of a unit-cell is not affected by this slight modification, and the bandwidth covered per unit-cell is still the maximum this structure could offer (52.5%) since the ratio of the width of the top strip to that of the bottom strip is close to unity, according to the design procedure in [11] and the ratio of $\frac{h}{p}$ is close to 0.635, slightly different from that in the previous structure. It is worth noting that PPW, microstrip and PSL unit-cells exhibit a very similar response under normal incidence, but the PSL will have the edge when a large bandwidth is to be covered with stable angular response, which is the case in this design.

A unit-cell has a substrate thickness of h , strip lengths $S1$ and $S2$ for two individually designed copper strips, a gap g that separates the cascaded PSLs, and a period length p , which determines the period of a unit-cell ($h + \text{air region thickness}$). The substrate thickness chosen is 1.27 mm, with a dielectric constant of $\epsilon_r = 9.8$ and a tangent loss of $\delta = 0.0035$. The separation gap s is kept very small and it is optimized in the last step

of the design process for tuning the rejection magnitude (weak effect), bandwidth, and stability under oblique incident angles.

Figure 4.2: Circuit model of a PSL unit-cell.

The circuit model is shown in Figure 4.2. The impedance values (Z_a , Z_s and Z_o) represent the PPW characteristic impedances, which are very close to those of the PSLs, for the air region, substrate region, and the overall characteristic impedance of the structure as seen from the port, respectively. The electrical length values θ_x for the strip lines are then calculated such that their length = 180° at the guided wavelength λ_g , except for θ_t values where calculation is performed for 90° . The capacitor values at the air-to-strip and strip-to-air discontinuities can be calculated using the equations in 3.2.2.1.

It should also be noted that the gap g separating the strips works as a matching part and is modeled in Figure 4.2 as a very short resonator. Its value is chosen such that its electrical length would be short enough not to introduce any reflection/transmission zero and affect the overall response, such that it only matches the impedances of both resonators at its sides while acting as an all pass section, and keeping the overall structure as short as possible. Also, it should not be too small to avoid excessive end-coupling between the strips $S1$ and $S2$. The optimal value is found to be $\sim \frac{\lambda_g}{4}$ of the shortest

resonator, which makes the design similar to that of a quarter wavelength transformer. Moreover, having a substrate as a medium between both PSLs helps to avoid undesirable reflections since the electric field propagates along a structure with uniform ϵ_r . This is besides giving the structure better mechanical stability.

4.1.2 Design Example

4.1.2.1 Theory and Procedures

Given a substrate of thickness $h = 1.27$ mm, $\epsilon_r = 9.8$, a small unit-cell size is chosen with width $W = 2$ mm with a small separation gap $s = 0.3$ mm for better stability under oblique incidence. The optimal ratio of $\frac{h}{p}$ is chosen to be 0.65 to choose an initial value of the period length. Considering both conductors' thickness to be 0.035 mm each, $p = 2.02$ mm ~ 2 mm, which represents the distance from the bottom of the substrate to that of the next.

As for the circuit parameters, the design follows the same procedure and equations used in the previous section (3.1 to 3.6). The impedances and discontinuity capacitances can then be calculated. However, for this design, a substrate with $\epsilon_r = 9.8$ is used. Figure 4.3 shows the values of the length of a unit-cell versus the -10 dB transmission response and transmission zero positions. The center frequency curve is also shown to facilitate the design process by showing the position of the center and upper -10 dB frequency points of the response during the design process. The fractional bandwidth provided is fixed at about 52.5% for a single element of the unit-cell for the whole range of frequencies shown. It is then easy to find the appropriate length that will provide the required lower -10 dB or center frequency points of interest, according to the design requirements.

This curve is easily reproducible for different ϵ_r by the full-wave simulation of different lengths versus the given ϵ_r .

Figure 4.3: PSL unit-cell length versus -10 dB transmission frequency for $\epsilon_r = 10$.

