Improved 3-D frequency-selective structures and their applications

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2019

Ahmed Abdelmottaleb Abdelfattah Attia Omar. (2019). Improved 3-D frequency-selective structures and their applications. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/88524>

<https://doi.org/10.32657/10220/47615>

**NANYANG
TECHNOLOGICAL
UNIVERSITY**

SINGAPORE

**IMPROVED 3-D FREQUENCY-SELECTIVE
STRUCTURES AND THEIR APPLICATIONS**

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OMAR**

**SCHOOL OF ELECTRICAL & ELECTRONIC
ENGINEERING**

2019

IMPROVED 3-D FREQUENCY-SELECTIVE STRUCTURES AND THEIR APPLICATIONS

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School of Electrical and Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2019

In Memory of my father,

Abdelmottaleb

To my mother, Mervana, my wife, Sarah, and my daughter, Maria

STATEMENT OF ORIGINALITY

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

10-01-2019

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SUPERVISOR DECLARATION STATEMENT

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

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[Prof. Zhongxiang Shen]

AUTHORSHIP ATTRIBUTION STATEMENT

This thesis contains material from 5 papers published in the following peer-reviewed journals where I was the first and/or corresponding author.

Chapter 3 is published as **A. A. Omar** and Z. Shen, “Multiband high-order bandstop 3-D frequency-selective structures,” *IEEE Trans. Antennas Propag.*, vol. 64, no. 6, pp. 2217–2226, Jun. 2016. DOI: 10.1109/TAP.2016.2546967.

The contributions of the co-authors are as follows:

- Prof. Shen provided the initial research direction and edited the manuscript.
- I prepared the manuscript, simulate the proposed structures, and measured the fabricated prototypes.
- I designed the proposed structures and analyzed the results.
- Prof. Shen helped with the analysis of the results.

Chapter 4 is published as **A. A. Omar** and Z. Shen, “Thin bandstop frequency-selective structures based on loop resonator,” *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 7, pp. 2298–2309, July. 2017. DOI: 10.1109/TMTT.2017.2651812.

The contributions of the co-authors are as follows:

- I prepared the manuscript, simulate the proposed structures, and measured the fabricated prototypes.
- Prof. Shen edited the manuscript.
- I designed the proposed structures and analyzed the results
- Prof. Shen helped with the analysis of the results.

Chapter 5 is published as **A. A. Omar** and Z. Shen, “Thin bandpass 3-D frequency-selective structure based on folded substrate for conformal radome application,” *IEEE Trans. Antennas Propag.*, vol. 67, no. 1, pp. 282–290, Jan. 2019. DOI: 10.1109/TAP.2018.2876706.

The contributions of the co-authors are as follows:

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- I prepared the manuscript, simulate the proposed structures, and measured the fabricated prototypes.
- Prof. Shen edited the manuscript.
- I designed the proposed structures and analyzed the results.
- Prof. Shen helped with the analysis of the results.

Chapter 6 is published as **A. A. Omar** and Z. Shen, “Double-sided parallel-strip line resonator for dual-polarized 3-D frequency-selective structure and absorber,” *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 10, pp. 3744–3752, Oct. 2017. DOI: 10.1109/TMTT.2017.2700301.

The contributions of the co-authors are as follows:

- I and Prof. Shen provided the idea of the proposed structures.
- I prepared the manuscript, simulate the proposed structures, and measured the fabricated prototypes.
- Prof. Shen edited the manuscript.
- I designed the proposed structures and analyzed the results.
- Prof. Shen helped with the analysis of the results.

Chapter 7 is published as **A. A. Omar**, Z. Shen, and H. Huang, “Absorptive frequency-selective reflection and transmission structures,” *IEEE Trans.*

Antennas Propag., vol. 65, no. 11, pp. 6173–6178, Nov. 2017. DOI: 10.1109/TAP.2017.2754463.

The contributions of the co-authors are as follows:

- Prof. Shen provided the initial research direction and edited the manuscript.
- I prepared the manuscript, simulate the proposed structures, and measured the fabricated prototypes.
- Mr. Huang revised the manuscript and helped with the initial idea of the manuscript.
- I designed the proposed structures and analyzed the results.

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ACKNOWLEDGMENTS

I would like to thank my supervisor, Prof. Shen Zhongxiang, for his strong support and encouragement during the past four years. I am thankful for his efforts to guide me and provide me with valuable discussions and critical suggestions.

I would also like to thank all colleagues in Communication Research Lab II, especially every group member under Prof. Shen, for engaging in technical discussions and helping in measurements.

My thanks are also due to our technical staff, Ms. Thung Lina, Mr. Ng Teng Kwee, Ms. Teng Siew Juan Catherine, and Ms. Foo Wenney, for their assistance regarding our lab arrangements, document requests, and measurements.

In addition, I would like to acknowledge the SINGA Scholarship and Nanyang Technological University for providing the research scholarship for my Ph.D. study.

I would also like to thank the anonymous examiners for their positive comments and valuable suggestions that improved the quality of the thesis.

My special gratitude goes to my family, especially my mother, Mervana, my wife, Sarah, my daughter, Maria, my sisters, and my brothers. I could not have completed this work without their endless and unconditional love and support.

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ABSTRACT

Frequency-selective surfaces are 2-D infinite arrays that act as a filter for the incident electromagnetic waves. They have been used in hybrid radomes, dichroic main reflector, sub-reflector, microwave absorbers, absorptive frequency-selective surfaces, high impedance surfaces, and many other applications. These 2-D arrays consist of slots in a conducting plate or conducting strips printed on a dielectric layer, which are able to produce bandpass or bandstop frequency response, respectively. The traditional 2-D frequency-selective surface suffers from some drawbacks, such as poor selectivity and for some cases poor angular stability under oblique incidence. Moreover, it is difficult to design high-order bandstop 2-D frequency-selective surface with good response and by adding the multiband response on the design requirement, the overall design becomes sophisticated. For absorber design, it is very challenging to achieve a wide absorption band with a thin structure. The reported absorptive frequency-selective transmission structures have narrow absorption bands, which limits their radar cross-section reduction bandwidth. Recently, 3-D frequency-selective structure was presented in the literature. By employing the 3-D concept, a multiband high-order bandstop frequency-selective surface, a wideband absorber with a thin structure, and an absorptive frequency-selective transmission structure with wide absorption bands are accomplished. Moreover, some limitations associated with the 3-D frequency-selective structure are solved.

First, a general and systematic methodology is proposed to design multiband high-order bandstop 3-D frequency-selective structures. By adjusting the coupling between identical resonators, the high-order response is achieved. Through introducing multiple resonators in a single unit cell, one can readily obtain a multiband filtering response. Detailed analysis of the coupling between identical resonators is provided, so are equivalent circuit models to explain the operating principle. Two structures are designed, fabricated, and measured to verify the proposed methodology. The first structure is a tri-band second-order

frequency-selective structure, while the second structure exhibits a dual-band third-order response.

Although the 3-D frequency-selective structure has a good performance, it has two problems: one is the thickness problem, which may limit its practical applications, and the other is the operation under a dual-polarized incident wave. A loop resonator is then proposed in the unit cell of the bandstop structure instead of microstrip resonator aiming to achieve maximum reduction in the structure thickness. Using the elliptical loop with stepped-impedance resonator, an 80% reduction in structure thickness is achieved without degrading the performance compared to the original structure, which makes it suitable for practical applications.