As a starting point to design the cascaded structure, independent elements of the unit-cell are to be designed first, prior to cascading. Later, they are cascaded leaving a small gap g in between, which acts as the matching part which slightly lengthens the total structure keeping the overall response unaffected. Its electrical length is $\theta_t = 90^\circ$ at the highest frequency among the two resonators. This rule applies to each of the air and substrate regions such that the response of the matching part acts as an all pass for bandstop region without disrupting the response. Moreover, it has an impedance of $Z_t = \sqrt{Z_1 Z_2}$ where Z_t is the impedance of the transformer, Z_1 is the circuit impedance seen from the left side of the transformer, and Z_2 is the circuit impedance seen from the right side of the short resonator. As long as both impedances Z_1 and Z_2 are similar

since they depend only on W , h , and ϵ_r , the calculated Z_t for each of the substrate and air regions would be identical to the previously calculated Z_s and Z_a , respectively.

4.1.2.2 Simulated and Measured Results

Similar to the previous structure, the simulation results are obtained using ADS package for circuit simulation, and CST package for full-wave simulation. The computer used is the same. Fig. 4.4 shows the model used in CST, with the following differences in simulation setup parameters:

1. Average number of tetrahedral mesh cells (tetrahedrons) that have been adaptively meshed through multiple passes is 16,500
2. Average time taken for each simulation is 4 minutes and 20 seconds
3. Peak memory used on average in each simulation is 328 MB

For our design example, a target bandwidth of 100% is to be rejected, starting from 5.6 GHz at -10 dB of the transmission curve. The design specifications are shown in Table 4.1. The two resonators are individually designed such that their transmission

Figure 4.4: 3-D model designed in CST simulation package.

curves are placed in an adjacent manner as shown in Figure 4.5. Note that the ratio $\frac{h}{p}$ is 0.66 here and the bandwidth targeted is the maximum this structure can exhibit.

Table 4.1: Circuit and physical parameters of the cascaded bandstop structure.

Circuit parameters: C/pF, Z/ Ω , θ /GHz@180 $^{\circ}$							
C_{a1}	C_{a2}	C_{s1}	C_{s2}	Z_o	Z_a		
0.0087	0.0086	0.0179	0.0161	443.52	161.884		
Z_s	θ_{a1}	θ_{a2}	θ_{s1}	θ_{s2}	θ_{ta}/θ_{ts} @80 $^{\circ}$		
89.965	30	16.667	9.583	5.324	30/9.583		
Physical Dimensions / mm							
ϵ_r	s	g	h	p	S1	S2	W
9.8	0.3	2.5	1.27	2	9	5	2

Figure 4.5: Full-wave vs circuit model transmission response for two independently designed PSL unit-cells before cascading.

The curves from both circuit and full-wave simulations show good agreement, especially for the lower part of the band. Slight frequency shifting appears in the upper part since the capacitors in the circuit model are considered perfect and fixed, while they are

variable with the frequency.

Figure 4.6: Full-wave vs circuit model transmission response of the cascaded bandstop structure proposed.

In circuit simulation, which was carried out using ADS, the gap may also be regarded as two physical transmission lines, each with an initial length equals to half of the shortest strip. The gap can then be optimized using circuit or full-wave simulation to obtain the best overall performance. Figure 4.6 shows the overall response after cascading using both circuit model and full-wave simulation after optimizing g .

Since one of the goals of our proposed designs is to obtain a stable performance under a large variation of oblique incident angles, a simple evaluation was performed on the PPW structure proposed in [11], which involved vertically stacked shielded microstrip lines rather than PSLs. Figure 4.7 demonstrates the performance under oblique incidence for the same proposed structure, using both shielded microstrip lines and PSLs.

Figure 4.7: Response under oblique incidence for the cascaded bandstop structure based on microstrip line, using full-wave simulation.

It is clear that the higher part of the band deteriorates significantly under oblique incidence when using microstrip lines. Therefore, a PSL structure with separation gap s is crucial in obtaining a better angular stable response. The gap s allows the current to be confined within each of the PSL unit-cells, and partitions the unit-cells into smaller individual elements as seen by the incident wave for.

Lastly, and after introducing the separation gap s , the structure is simulated using HFSS and tested under large oblique incident angles. The result is shown in Figure 4.8. The structure exhibits a fractional bandwidth of 100% at $S_{12} \leq -10$ dB, ranging from 5.71 to 17.17 GHz. Under oblique incidence, it remains lower than -10 dB over most of the band. At around 7.5 GHz, it deteriorates to -8.35 and -6.8 dB under 40° and 60° incidences, respectively.

Figure 4.8: Full-wave simulated response of the optimized cascaded bandstop structure based on PSL under oblique incidence.

Moreover, the small dip of -1.6 dB in the reflection response at 9.18 GHz is attributed to a small conductance that generates between the parallel strips, according to the circuit model in page 179 in [51], which could be simulated as parallel resistors of very high values. The transmission response has no peak at that point, asserting that the power is being absorbed rather than partially transmitted, leaving the stop band undisturbed since the transmission response is of utmost importance in this structure.

As for the out-of-band performance, reflection ears rise and the transmission worsens while oblique incident angles increase. The port impedance Z_o is a function of the angle of incidence $Z_{oblique} = Z_o / \cos(\theta)$, where θ is the angle of incidence. Higher angles of incidence increase the port impedance, thereby introduce a larger impedance mismatch. Furthermore, under oblique incidence, the propagation constant also varies for both of the air and substrate regions. Yet, it varies almost constantly for both regions with

fundamental transmission zeros, which is why the zeros are almost fixed under large oblique incidences. Under the normal incidence, the highest out-of-band insertion loss occurs at 2.47 GHz and reaches -0.8 dB at worst. While under the incidence angles 20° , 40° , and 60° , it deteriorates to -0.93, -1.5, and -3.15 dB at worst, respectively.

4.2 Fabrication and Measurement

(a) A strip of adjacent unit-cells

(b) Backbone comb structure

(c) Assembled structure

Figure 4.9: Fabricated cascaded structure prototype showing individual parts and the final assembly

The fabrication and measurement are both carried out in the same fashion as those of the previous structure in Section 3.3, except that the measurement here is slightly more difficult as two pairs of horn antennas have been used to measure the entire band. The

Figure 4.10: Full-wave simulated vs measured transmission response of the cascaded bandstop structure under normal incidence.

Figure 4.11: Full-wave simulated vs measured transmission response of the cascaded bandstop structure under 20° incidence.

final individual and assembled structures are shown in Figure 4.9. A good agreement can be seen in Figure 4.10 when the structure is tested under the normal incidence. Oblique

Figure 4.12: Full-wave simulated vs measured transmission response of the cascaded bandstop structure under 40° incidence.

Figure 4.13: Full-wave simulated vs measured transmission response of the cascaded bandstop structure under 60° incidence.

incidence tests were performed up to 60° since the structure is tolerant to high oblique incidences, and the results are shown in Figures 4.11, 4.12, and 4.13. Under 60° oblique

incidence, it is slightly difficult to get an accurate reading since the tilted window is not big enough to keep the space between the two horns blocked, besides some significant blockage introduced by the absorbing material. Thus, the stop band is slightly higher in magnitude than it should be since the structure is expected to show good rejection performance even under 60° , according to the full-wave simulation.

4.3 Summary

A cascaded structure with a large fractional bandwidth of 100% has been proposed, featuring a unit-cell size of $0.08\lambda_o$, and thickness of $0.63\lambda_o$ at the center frequency of operation. It offers a good out-of-band response under both normal and oblique incidences. The structure is mechanically stable, easy to assemble, and features very small unit-cell sizes as seen by the incident waves, rendering it stable under large incident angles. Design procedures have been provided with the necessary equations and curves to facilitate the design process. Designs with various relative bandwidths ($>52.5\%$ and $\leq 100\%$) can be provided and for a wide range of frequencies, according to the requirements. In terms of bandwidth, out-of-band performance and stability, this structure outperforms the alternatives available in the literature, according to the author's knowledge.

The structure is designed to cover this range and avoid higher frequencies that have the potential to raise accuracy issues due to small fabrication errors, higher dielectric constant losses, and increased far field distances which can complicate the measurement process.

The design process is simplified and the initial simulation results are made possible

using a transmission line/circuit simulation package before the optimization, which is carried out using full-wave simulation.