In addition, the folded substrate technique is employed for the bandpass frequency-selective structure to reduce its thickness. A 67% and 79% thickness reduction compared to the basic bandpass 3-D frequency-selective structure is achieved by employing three- and five-layer folded substrate, respectively. Moreover, a half-cylindrically shaped radome as an example of a conformal frequency-selective surface is constructed based on the proposed technique and integrated with a broadband horn antenna. Besides working as a radome it shows good filtering properties. A fabricated prototype of this radome is measured in the presence of a broadband horn antenna.

Furthermore, double-sided parallel-strip line resonator is employed in the unit cell of a 3-D frequency-selective structure in order to exhibit the desired response (bandstop, bandpass and absorption) under an incident wave of arbitrary polarization. It is found that this unit cell supports the propagation of three quasi-TEM modes. A dual-polarized bandstop 3-D frequency-selective structure is then proposed. Moreover, a dual-polarized absorber is also realized by employing a unit cell with double-sided parallel-strip line resonator. By properly exciting and absorbing the second mode of the modified unit cell, an ultra-wideband dual-polarized absorber is constructed.

As the last application example, which is absorptive frequency-selective transmission structure, a tailoring methodology is proposed to achieve arbitrary frequency response within a wide absorption band. An ultra-wideband absorber utilizing multiple resonances in the unit cell of 3-D frequency-selective structure is utilized as the textile for the proposed designs. Two structures are proposed; one is the absorptive frequency-selective reflection structure and the other is the absorptive frequency-selective transmission structure. Two prototypes are designed, fabricated, and measured to validate the proposed methodology. The first one is an ultra-wideband absorber, while the second one is an absorptive frequency-selective transmission structure with wide absorption bands at both sides of its transmission window. A good agreement is achieved between simulated and measured results.

Finally, a few suggestions and recommendations are made for further investigation. For example, a tunable absorptive frequency selective transmission is proposed, which may be integrated with a tunable antenna to achieve a low radar cross-section tunable antenna system. The structure employing double-sided parallel-strip line resonator may be suitable for transmit/reflect arrays as well.

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CHAPTER 1

INTRODUCTION

1.1 Background

Frequency-selective surfaces (FSSs) are two-dimensional periodic arrays that act as a filter for the incident electromagnetic waves. It is the wave version of the filters in electric circuits. It is characterized by its reflection and transmission coefficients as in filters but they have to fulfil the stability of the performance under various angles of incidence and different polarizations of the incident electromagnetic field. One of the oldest FSSs was fabricated to replace the parabolic reflector [1]. It has been used in hybrid radomes, dichroic main reflector, sub-reflector, microwave absorbers, absorptive FSSs, high impedance surfaces, and many other applications. Moreover, it can be integrated with antennas to achieve unique characteristics to be presented below.

Hybrid radomes are one of the fundamental applications of the FSSs in which they are integrated with radomes to reduce the monostatic radar cross-section (RCS) of the antenna outside its operating band. Fig. 1.1 shows an antenna mounted at the nose of an airplane and covered by a shaped radome with bandpass FSS, which reflects the out-of-band incident wave and transmits the in-band incident wave. When the FSS operates in its reflection mode, the wave is reflected away from the incoming direction due to the shape of the radome while the incident wave goes through the radome when the FSS becomes transparent [1].

Not only bandpass FSSs are used as radomes, but bandstop one could be also used. For example, it can be used in shipboard applications [1]. In such application, frequencies below 3 GHz are often used. If bandpass FSSs are employed, grating lobes may appear at high frequencies, which leads to high

Figure 1.1: Hybrid radome used to reduce antenna's out-of-band radar cross section [1].

Figure 1.2: Using frequency-selective surface as dichroic sub-reflector [1].

level of RCS at these frequencies. Therefore, it is better to utilize the bandstop one under such a situation.

Dual-band communication can be achieved using one main reflector and dichroic sub-reflector, which employs an FSS with bandstop response at a certain frequency and bandpass at a different frequency, as shown in Fig. 1.2 [1]. FSSs can also be employed with the main reflector, which is called dichroic

Figure 1.3: Circuit analog absorber along with its circuit model [1].

main reflector. By loading a Salisbury screen with bandstop FSS, as shown in Fig. 1.3, its thickness can be reduced with an enhanced bandwidth [1]. The enhanced bandwidth was accomplished by exploiting the feature of a series resonant circuit that acts as a capacitor before the resonant frequency and as an inductor after the resonant frequency to match the performance of the conductor-backed substrate.

FSSs can be also integrated with antennas in several forms. Bandstop FSSs were used instead of the ground plane to reduce the out-of-band RCS of the patch antenna [2]. Bandstop FSS reflects the in-band incoming signal, which is suitable to work as a ground plane for the antenna and passes the out-of-band incoming signal. Therefore, a low out-of-band RCS antenna is achieved. However, this feature vanishes if a metallic object is presented beyond the bandstop FSS. FSS was also used as a superstrate, wide bandwidth and high directivity was achieved [3]–[5]. In [3] an aperture coupled microstrip antenna was employed as a source and an FSS was used as a superstrate layer. Separate resonance frequencies and therefore a wide frequency band was accomplished with the proper use of the superstrate layer, microstrip patch and coupling aperture simultaneously. Furthermore, high directivity was achieved only by using the superstrate layer. Dual-band high directivity antenna was reported in

Figure 1.4: Principle of beamwidth control [7].

[6] by employing high-impedance surface in the ground and superstrate layer, which were based on FSS design. In this design, single-layer superstrate and artificial surface as ground plane were employed. Superstrate layer is enough to obtain the lower operating frequency band, but both the superstrate layer and artificial surface as a ground plane are necessary to generate the upper operating frequency band. Controlling beamwidth of antennas was presented [7] by utilizing active FSS as a superstrate, as illustrated in Fig. 1.4. Active FSSs were also used to electronically steer antenna beam either by controlling the transmission phase of the surface to achieve gradient phase distribution [8] using varactor diodes or employing reconfigurable bandstop FSS using PIN diodes [9, 10], as shown in Fig. 1.5.

By controlling the transmission/reflection phase of the incident wave at each unit cell of the FSS, transmit/reflect arrays were implemented [11 – 14]. Multiband bandstop FSSs may be also useful for suppression of antenna harmonics to reduce electromagnetic interference, especially in multifunction communication platforms.

This thesis is concerned with the frequency-selective surface/structure (FSS) design and its applications. The performance of the periodic structures such as FSSs can be determined from its unit cell response as stated by the Floquet theorem [15]. In our discussion, we focus on the unit cell characteristics. By employing a parallel resonance circuit between input and output ports, one can

Figure 1.5: Control methods for: (a) Single-beam; (b) Multi-beam [10].

achieve bandpass response at the resonant frequency of the circuit. While one can accomplish bandstop response if series resonance circuit is utilized. Previously in [1], the elements employed in the unit cell are of the resonant type. If one considers the shape, dipole and loop are considered as the main elements. The patch and slot elements are two types employed to achieve bandstop and bandpass responses, respectively.