Chapter 5

Angular-Stable Polarization Rotator

5.1 Introduction

Polarization rotating structures were briefly reviewed in Chapter 2, and were found to exploit a multitude of layers, each designed independently to contribute in the rotation. Another solution employed an SIW structure and was found to exhibit performance limitations in terms of angular stability, though it solved the multi-layer requirement in the design. Therefore, a new variation of polarization rotator is proposed in this chapter, based on the fundamental bandstop unit-cell design proposed in the previous chapters. The rotator exploits the inherently stable bandstop structure by adding an extra function for linear polarization rotation, which is achieved by manipulating the E-field using an L-shaped slot etched on one of the conductor layers to generate a 90° rotated E-field component at the output.

What distinguishes this structure from the previously proposed bandstop solutions is that the unit-cell is considered as a PPW instead of a PSL. Small and individual unit-cells are not necessary since the angular stability will still be maintainable as the bandwidth targeted is relatively narrower, as well as to facilitate the design.

An example design is presented here, operating at 10 GHz as a center frequency, with a half-power (3 dB) fractional bandwidth of 2.9% for the transmission coefficient of the orthogonal component. The insertion loss is 1.3 dB at the resonant frequency. The proposed rotator exhibits good stability over a large range of oblique incidence and it solves the multi-layer requirement with a relatively thin profile ($<0.25\lambda_o$). The operating principle of the L-slot is explained, along with an illustration of the field behavior, and the design details are briefly discussed. The presented design operates in the X-band, exhibiting an angular stable response under large oblique incidences, besides a simple design procedure.

5.2 Description of the Structure

The design follows the principle developed in [11] to realize a dual-mode resonator to achieve band rejection. The dual-mode resonator is modified for a simpler design and for accommodating the slot, which is done by extending the strip to be as wide as the ground, realizing a simple PPW. An equivalent circuit model can be established and the circuit parameters can be calculated in a similar fashion as in Subsection 3.2.2.1. The structure is shown in Fig. 5.1. Similar to the previous structures, two fundamental TEM modes are excited between the two parallel plates: one passes through the substrate region and the other through the air region. They both realize the band reject performance, which basically restricts the transmission of the co-polarized E-field component, which

is orthogonal to the (x-y) plane of the PPW array.

Figure 5.1: Geometry of the proposed polarization rotator.

An additional L-shaped slot is etched on one of the PPWs such that proper slot lengths are chosen as follows, while maintaining a slot width of 0.3 mm:

$$H\text{-slot} + V\text{-slot} = \frac{\lambda_g}{2} = \frac{\lambda_o}{2\sqrt{\epsilon_{eff}}} \approx \frac{\lambda_o}{2\sqrt{\epsilon_r}} \quad (5.1)$$

where λ_g is the guided wavelength of the E-field at the point where the H-slot intercepts the field. As shown in the equation, the effective dielectric constant that is apparent to the propagating field at the area where the slot is etched is almost equal to the relative dielectric constant of the substrate material itself. Thus, calculating the approximate length of the slot becomes easy using the above equation. Optimization is then done to tune the values of *H-slot* and *V-slot* such that the desired performance is attained.

5.3 Design Example

5.3.1 Theory and Procedures

As mentioned before, the dual-mode resonator bandstop structure which the rotator exploits is designed according to the same procedures in Chapter 3. A relatively high $\varepsilon_r = 9.8$ is used in order to yield a thin profile. The thickness of the structure's profile (L) is less than $0.25\lambda_o$ at 10 GHz and is chosen according to Figure 3.10. An L-shaped slot is etched on one of the two plates, as shown in Fig. 5.1. Its purpose is to rotate the field that is coupled to L-slot's horizontal part $H - Slot$ by 90° , so that it is later re-radiated at the open end (aperture) of L-slot's vertical part $V - Slot$.

The polarization rotation effect is shown in Figure 5.2, where the E-field behavior is depicted in the air, substrate and conductor planes. The slot introduces a vertical E-field transmission zero at 10 GHz, then transforms it into a horizontal E-field transmission pole at the given frequency. Note that the vertical E-field polarization lines are aligned

Figure 5.2: Distribution of the E-field in the x-z plane for different regions of the unit-cell.

with the y-axis, and the horizontal E-field polarization is aligned with the x-axis. It is worth emphasizing that the source of the re-radiated rotated component is neither of the slots, but the aperture (open end) at the end of the *V – Slot* and as shown in Fig. 5.2. The stop band exhibited by the PPW also means that no radiation of the co-polarized field by the PPW itself is possible. In order for the rotator to operate at 10 GHz, Table 5.1 lists the values chosen for the design to achieve the target performance.