1.1.1 2-D Bandpass Frequency-Selective Surface

The 2-D periodic array of slot element is employed to produce bandpass response and can be modeled as a parallel resonance circuit in which the capacitor represents the electric field in the slot and the inductor represents the current flow around the slot. Fig. 1.6 shows the 2-D bandpass FSS employing

Figure 1.6: Structure of bandpass frequency-selective surface. (a) 3-D array view. (b) Unit cell excitation and boundary conditions used in full-wave simulation. (c) Top view of a unit cell with element dimensions ($b = 16$ mm, $L_d = 14$ mm, $w = 0.4$ mm, $\epsilon_r = 1$).

crossed slot periodic array in vacuum operating at 10 GHz. Fig. 1.7 shows a comparison between the simulated reflection and transmission coefficients using the full-wave High-Frequency Structure Simulator (HFSS) with Master and Slave boundary conditions as the boundaries of the unit cell and Floquet ports as the excitation ports, as shown in Fig. 1.6(b), and those obtained from circuit model using Agilent Advanced Design System (ADS). Throughout the thesis, the full-wave HFSS and CST Microwave Studio (MWS) simulators are used as the full-wave simulation tools. Moreover, variables Θ and Φ are used to represent the angle of incidence while φ is used to represent the polarization of the incident electromagnetic wave.

Figure 1.7: Modeling of the bandpass frequency-selective surface. (a) Equivalent circuit model for the unit cell presented in Fig. 1.6. (b) Simulated electric field and surface current distribution at resonant frequency for TE polarization (E in the y -direction). (c) Simulated reflection and transmission coefficients using HFSS and ADS with $L = 0.588$ nH, $C = 0.4$ pF and $Z_0 = 377$ Ω .

Figure 1.8: Various element shapes used in the unit cell [1].

For the dipole element, the resonance occurs when its length equals to a half-wavelength of the incident electromagnetic wave. While for the loop, resonance occurs when its circumference equals to the wavelength of the incident electromagnetic wave. By adding a substrate, the resonant frequency is lowered depending on the dielectric constant value. There are a lot of element shapes that can be used in the unit cell [1], as shown in Fig. 1.8.

Based on the comparison made in [1, 16], the loop type may be the best element when the stability of performance under various angles of incidence is concerned. As shown in Fig. 1.9, the resonant frequency is stable under oblique incidence for TE polarization (E in the y -direction) and TM polarization (E in the x -direction) by employing slot loop as the element in a unit cell although the selectivity is not sharp.

Figure 1.9: Frequency-selective surface employing slot loop element. (a) Unit cell 2-D view, simulated reflection and transmission coefficients under oblique incidence for (b) TE polarization and (c) TM polarization ($b = 12$ mm, $r = 5$ mm, $w = 0.1$ mm, $\epsilon_r = 1$).

To improve the selectivity, the cascading technique [1] can be used, which leads to an increase of the filter order by employing direct coupling between resonators. As shown in Fig. 1.10, the selectivity is improved compared to the single sheet case but the response becomes sensitive to oblique incidence, as shown in Fig. 1.11, because of the large spacing between the two sheets. The equivalent circuit model of the cascaded case is shown in Fig. 1.10(b) with a comparison between the simulated results obtained using HFSS and ADS shown in Fig. 1.10(c). The sensitivity to oblique incidence is reduced by adding a substrate between these two arrays. However, the spacing between the arrays has to increase to obtain -10 dB reflection coefficient between the two resonant frequencies. By including the substrate, the periodicity reduces compared to the operating wavelength leading to the stability under oblique incidence. The problem is that as the dielectric constant increases, it is difficult to obtain -10 dB reflection coefficient between the two resonant frequencies regardless of the spacing between the two arrays. For the bandpass response, by adding two cover dielectric layers with a thickness of a quarter-wavelength and low dielectric constant, better stability can be obtained for the operating frequency and bandwidth at the expense of an increased total structure thickness.

Figure 1.10: Cascading two bandpass arrays in vacuum. (a) Unit cell 3-D view.

(b) Equivalent circuit model for the structure. (c) Simulated reflection and transmission coefficients using HFSS and ADS with $b = 12$ mm, $r = 5$ mm, $w = 0.1$ mm, $d = 5$ mm, $\epsilon_r = 1$ and the corresponding circuit model parameters $L = 1$ nH, $C = 0.26$ pF, $Z_0 = 377 \Omega$ and $\beta \cdot d = 90^\circ$ @ 15 GHz.

Moreover, better performance can be achieved regarding the reflection coefficient level between the two resonant frequencies, as seen from Fig. 1.12.

Figure 1.11: Simulated reflection and transmission coefficients for the structure presented in Fig. 1.10 under oblique incidence for TE and TM polarizations.

(a)

(b)

(c)

Figure 1.12: Utilizing dielectric covers. (a) Unit cell of cascaded arrays with cover layers. Simulated reflection and transmission coefficients under oblique incidence for (b) TE polarization and (c) TM polarization. ($b = 12$ mm, $r = 5$ mm, $w = 0.1$ mm, $\epsilon_r = 1.9$, $d = 6$ mm, $\epsilon_{rcover} = 1.3$, $d_{cover} = 10$ mm).

Figure 1.13: Structure of bandstop frequency-selective surface. (a) The 3-D array view. (b) Unit cell excitation and boundary conditions used in simulation. (c) Top view of a unit cell with element dimensions ($b = 16$ mm, $L_d = 14$ mm, $w = 0.4$ mm, $\epsilon_r = 1$).

1.1.2 2-D Bandstop Frequency-Selective Surface

Fig. 1.13 shows 2-D periodic arrays of the cross dipole as an example for bandstop FSS. It can be modelled as a series resonance circuit, in which the inductor represents the current in the dipole and the capacitor represents the electric field in the gap between two dipoles in two adjacent unit cells. Fig. 1.14 shows a comparison between the simulated reflection and transmission coefficients using HFSS with boundary conditions and excitations, as shown in Fig. 1.13 and those obtained from the circuit model using ADS.

As stated above, the loop resonator is the best candidate as long as the stability under oblique incidence is concerned. Fig. 1.15 shows the bandstop response

Figure 1.14: Modeling of the bandstop frequency-selective surface. (a) Equivalent circuit model for the structure presented in Fig. 1.13. (b) Simulated electric field and surface current distribution at resonant frequency for TE polarization and (c) simulated reflection and transmission coefficients using HFSS and ADS with $L = 16.6$ nH, $C = 0.015$ pF and $Z_0 = 377 \Omega$.

Figure 1.15: Frequency-selective surface employing conductor loop element. (a) Unit cell 2-D view, simulated reflection and transmission coefficients under oblique incidence for (b) TE polarization and (c) TM polarization ($b = 12$ mm, $r = 5$ mm, $w = 0.1$ mm, $\epsilon_r = 1$).

employing loop unit cell with the response under oblique incidence for both polarizations.

Like all the traditional 2-D FSS, the selectivity is not sharp and the known method to increase it is to increase the filter order by using the cascading technique, which does not exhibit a good performance in the bandstop case, as shown below. Because in the case of the cascaded bandstop array, all the elements in the front and back arrays contribute to the total reflection. While in the bandpass case, the transmission occurs only from the last array [1, pages 242 and 312]. Fig. 1.16 shows the simulated transmission and reflection coefficients for cascading two loops with a half-wavelength spacing of vacuum. The selectivity is sharper than the bandpass case due to the existence of two zeros outside the stopband, which are called array interface nulls [1, Ch. 8], but the out-of-band performance is not good either in upper or lower band. By including a dielectric layer in between the two arrays, the stability under oblique incidence is improved, but the out-of-band performance is degraded due to the reflection from the dielectric layer (e.g., see [1, Fig. 8.6]). The lower band performance could be improved by adding a cover dielectric layer with a thickness of a half-wavelength that acts as a matching section, which makes the

Figure 1.16: Cascading two bandstop arrays in vacuum. (a) Unit cell 3-D view.

(b) Equivalent circuit model for the structure. (c) Simulated reflection and transmission coefficients using HFSS and ADS with $b = 12$ mm, $r = 5$ mm, $w = 0.1$ mm, $d = 5$ mm, $\epsilon_r = 1$ and the corresponding circuit model parameters $L = 6.1$ nH, $C_1 = 0.043$ pF, $C_2 = 0.0375$ pF, $Z_0 = 377 \Omega$ and $\beta \cdot d = 90^\circ$ @ 14.5 GHz.

Figure 1.17: Salisbury screen absorber. (a) Unit cell 3-D view. (b) Equivalent circuit model for the structure. (c) Simulated reflection coefficient using HFSS for different resistance value with $d = 20$ mm, $\epsilon_r = 1$.

structure very thick. All these techniques used in both cases of bandpass and bandstop are difficult to be implemented because it may end with a design that has dielectric constant and thickness not commercially available.

1.1.3 Microwave Absorber

Microwave absorbers have been used in many civil and military applications. Two primary designs were presented in the literature. The first one is the Dallenbach screen [17], which involves lossy substrate of a quarter wavelength backed with a conducting ground plane. The second one is termed as, Salisbury screen [18], which uses a resistive sheet on a lossless substrate with a quarter-wavelength thickness backed by a conducting surface, as shown in Fig. 1.17(a).

Table 1.1 Comparison between Salisbury and Dallenbach screens.

	Salisbury screen	Dallenbach screen
Construction	<ol style="list-style-type: none"> 1. Perfect electric conductor. 2. Lossless spacer/substrate. 3. Lossy top layer. 	<ol style="list-style-type: none"> 1. Perfect electric conductor. 2. Lossy layer/substrate. <ul style="list-style-type: none"> • Based on the type of the lossy layer, it is classified into two categories; electric and magnetic types.
Bandwidth	<ul style="list-style-type: none"> • Very narrow band. • Bandwidth is enlarged by cascading technique (Jaumann absorber) 	<ul style="list-style-type: none"> • The electric type is narrow band. • The magnetic type is wideband with a thinner structure.
Weight	Lightweight.	The electric type is light, while the magnetic one is heavy.

The equivalent circuit model is shown in Fig. 1.17(b). The input impedance at the surface of the absorber is a parallel combination between the resistance of the lossy layer and the reactance resulting from the conductor-backed lossless substrate. The absorption occurs when the thickness of the lossless substrate equals to a quarter wavelength of the incident electromagnetic wave. At resonance, the input impedance of the conductor-backed substrate equals to infinity and the total input impedance is equal to the pure resistance, which absorbs the incident power when it matches the wave impedance in free space. The simulated reflection coefficient for different resistor values is shown in Fig. 17(c). It is noted that the bandwidth is narrow for the classical Salisbury screen. Multi-layered Salisbury screen (Jaumann absorber) can be used to improve the bandwidth, but it ends with a thick structure [19]. A comparison between Salisbury and Dallenbach screens is summarized in Table 1.1. The comparison is focused on their construction, performance, and weight.

1.1.4 Absorptive Frequency-Selective Structure

FSS exhibits in-band transmission/reflection and out-of-band reflection/transmission. For certain applications aiming to reduce radar signature of antenna systems, it is desirable to obtain out-of-band absorption for both cases. These new structures termed as absorptive frequency-selective transmission structure (AFST) and absorptive frequency-selective reflection structure (AFSR) because they exhibit absorption and transmission or absorption and reflection responses, respectively. More detailed explanation will be provided in Chapters 2 and 7.

By cascading a lossy layer and lossless layer, AFST response is achieved, as shown in Fig 1.18. The lossless layer acts as a bandpass FSS. More details about the losses in the passband and the techniques used to bypass them will be discussed in Chapter 2.

1.2 Motivation

There are several challenges facing the design of high-order bandstop 2-D FSSs, as demonstrated above. Moreover, accomplishing multiband operation is more challenging. The 3-D FSS can outperform the traditional 2-D one. However, the 3-D FSS has some drawbacks as operation under single-polarized electromagnetic waves and the structure is thick. The thickness problem can be solved by utilizing new resonators in the unit cell. By utilizing double-sided parallel strip-line instead of the microstrip line in the unit cell, dual-polarized performance can be achieved.

Most of AFST structure presented in the literature have one-sided absorption band, in which the absorption band is located in either the upper band or lower band of the transmission window. Moreover, those have two-sided absorption, their absorption bandwidth is narrow. The 3-D FSS concept may be useful if it is applied to the design of AFST structures with two-sided and wide absorption bands.

Figure 1.18: AFST structure. (a) Unit cell 3-D view. (b) Equivalent circuit model for the structure. (c) Simulated reflection and transmission coefficients using HFSS for the normal incidence with $d = 20$ mm, $\epsilon_r = 1$.

1.2.1 Limitations of Frequency-Selective Surfaces/Structures

From the demonstration shown in Section 1.1.2, it is clear that the difficulty of achieving high-order bandstop response with sharp skirt using 2-D FSS even with cascading technique. Recently, 3-D FSS that can be considered as the thick form of the traditional 2-D FSS was proposed [20], [21]. This new structure

employs multi-mode resonator as a unit cell to produce a quasi-elliptic response, which may be the best among other responses when the selectivity is concerned. Moreover, because the transverse dimensions do not affect the resonant frequency, a stable response can be achieved even under a large angle of incidence. Despite its good performance, the new structure suffers from two main drawbacks: (i) the operation under only a single polarization and (ii) the structure is thick, which may limit its practical applications. Because it is still relatively new, more research work is required to address these problems and to carry out further research especially for the bandstop response, which is difficult to be implemented with a good performance by employing the traditional 2-D FSS as shown above.

1.2.2 Limitations of Absorber Design

The challenge in the absorber design is to achieve a wide absorption band concurrently with a small thickness. Although the existence of good 2-D absorber designs in the literature, achieving thin absorber design with ultra-wideband performance is still a challenge. One absorber design was presented in the literature, which is based on the 3-D FSS concept [22]. However, the design suffers from the drawback that it operates under a single-polarized wave and the absorption bandwidth is not very wide.

1.2.3 Limitations of Absorptive Frequency-Selective Transmission Structure

It is noted that most of the AFSR designs presented in the literature have one-sided absorption band close to the transmission window with a high reflection level at its boundary for both cases (lower/upper absorption band). Based on 3-D FSS, two AFST structures were presented in [23] and [24]. In [23], only upper absorption band was achieved and the insertion loss was high. A new concept of 3-D AFST was presented in [24] to achieve an absorption at both sides of the transmission window. Moreover, low insertion loss was achieved by bypassing the losses within the transmission window as it was connected in

shunt with a short-circuited half-wavelength transmission line. However, its absorption rate at both sides is 80% because it relies on lossy resonators and narrow absorption bands can then be achieved.

1.3 Objectives

It has been shown in the previous discussion that the existing high-order bandstop 2-D FSSs, 3-D FSSs, traditional 2-D absorbers, and AFST structures suffer from a number of limitations, which may limit their practical applications. In this thesis, we aim to present a comprehensive study of the state-of-the-art of these designs, propose possible solutions and design methodologies to solve these mentioned limitations.

The objectives of the thesis are summarized below.