5.3.2 Simulation Results

Figure 5.3: 3-D model designed in CST simulation package.

As it was with the two previous structures in Sections 3.2.2.2 and 4.1.2.2, the simulation results are obtained using ADS package for circuit simulation, and CST package for full-wave simulation, using the same computer system. Fig. 4.4 shows the model used in CST, with the following differences in simulation setup parameters:

1. Gaussian input pulse that extends from 7 to 13 GHz
2. Average number of tetrahedral mesh cells (tetrahedrons) that have been adaptively meshed through multiple passes is 11,300
3. Average time taken for each simulation is 3 minutes and 30 seconds

4. Peak memory used on average in each simulation is 256 MB

Simulation results are split into two categories: one is concerned with the behavior of the slot's dimensions and their effect on the response and the other provides the optimized results.

5.3.2.1 Slot Related Parametric Studies

Multiple variables have a direct effect on the slot's performance, including: $H - Slot$, $V - Slot$, s , ϵ_r , h , and $p - h$ (air gap). Since ϵ_r , h , and $p - h$ have also been involved as crucial variables in designing bandstop response of the structure, they are not considered here in the slot's parametric studies. Only $H - Slot$, $V - Slot$, and s are considered after realizing the desired bandstop performance. Thus, a series of parametric studies have been conducted to observe the relative performance difference caused by these fundamental dimensions.

First, the effect of the parameter W on the performance is studied, and the results are shown in Figure 5.4. The lowest value the parameter can have is 3.8 mm since $H - slot$ is set to 2.7 mm in this study. The value of $V - slot$ is 0.45 mm. Varying W would mean varying the separation distance s between each $H - slot$ in the periodic structure. The result does not change for this variation and the center frequency of operation is almost fixed. In theory, the width W of the unit-cell does not affect the frequency of operation of the co-polarization component as the unit-cell can be designed to have virtually any width, which keeps $S_{21}(V)$ fixed too.

Second, the effect of the parameter $H - slot$ is tested and illustrated in Figure 5.5. It

Figure 5.4: Parametric study showing the effect of unit-cell width (W) on the structure's response.

shows that increasing the horizontal length of the slot will cause a drop in the slot's center frequency of the operation. It should be noted that the cross-polarization component excited can weaken the co-polarization component such that it helps to maintain a better cross-polarization isolation. A suitable value was chosen to be 3.7 mm, which centers the cross-polarized component in the middle of the co-polarization transmission band. The study shows that shifting in the response is highly sensitive to a small change in the horizontal slot length, which is intuitive.

Third, another important study is made on the slot dimension V-slot, which shown in Figure 5.6. It also has a crucial role in deciding the center frequency of operation of the cross-polarization component. Similar to the previous case with H-slot, the point at which the cross-polarization component is excited cause a drop in the co-polarization

Figure 5.5: Parametric study showing the effect of H-slot on the y-polarization (vertical, co-polarized) and x-polarization (horizontal, cross-polarized) E-field components. V-slot is fixed at 0.45 mm.

component such that it enhances the isolation. The same amount of change in the slot's length will correspond to almost similar differences in frequency shifting when compared with the case of H-slot, showing a significant contribution to the frequency response of the slot.

Finally, a study was conducted on the length of the substrate, considering special fabrication conditions that may need to be considered. It is expected that the substrate may be designed with a length value that exceeds that of the parallel strip lines (conductor layers) to facilitate the fabrication process and prevent trimming the L-slot due to the milling machine's tolerance issues. The study shows that when the substrate becomes longer, the response tends to shift downwards. This causes the field to travel

an extra distance inside the higher dielectric constant medium, resulting in a lower frequency component at the output.

Figure 5.6: Parametric study showing the effect of V-slot on the y-polarization (vertical, co-polarized) and x-polarization (horizontal, cross-polarized) E-field components. H-slot is fixed at 3.7 mm.