1. Conduct a comprehensive literature review of 2-D frequency-selective surfaces, 3-D frequency-selective structures, absorbers, and absorptive frequency-selective transmission structures. The primary target of this step is to understand the theory of operation for each type and study techniques to overcome the limitations of previous work.
2. Propose a design methodology to achieve multiband high-order bandstop 3-D frequency-selective structure. The multiband response is achieved by employing different resonators in a unit cell, while the high-order response is accomplished by utilizing the coupling between identical resonators.
3. Design a thin 3-D frequency-selective structure based on a new resonator aiming to solve the thickness problem associated with this structure for both bandstop and bandpass responses.
4. Propose a new unit cell to solve the dual-polarization problem associated with the 3-D frequency-selective structure. By carefully studying and understanding its properties, the dual-polarized problem

may be solved and new features associated with this new unit cell are then discovered.

5. Introduce a methodology to accomplish arbitrary frequency response within an ultra-wide (UW) absorption band. With this, it is possible to achieve absorptive frequency-selective transmission structure with two-sided and wide absorption bands.

1.4 Major Contributions of the Thesis

This thesis introduces several frequency-selective structure applications with novel designs and good performance based on solving the limitations of the existing designs. These proposed structures can achieve multiband high-order bandstop response with a good performance, very thin 3-D structure with bandstop and bandpass performance, ultra-wideband dual-polarized absorber, and absorptive frequency-selective transmission structure with two-sided absorption bands. All of these designs have a stable performance under oblique incidence. Major contributions of the thesis are specified as follows.

1. A general methodology for the design of multiband high-order bandstop 3-D frequency-selective structures is proposed. The methodology is based on employing multiple resonators inside a unit cell to obtain a multiband response and by controlling the coupling between identical resonators a high-order filtering performance is achieved. This provides more transmission zeros in the operating bands, which leads to an improved bandwidth and selectivity. Detailed analysis of the coupling between identical resonators is provided as well as equivalent circuit models to explain the operating principle. Two structures are designed, fabricated, and measured to verify the proposed methodology.
2. Circular and elliptical loop resonators are proposed as the unit cell of bandstop 3-D frequency-selective structure aiming to reduce the structure thickness. The structure exhibits a pseudo-elliptic response by utilizing the coupling between two quasi-TEM modes, which are

concentrated in air and substrate regions of the unit cell, respectively. Using the elliptical loop with stepped-impedance resonator, the thickness of the structure can be reduced by 80% compared to previous structures. A comparison between the response of the structure employing the microstrip line as the resonator and the structure using the proposed loop resonators is performed. Three prototypes are designed, fabricated, and tested.

3. Folded substrate method is proposed to reduce the thickness of 3-D bandpass frequency-selective structure. By employing three- and five-layer folding of the substrate, a 67% and 79% thickness reduction compared to the basic bandpass 3-D FSS is accomplished. One structure with the five-layer folding substrate is designed, fabricated and measured. It exhibits a center frequency of 3.57 GHz with 26.9% transmission bandwidth. The structure thickness equals to $0.06 \lambda_0$, where λ_0 is the free-space wavelength at the center frequency of the passband. Moreover, a half-cylindrical shaped radome as an example of a conformal radome is constructed based on the thin 3-D frequency-selective structure and integrated with a broadband horn antenna. Besides working as a radome it shows good filtering properties. The radiation properties of the whole system are finally examined.
4. Double-sided parallel-strip line resonator is employed in the unit cell of a 3-D frequency-selective structure in order to exhibit the desired response (bandstop, bandpass and absorption) under an incident wave of arbitrary polarization. It is found that this unit cell supports the propagation of three quasi-TEM modes. The first and third modes are concentrated in the substrate and air regions of the unit cell, respectively. While the second mode is concentrated in the interface between the air and substrate of the unit cell. The polarization of the second mode is perpendicular to the polarization of the first and third modes. Detailed analysis of these propagating modes is provided with equivalent circuit models to explain the operating principle. A dual-polarized bandstop 3-

D frequency-selective structure is then proposed. It exhibits a stable frequency response with a constant out-of-band bandwidth under a large range of oblique incident angle. A dual-polarized absorber is then realized employing a unit cell with double-sided parallel-strip line resonator. By properly exciting and absorbing the second mode of the modified unit cell, an ultra-wideband absorber is constructed, which has -10 dB fractional bandwidth of 148% from 1.51 GHz to 10.1 GHz with an absorber thickness of $0.1 \lambda_0$ at the lowest absorption frequency.

5. Tailoring methodology is proposed to achieve arbitrary frequency response within a wide absorption band. An ultra-wideband absorber utilizing multiple resonances in the unit cell of 3-D frequency-selective structure is utilized as the textile for the proposed designs. Two types are proposed; one is the absorptive frequency-selective reflection structure and the other is the absorptive frequency-selective transmission structure. These structures have reflection/transmission band within a wide absorption band. Two prototypes are designed, fabricated, and measured to validate the proposed methodology. The first one is an ultra-wideband absorber with 155.7% absorption bandwidth from 1.8 GHz to 14.445 GHz with a thickness of $0.12 \lambda_0$ at the lowest absorption frequency. The second one is an absorptive frequency-selective transmission structure with wide absorption bands at both sides of its transmission window. It exhibits 54% and 76% for the lower and upper absorption bands, respectively, with 1 dB insertion loss in the transmission window.

1.5 Organization of the Thesis

The thesis is organized into seven chapters. The first chapter primarily introduces the background and motivation behind this work. Based on the limitations, the objectives of the thesis are established. In addition, the major contributions of the thesis are summarized.

Chapter 2 is a literature review of the research work conducted on the 2-D

frequency-selective surface, 3-D frequency-selective structure, absorbers, and absorptive frequency-selective transmission structures. Some techniques are reviewed to improve the stability of performance under various incident angles, improve the selectivity, and achieve the multiband response. This chapter also presents a review of the traditional 2-D absorber designs and existing techniques to achieve the absorptive frequency-selective transmission response.

Chapter 3 proposes a general methodology to design a bandstop 3-D frequency-selective structure that exhibits a multiband high-order response. By employing multiple resonators inside a unit cell, a multiband response is accomplished and high-order filtering performance is achieved by controlling the coupling between identical resonators inside the unit cell. This provides more transmission zeros in the operating bands, which leads to improved bandwidth and high selectivity. Equivalent circuit models to explain the operating principle and detailed analysis of the coupling between identical resonators are provided. Two structures are designed, fabricated, and measured to verify the proposed methodology.

Chapter 4 presents circular and elliptical loop resonators to be employed inside the unit cell of 3-D frequency-selective structure aiming to reduce the structure thickness. Utilizing elliptical loop with stepped-impedance resonator, 80% reduction in the thickness of the structure is achieved compared to previous structures with microstrip line resonator. Three prototypes are designed, fabricated, and tested. Two of them are single-polarized designs and the last one is dual-polarized design.

Chapter 5 proposes a design based on a folded substrate to reduce the thickness of the 3-D bandpass frequency-selective structure. A 67% and 79% thickness reduction compared to the basic bandpass 3-D FSS is achieved by employing three- and five-layer folding of the substrate, respectively. One structure with the five-layer folded substrate is designed, fabricated and measured using a parallel-plate waveguide measurement setup. It has a center frequency of 3.57 GHz with 26.9% -3 dB transmission bandwidth. Moreover, a half-cylindrical shaped radome as an example of a conformal 3-D frequency-selective structure

is constructed based on the proposed technique and integrated with a broadband horn antenna. Besides working as a radome it shows good filtering properties. The radiation properties of the whole antenna-radome system are characterized.

Chapter 6 introduces the double-sided parallel-strip line resonator in the unit cell of a 3-D frequency-selective structure to achieve the desired response (bandstop, bandpass and absorption) under an incident wave of arbitrary polarization. Detailed investigation and analysis of the propagating modes supported by the unit cell are conducted with equivalent circuit models. A dual-polarized bandstop 3-D frequency-selective structure is then presented. Moreover, a dual-polarized multiband high-order bandstop structure is designed, fabricated and tested. By properly exciting and absorbing the second mode of the modified unit cell, an ultra-wideband absorber is constructed, which has -10 dB fractional bandwidth of 148% from 1.51 GHz to 10.1 GHz with an absorber thickness of $0.1 \lambda_0$ at the lowest absorption frequency.

In Chapter 7, arbitrary frequency response within a wide absorption band is achieved. By employing multiple lossy resonators in the unit cell of 3-D frequency-selective structure, an ultra-wideband absorber is designed and fabricated. This ultra-wideband absorber is then utilized as the textile for the presented designs. Two types are proposed; one is the absorptive frequency-selective reflection structure and the other is the absorptive frequency-selective transmission structure. These structures have reflection/transmission band within a wide absorption band.

Chapter 8 concludes the thesis by summarizing the achievements of this research. Suggestions for future work are also provided in this chapter.

CHAPTER 2

LITERATURE REVIEW

2.1 Introduction

Frequency-selective surfaces (FSSs) have been employed in hybrid radomes, dichroic main reflector, sub-reflector, microwave absorbers, absorptive frequency-selective surface, high impedance surfaces, and many other applications. A comprehensive literature review on the existing FSSs and their applications is conducted in this chapter. The review is divided into several parts: stability and selectivity improvement of the 2-D FSS, existing techniques to achieve a multiband response, microwave absorber designs, techniques used to design absorptive frequency-selective transmission structure (AFST) and 3-D FSS concept and its limitations.

As demonstrated in Chapter 1, the traditional FSSs employing resonant element [1] exhibit single-order Butterworth response, which has monotonic increase/decrease in out-of-band performance. Most of these designs have unstable performance under oblique incidence because the period is comparable to half-wavelength at the resonant frequency. This chapter begins with a review of several techniques presented in the literature to enhance performance stability under oblique incidence, improve the selectivity, and achieve a multiband response.

In addition, existing microwave absorber designs are reviewed. The challenge in the absorber design is to achieve a wide absorption band concurrently with a small thickness. Moreover, AFST designs and techniques used to enhance the insertion loss inside the passband presented in the literature are briefly reviewed. The basic concept of the 3-D FSS is revisited with several designs and examples. Finally, some limitations of 3-D FSS are listed.

2.2 Stability Improvement

One needs to reduce element size in the unit cell compared to the operating wavelength in order to decrease the period to achieve better stability under oblique incidence [1]. Several techniques were used in the past as listed below.

2.2.1 Dielectric Slabs at Both Sides of the Screen

The traditional solution to solve the stability problem is to add a dielectric layer to the surface of the array at both sides [1, pages 241, 248 and 254]. However, this technique works well for slot array, but it does not work well for dipole array (bandstop response) [1, pages 243, 244 and 308] because it adds another source of reflection in the dipole case. Moreover, the bandpass (slot-type) design requires dielectric constant values for the slabs that may be not commercially available as illustrated in Chapter 1.

2.2.2 Non-Resonant Element

One of the attractive solutions is to use non-resonant elements (spatial lumped elements) [25] instead of the resonant one. Unfortunately, the response is still sensitive to the oblique incidence (e.g., see [25, Fig. 6, 10]). A stable first-order bandpass response was presented in [26] using single-layer non-resonant elements.

Basically, the non-resonant elements (C/L) are resonant elements (rectangular patch/slot) operating at a frequency lower than their resonant frequency. To achieve the resonant condition, two non-resonant elements have to be employed. It is difficult to achieve bandstop response using non-resonant elements. A bandpass response is only achieved using such elements because it is easy to construct a parallel resonant circuit, as shown in Fig. 2.1. There is no bandstop FSS based on non-resonant elements reported in the literature until now.

Figure 2.1: FSS based on non-resonant element presented in [25]. (a) Adjacent edges of the patch form a capacitance and inductance associated with a metallic strip. (b) 3-D view and side view of one-pole FSS. (c) Equivalent circuit model of one-pole FSS.

2.2.3 Lumped-Reactive Elements

Using lumped reactive element to create the resonant condition leads to miniaturize unit cell size [27]. A stable bandpass response is obtained under various incident angles. The resonant frequency is mainly controlled by the lumped element values. It has some drawbacks because of the used lumped components such as, adding more cost to the fabricated structure, drift in the operating frequency as the component value changes with time and the

Figure 2.2: Dual-polarized bandpass FSS based on lumped component presented in [28]. (a) 3-D view and front view of the unit cell. (b) simulated reflection and transmission coefficients under the normal and oblique incidence for TE and TM polarizations.

proposed design operates only under TE polarization. This design was extended to operate for dual-polarization in [28] and is also shown in Fig. 2.2. Moreover, a bandstop FSS was also presented in [28].

2.2.4 Distributed Capacitance and Inductance

Meandering the resonant element [29] to increase the inductance or capacitance is another effective way to miniaturize the element size. Moreover, increasing the edge length of the element or utilize the face coupling between patches to increase the capacitance with the neighbor unit cell [30, 31], employing fractal configuration [32, 33] are other alternative methods. The via-based frequency-selective surface was proposed [34], based on vertical via holes with the resonant elements to increase the equivalent capacitance and inductance of the element. The diameter of via hole controls its inductance, while the spacing between via holes in two adjacent unit cells controls the mutual capacitance. This can achieve good stability at the expense of the selectivity performance.

2.3 Selectivity Enhancement

To improve the selectivity of the traditional 2-D FSSs, one of the following techniques may be used.

2.3.1 Higher-Order Filter

The traditional technique to increase the filter order is the cascading technique [1] in which two or more first-order screens are cascaded with spacing close to a quarter-wavelength. However, this ends with thick and bulk structure, especially when a dielectric cover is utilized for stabilization, as demonstrated in Chapter 1. Moreover, this technique is useful only for bandpass response, as illustrated in Chapter 1. This spacing can be reduced by employing aperture-coupling interlayer [35]. The aperture-coupling interlayer controls the coupling between the cascaded resonant layers. Low-profile high-order bandpass FSS utilizing non-resonant element was presented in [36, 37] or using miniaturized resonant element in conjunction with non-resonant element [38, 39], which can be seen as antenna-filter-antenna configuration. Butterworth and Chebyshev responses were realized under dual-polarization [36, 37] or single-polarization [38, 39]. Better stability compared to [1 (without substrate between the arrays)] was achieved without using dielectric slabs.

A double-layer thin substrate high-order bandpass FSS was presented in [40] by utilizing the coupling between the elements on two sides of a thin substrate and using lumped capacitors to enhance the capacitive coupling between two neighboring elements. However, the resonant frequency changes with the incident angle.

One-sided sharp roll-off was reported in [41] using meta-surface. However, the performance is limited to operation under single polarization for two proposed unit cells.

2.3.2 Antenna-Filter-Antenna

High-order Chebyshev response using antenna-filter-antenna was proposed in [42]. The antenna-filter-antenna unit cell is formed by combining two antennas and non-radiating resonator in between. A higher-order bandpass response can be achieved by adding more resonators to the intermediate layer instead of adding more layers as in cascading technique. However, its response was polarization sensitive and its performance changes under the oblique incidence.