5.3.2.2 Optimized Results

As seen before, the resonant frequency of the rotated E-field component is affected by multiple parameters, mainly $H - Slot$, $V - Slot$, and s (or W). The optimized values for L-slot's dimensions are chosen as: $H - Slot = 3.7$ mm, $V - Slot = 0.45$ mm, $s = 0.15$ mm, which is achieved by having $W = 3.85$ mm. Adding $V - Slot$ and $H - Slot$ yields almost $0.5\lambda_g$ (guided wavelength) at 10 GHz and as shown in Equation 5.1. Note that the $V - Slot$ is kept open at its end to re-radiate the rotated field component. The final design parameters are tabulated in Table 5.1

Table 5.1: Physical parameters of the polarization rotator.

Unit-cell's physical parameters / mm				
ϵ_r	L	W	h	p
9.8	7.3	3.85	1.27	1.9
Slot parameters / mm				
$H - slot$	$V - slot$	Slot thickness		s
3.7	0.45	0.3		0.15

Figure 5.7: Parametric study showing the effect of the length of the substrate L_{sub} on the y-polarization (vertical, co-polarized) and x-polarization (horizontal, cross-polarized) E-field components.

The simulated response is shown in Fig. 5.8. The insertion loss achieved is 1.23 dB at 10 GHz with a 3 dB fractional bandwidth of 2.9%. A slight shift in the center frequency of the stop band response shows when the structure is with and without the L-slot. This can be attributed to the slight change in ϵ_{eff} of the substrate filled PPW since the etched slot exposes an area of the PPW to the air filled region, lowering the overall ϵ_{eff} , and

Figure 5.8: Simulated reflection and transmission coefficients of the polarization rotator with and without the L-slot. V and H are the y-polarization (vertical, co-polarized) and x-polarization (horizontal, cross-polarized) E-field components, respectively.

thereby the coupling between the plates compared to the case without the L-slot when $\epsilon_{eff} = \epsilon_r$ and the substrate region is homogeneously filled with the dielectric material.

The performance against a wide range of oblique incident angles is shown in Fig. 5.9. The transmission pole is stable under 25° incidence. Slight shifting by around $+0.065$ GHz starts at 50° , yielding a 1.7 dB insertion loss for the rotated E-field component at 10 GHz, indicating good tolerance toward a large variation of oblique incidence angles. It should be noted that this structure is single polarized and the performance may deteriorate if the incident E-field is not normal.

The advantage of embedding the L-slot into this bandstop structure is to efficiently isolate the two orthogonal E-field components since the cross-polarized component arises

Figure 5.9: Simulated reflection and transmission coefficients of the polarization rotator under oblique incidence. V and H are the y-polarization (vertical, co-polarized) and x-polarization (horizontal, cross-polarized) E-field components, respectively.

within a stop band of the co-polarized component. The cross-polarization isolation under the normal incidence at 10 GHz is 19.67 dB, and it decreases to 15.78 and 7.66 dB under 25° and 50° , respectively. It is worth noting that the cross-polarization isolation can be further tuned by changing the ratio $\frac{h}{p}$. When the air gap is reduced, higher rejection magnitude with steeper skirts and lower overall -10 dB bandwidth can be obtained, yielding higher cross-polarization isolations under both normal and oblique incidences.

5.4 Fabrication and Measurement

The fabricated prototype is shown in details in Figure 5.10. Fabrication tolerances impose a slight effect on the performance, especially for filing/cutting the substrate ma-

Figure 5.10: Fabricated polarization rotator showing individual parts and the final assembly

terial accurately. The measurement is carried out using a similar setup as that used for the previous structures in this thesis, except that the horn antenna is rotated by 90° when receiving the rotated (cross-polarization) field. As it can be seen in Figures 5.11, 5.12 and 5.13, a good bandstop response is obtained in the co-polarized position, providing an improved cross-polarization isolation. It is noticeable that the frequency at which the cross-polarization component is excited is shifted downwards compared to the simulation results, which was explained using the parametric study in Section 5.3.2.1 in Figure 5.7.

Figure 5.11: Simulated reflection and transmission coefficients of the polarization rotator under oblique incidence.

Figure 5.12: Simulated reflection and transmission coefficients of the polarization rotator under oblique incidence.

Figure 5.13: Simulated reflection and transmission coefficients of the polarization rotator under oblique incidence.