2.3.3 Quasi-Elliptic Response

Using the previous techniques, one can achieve Butterworth or Chebyshev responses, but not elliptic response, which has the sharpest roll-off among all the filtering responses. It is not easy to achieve elliptic response using 2-D surfaces due to the lack of cross-coupling between multiple layers [43].

Several designs have been developed to achieve quasi-elliptic bandpass FSS response. One was based on substrate-integrated waveguide [43] and others were based on dual-behavior resonator [44, 45].

2.4 Multiband Response

In the past, several techniques were proposed to design multiband FSS. Most of these techniques are suitable for bandpass response, such as those summarized in the following.

2.4.1 Perturbation

The perturbation technique is often used when the multiband response is needed. For example, using multiple resonant elements [46-48], perturbing both element and spacing of a single band array [49], loading the unit cell with different lumped element values [50], and by perturbing unit cell size [51]. A tri-band second-order bandpass response was reported in [52], which utilizes the unit cell size perturbation with aperture-coupled patch configuration [53]. Unfortunately, because its unit cell size is comparable to the operating wavelength, two problems occur at a large incident angle: (1) the band spacing is limited as grating lobes will significantly affect the performance; (2) the insertion loss will be significant for large incident angles ($\Theta > 30^\circ$). A first-order multiband bandpass design based on LC resonator was presented in [112]. The parallel LC resonator was constructed from a via-connected metallic line and a parallel-plate capacitor, which exhibits a single-band bandpass filtering response. By employing different resonators with different resonant frequencies, the multiband response was achieved. A systematic technique to design multiband multilayer bandstop FSS with low-profile was introduced in [113]. Periodically loaded dipoles are used on each layer to create a single stopband response. Several layers are cascaded together to create the desired first-order multi stopband response.

2.4.2 Fractal Element

Utilizing the multiband feature of the self-similar fractal elements [54], dual-band response was achieved. The geometry of the Sierpinski dipole offers sufficient degrees of freedom to modify its shape in order to tune its response.

2.4.3 Complementary Frequency-Selective Surface

It consists of two closely coupled FSSs: one is formed using conducting elements and the other is formed by aperture elements, placed at both sides of the substrate. Using this surface [55] one is able to achieve dual-band response. A tri-band response was obtained by combining these surfaces with non-

resonant elements [56]. The complementary pattern was used with a minimized element to achieve dual-band response [57].

2.4.4 Non-Resonant Element

By employing non-resonant elements, reference [58] achieved dual-band first-order bandpass response using a single substrate. Most of the above designs exhibit a multiband first-order response. Based on non-resonant elements, several designs have been developed to achieve multiband high-order frequency-selective surfaces. One was based on synthesizing the FSS from dual-band second-order bandpass microwave filter topology [59] and another was based on a fourth-order coupled resonator filter, where the four poles of the transfer function were split into two groups [60]. These two designs have dual-band second-order bandpass response.

2.5 Absorber Designs

As introduced in Chapter 1, Dallenbach [17] and Salisbury [18] screens are considered as the earliest absorber designs. The optimum design of Salisbury screen was reported in [61]. By capacitively loading the screen [62], its thickness can be reduced. By using capacitive loading with Jaumann absorber, the total thickness can be reduced concurrently with an enhanced absorption bandwidth [63], [64]. Circuit analog absorber was presented in [1] by exploiting the feature of the series resonant circuit that acts as a capacitor before the resonant frequency and as an inductor after the resonant frequency to match the performance of a conductor-backed substrate. The challenge in the absorber design is to achieve a wide absorption band concurrently with a small thickness.

A single-layer wideband absorber was proposed in [65]–[67] by employing multiple resonances. Several designs were presented in the literature based on this concept. An ultra-thin and broadband absorber was presented in [68] by employing resistive treble-square loops. Two and three resonances were produced to widen the absorption bandwidth in [65] and [69], respectively, by employing double-square-loop array with resistors. By employing a second

lossy layer above the one presented in [69], a nearly doubled bandwidth was achieved in [114] with a little increase in the structure's thickness. Utilizing three lossy layers, a broadband polarization-insensitive absorber was accomplished in [115]. Each layer consists of a periodic structure of resistive patterns printed on a low-cost FR4 sheet using resistive ink. An ultra-thin absorber was reported in [70] by loading a resistive sheet directly to a high-impedance surface, but the bandwidth was narrow. An ultrathin and broadband absorber was introduced in [116] utilizing a lossy FSS layer and frequency-dispersive magnetic material. The frequency-dispersive magnetic material has both high permeability and permittivity for thickness reduction and near frequency-nondispersive input susceptance for bandwidth enhancement at low frequencies. A non-Foster active absorber with an ultrawideband and extremely low profile, which can effectively break the Rozanov limit for passive absorbers was presented in [117]. A negative inductor is used for actively matching the equivalent inductance of a thin conductor-backed spacer. All of the previous structures used a resistive sheet or lumped resistors to produce the absorption. By using a structure consisting of two metamaterial resonators that couple separately to electric and magnetic fields [71], the absorption takes place due to the ohmic and dielectric losses, which is called the perfect metamaterial absorber.

Based on 3-D FSS, a wideband microwave absorber was presented in [22]. This structure works as an absorber when the polarization of the incident electric field is perpendicular to the microstrip line in the unit cell in order to excite two quasi-TEM modes. One of these two modes is mainly concentrated in the air region, while the other is primarily concentrated in the substrate region. This absorber is a bandstop 3-D FSS [20] with a backed conducting plate and lumped elements to match it to free space and absorb the incident power. A broadband impedance matching to the air transmission line was achieved using a series R-C network and a resistor in the third dimension (the z-axis) to absorb the power in the air and substrate transmission lines. The equivalent circuit model is constructed and its simulated reflection coefficient is compared with the simulated reflection coefficient presented in [22]. Fig. 2.3(a) shows the

Figure 2.3: (a) Equivalent circuit model for the absorber reported in [22]. (b) A comparison between CST and ADS results ($Z_p = 434 \, \Omega$, $C_{\text{sub}} = 0.027 \, \text{pF}$, $Z_a = 340 \, \Omega$, $\beta_a l = 180^\circ @ 7.8 \, \text{GHz}$, $Z_s = 40 \, \Omega$, $\beta_s l = 180^\circ @ 5.6 \, \text{GHz}$, $L_{\text{strip}} = 3.3 \, \text{nH}$, $C_a = 0.5 \, \text{pF}$, $R_a = 180 \, \Omega$, $R_s = 120 \, \Omega$).

equivalent circuit model of the absorber, which consists of two transmission lines representing the air and substrate quasi-TEM modes supported by the unit cell under the TE-polarized wave. Resistor (R_s) at the end of the microstrip line is shared between air and substrate transmission lines. This resistor contributes to the absorption of energy in both transmission lines. A comparison between CST and ADS simulated results is shown in Fig. 2.3(b).

2.6 Absorptive Frequency-Selective Transmission Structures

AFST structure exhibits in-band transmission and out-of-band absorption as illustrated in Chapter 1. This AFST structure was named as “frequency-selective rasorber” in [23], [24] and [72]–[74], “absorptive FSS” in [75]–[79] and “frequency-selective radome” in [80] and [81]. Throughout the thesis, the term AFST is used because it selects a specific frequency band to be transmitted within a wide absorption band. A complementary response to AFST response was presented in [82], [83], which exhibits transmission-absorption-transmission response.