5.5 Summary

An angular stable polarization rotator is proposed, based on a 3D bandstop structure, which was pioneered by our lab recently. The design concept is based on a PPW and its design procedure has been simplified to be ready for circuit simulation to obtain initial results, as explained in the previous chapter. The polarization rotation is achieved by employing an L-shaped slot, which serves to trap the co-polarized E-field passing through both the air and substrate regions of the PPW, then rotates it and re-radiates it from the open end of the L-slot. The insertion loss under normal incidence is 1.2 dB. Higher oblique incidence angles tend to introduce an impedance mismatch at the port side since its impedance is a function of the angle of incidence, thereby increasing the insertion loss under oblique incidence and also decreasing the cross-polarization isolation since the co-polarized component tends to increase in magnitude. The highest insertion loss is 1.7

dB at the center frequency of operation, 10 GHz. The unit-cell thickness is less than $0.25\lambda_o$.

Chapter 6

Conclusion and Future Work

6.1 Conclusion

This thesis focuses on exploiting a three-dimensional frequency selective structure that exhibits pseudo-elliptic filtering response to design novel ultra wide bandstop and polarization manipulation structures with stable performance under a large variation of oblique incident angles using various techniques. A comprehensive literature review have been conducted on wide band bandpass and bandstop filters and polarization rotators, revealing that it is a challenging task to design high performance ultra wide band filters using the few concepts studied in the literature to attain wide bandwidths with sharp selectivity and good angular stability over the entire band. Similarly, polarization rotators that are based on SIW cavities offer relatively angle dependent performance. In this thesis, we have presented three novel structures, containing two ultra-wide bandstop filters and one polarization rotation frequency selective structure, which can tackle the performance and design complexity limitations. They could outperform the current state of the art designs, according to our literature review, keeping robust mechanical characteristics and fabricated using the well established PCB technology.

Two ultra-wide bandstop structures are proposed in Chapter 3 and Chapter 4 using two different methods. Both exploit the same unit-cell of the array of vertically stacked shielded microstrip lines proposed earlier in the literature. Simplification is done on the design by transforming the unit-cell into two parallel strip line waveguides, each filled with a different medium having a different dielectric constant. This helps to provide a systematic approach to design the filter and allows the initial simulation to be carried out using the circuit model approach, and later comes the optimization using full-wave simulation, if necessary.

The first method involves introducing a transmission zero of a higher order resonance to combine it with the two pre-existing zeros to widen the rejection band. The ultimate fractional bandwidth is extended from 52.5% to 78% only by choosing an appropriate dielectric constant value for the substrate used, yielding a wider bandwidth, which happened to require a relatively cheaper laminate. The approach yields a good out-of-band response for both the lower and upper regions. However, its performance is less immune to oblique incidence because of the existence of a higher order resonance, which is more sensitive compared to the fundamental resonance. The overall thickness of the structure is $0.5 \lambda_o$ at the center frequency of operation. It is designed to cater for applications with small incident angle variations.

The second method involves a cascade of two independently designed parallel strip line waveguides with a certain gap (dielectric section) designed to separate the two resonators and ensure proper matching. The fractional bandwidth acquired is 100% and exhibits a stable performance under highly oblique incidences, providing a good solu-

tion to overcome the stability issue when ultra-wide bandwidths are considered since the unit-cell size as seen by the incident field is $0.08\lambda_o$. The thickness is $0.63\lambda_o$, which is slightly larger than that in the previous design. The in-band insertion loss over the whole range of operation is as low as -10 dB under normal incidence for both designs, while it reaches -8.35 at 40° and -6.8 at 60° . The out-of-band insertion loss is low under normal incidence. However, it starts to increase while oblique incident angles increase, reaching -1.8 at 40° and -3.15 dB at 60° .

An angular stable linear polarization rotating frequency selective structure is proposed, deploying a PSL unit-cell with strip width equivalent to the unit-cell width, with an added L-shaped slot such that the field passing through the air and substrate regions is trapped and rotated. A 90° E-field cross-component is re-radiated from the open end of the L-slot. Since the structure of this variant exhibits a bandstop response with the co-polarized field, while it excites a cross-polarization field at 10 GHz with a -3 dB bandwidth of 2.9%, it helps to enhance the polarization isolation at the center frequency. The response shows tolerance under a wide range of oblique incident fields with +0.065 GHz shifting at 50° at the center frequency. The insertion loss exhibited is 1.23 and 1.7 dB under normal and 50° incidences, respectively. Cross-polar isolation is 19.67, 15.78 and 7.66 dB under normal, 25° and 50° incidences, respectively.

Conclusively, the first challenge of having an ultra-wide band structure has been tackled in two different ways, each featuring a different level of tolerance to oblique incidence and a different substrate material. The last challenge of providing a stable and simple polarization rotator has been overcome as well by exploiting a PPW unit-cell and etching an L-shaped slot on one conductor layer. A comparison of performance and

features of all our proposed structures with their counterparts in the literature have been made in Table 6.1 and Table 6.2.

6.2 Future Work

1. A cascaded structure can be considered while exploiting a higher order resonance at the same time so that a wider bandstop response is acquired. The dielectric constant of the substrate should be significantly smaller as well, yielding a more cost effective structure with an even wider coverage.
2. Conformal variations of these designs could also be studied so that they could be stacked at the inner walls of dielectric radomes to filter/rotate the E-fields transmitted by their integrated antennas. The tolerance to oblique incidence and sharp filtering characteristics qualify them to realize high performance conformal radomes.
3. Using the cascaded structure, bandpass designs can be investigated by separating the responses of each of the individual resonators such that a pass band forms in between. This technique could also help to realize multi-band bandstop filters.

Table 6.1: Performance comparison of our two proposed bandstop structures vs. those in the literature.

Feature	Extruded [18]	Cylindrical [12, 19]	Square cylinder[20]	Spring [21]	4-legged loop[33]	3-Zeros Fig. 3.1	Cascaded Fig. 4.1
<i>Unit-cell size</i> / λ_o	19.5%	79%	88%	40%	20%	7%	8%
<i>Thickness</i> / λ_o	2.4%	50%	20%	16%	4.8%	50%	63%
<i>%BW/-10 dB</i>	50%	112%	18%	28% @ -20 dB	10%	78%	100%
<i>Angular performance</i>	Very good up to 60°	-	Good up to 45°	Angle dependent	Very good up to 60°	Good up to 40°	Very good up to 60°
<i>Selectivity</i>	Smooth	Sharp	Smooth	Smooth	Smooth	Elliptic	Elliptic
<i>Nearn out-of-band loss at 0°</i>	Spurious stop bands	High insertion loss (>10 dB)	High insertion loss (>10 dB)	High insertion loss (>10 dB)	≈ 2 dB	0.64 dB	0.8 dB
<i>Response (upper-lower) symmetry</i>	Symmetric	Symmetric	Symmetric	Symmetric	Asymmetric	Symmetric	Symmetric

Table 6.2: Performance comparison of our proposed polarization rotator vs. those in the literature.

Feature	Wire grid#1 [24]	Wire grid#2 [25]	SIW-based [14]	Our rotator Fig. 5.1
<i>Unit-cell size</i> / λ_o	-	14%	67%	13%
<i>Thickness</i> / λ_o	>75% (seven layers)	$\approx 23\%$ (three layers)	18.3%	25%
<i>%BW/-10 dB</i>	42%	120%	9.1%	2.9% @ -3dB
<i>Angular performance</i>	Very good up to 60°	-	Up to 20°	Very good up to 50°
Insertion loss	0.1 dB	<0.5 dB	0.2 dB	1.7 dB

List of Publications

- [1] A. Qasim and Z. Shen, "Design of Wideband Bandstop Frequency-Selective Structures Using Stacked Parallel Strip Line Arrays," *IEEE Trans. Antennas Propag.*, 2015. (Submitted)
- [2] A. Qasim and Z. Shen, "Angular-stable polarization rotator," *IEEE Antennas and Propagation Society International Symposium (APSURSI)*, 2014.
- [3] A. Qasim and Z. Shen, "Ultra-wide bandstop frequency selective structure using a cascaded array of vertically stacked parallel strip lines," *IEEE International Conference on Computational Electromagnetics (ICCEM)*, 2015.
- [4] B. Li, Y. Tang, A. Qasim, and Z. Shen, "Implementation of three-dimensional band-pass frequency selective structure using multilayer PCB technology," *IEEE Antennas and Propagation Society International Symposium (APSURSI)*, 2015 (accepted).

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