2.6.1 One-Sided Absorption with Lower Transmission Window

Most of AFST designs in the literature were realized by cascading a lossless 2-D FSS layer [1] with lossy layers. The initial idea was introduced in [84], though it did not provide any designs or results. A lossless layer of interdigitated Jerusalem cross element, which has a large reflection band, was utilized as a ground plane for a thin wideband absorber [80] and shown in Fig. 2.4. Low insertion loss was achieved by placing the passband away from the absorption band. Two lossy layers were employed to enhance the upper out-of-band absorption in [73]. Based on the concept of multiple resonances proposed in [65] and [67] to enhance the absorption bandwidth, a double-square loop layer was cascaded by a convoluted cross-slot layer to achieve AFST with a wide absorption band [75]. Two bandpass FSS layers were cascaded to improve the passband response [81]. A minimized AFST design was presented in [76] and due to its small periodicity, grating lobes in the absorption band were eliminated when oblique incidence was considered.

(a)

(b)

Figure 2.4: (a) 3-D view of AFST with antenna in [80]. (b) Reflection and transmission coefficients under the normal incidence in receiving mode.

Figure 2.5: Perspective view of the structure and unit cell of AFST in [77].

2.6.2 One-Sided Absorption with Upper Transmission Window

The frequency response of these previous designs only has one absorption band with lower transmission window, which leads to a strong reflection at the lower frequency band. Other designs in the literature have the absorption band below the transmission window [77]–[79]. The design presented in [77] is shown in Fig. 2.5, which is based on the conceptual design reported in [72]. This design employed a quarter-wavelength short-circuited stub in the middle of the lossy resonator to prevent the current from flowing through the resistor, hence bypassing the losses at the passband frequencies. The short circuit at the end of the quarter-wavelength stub is transferred to an open circuit at its beginning, hence cutting the current path. This method was described in [72], where the term choke was used instead of a short-circuited stub. A parallel resonator in series with the resistor was utilized in [78] to cut the current path, as shown in Fig. 2.6. A dual-polarized AFSR design with upper and wide transmission band was recently presented in [118]. This design utilizes a second-order bandpass

Figure 2.6: Perspective view of the structure and unit cell of AFST in [78].

FSS [53] to achieve a wide transmission band. It is noted that all of these previous designs have one-sided absorption band close to the transmission window with a high reflection level at its boundary for both cases (lower/upper absorption band).

2.6.3 Two-Sided Absorption

A complete conceptual design based on cascading 2-D FSS layers was reported in [72] to achieve AFST with two-sided absorption bands and is shown in Fig. 2.7. First, reference [72] presented a method to achieve a middle passband and a low absorption band by cascading a lossless bandpass FSS layer, passing the middle frequencies, with a lossy layer with a choke working at the passband frequencies, named as AFST#1. Then two FSS layers are placed in front of AFST#1 to absorb the high-frequency waves. These layers consist of a lossy layer to absorb waves at high frequencies cascaded by a lossless bandstop FSS, which serves as a ground for the front layer. Two designs were reported in [74], and one of them is shown in Fig. 2.8. Both designs have two-sided

Figure 2.7: Complete conceptual AFST structure with two-sided absorption bands presented in [72].

Figure 2.8: Geometry and unit cell details of the preliminary AFST design with two-sided absorption bands reported in [74].

absorption bands around the transmission window. The transmission window was achieved by utilizing a dual-band bandstop FSS in the first design, while employing a bandpass FSS in the second design. However, their absorption bands were narrow.

Several AFST designs with two-sided absorption bands were presented in the literature [119]–[125]. These designs were based on cascaded 2-D FSS layers and using different implementations to cut the current path, hence bypassing the losses in the passband frequencies. These implementations used a parallel LC resonator [121], quarter-wavelength choke [119], slot resonator [120], square-loop hybrid resonator [122], stub resonator [123], and split-ring hybrid resonator [125].

2.7 3-D Frequency-Selective Structures

3-D FSS employing multi-mode resonators as a unit cell to produce quasi-elliptic response was presented in [20, 21].

2.7.1 Operating Principle

Fig. 2.9 shows the structure presented in [20] as well as its performance under oblique incidence. The structure depends on the coupling between two quasi-TEM modes concentrated in the air and substrate regions to produce transmission and reflection zeros. The unit cell produces two reflection zeroes and one transmission zero. The two reflection zeroes are produced when the propagating path of each mode is equal to π . The one transmission zero occurs when the length of the microstrip line is approximately equal to $0.5 \lambda_{\text{eff}}$ at which microstrip line resonates, where $\lambda_{\text{eff}} = \lambda_0 / \sqrt{(\epsilon_r + 1)/2}$ and λ_0 is the free space wavelength at the operating frequency. This performance is only achieved when the electric field polarization is along the y -direction at which the electric field is normal to the microstrip.

The unit cell consists of two parallel TLs (air TL and substrate TL) and it can be modeled by the equivalent circuit shown in Fig. 2.10(a). With a unit cell's

Figure 2.9: 3-D FSS presented in [20]. (a) 3-D structure view. (b) Detailed unit cell excitation and dimensions. (c) Simulated reflection and transmission coefficients under various angles of incidence for TE polarization (E in y -direction) ($b = 5$ mm, $h = 3.524$ mm, $L = 10.5$ mm, $d = 1.524$ mm, $w = 1$ mm, $\epsilon_r = 3$).

Figure 2.10: Modeling of the 3-D FSS. (a) Equivalent circuit model of unit cell in Fig. 2.1. (b) Simulated reflection and transmission coefficients using HFSS and ADS.

physical dimensions of $b = 5$ mm, $h = 3.524$ mm, $d = 1.524$ mm, $w = 1$ mm, $L = 10.5$ mm and using a substrate with $\epsilon_r = 3$, the air TL resonates ($\beta_a l = 180^\circ$) at 14 GHz and has characteristic impedance $Z_a = 147 \Omega$ with $C_a = 0.01$ pF that represents the discontinuity between Port 1/Port 2 and air TL. The substrate TL resonates ($\beta_s l = 180^\circ$) at 10 GHz and has characteristic impedance $Z_s = 66 \Omega$ with $C_s = 0.04$ pF representing the discontinuity between Port 1/Port 2 and substrate TL. C_a and C_s are microstrip open-end capacitances for air and substrate TLs, respectively. Port 1 and Port 2 are two parallel-plate waveguide ports with characteristic impedance $Z_p = 265.7 \Omega$, which can be calculated with the standard formula for the characteristic impedance of parallel-plate waveguide. Fig. 2.10(b) compares the simulated reflection and transmission coefficients for the structure using HFSS and those obtained from the circuit model using ADS.

2.7.2 Previous 3-D FSS Designs

Based on this new 3-D structure, several designs have been reported to show the advantages of the new structure. By employing a waveguide with microstrip line inserted, a single-band second-order bandpass response was achieved in [85]. Reference [86] proposed a structure that has a bandpass response and produces multiple out-of-band transmission zeros by blocking the air path. A dual-polarized response with wide out-of-band rejection and fractional bandwidths of 18%, 78% for bandpass and band-reject, respectively was presented in [87]. One design with wideband harmonic rejection was presented [88] employing parallel plate stepped impedance resonator. A dual-band response was described in [89] by introducing a new longer path along with the parallel plate stepped impedance resonator. A tunable bandpass 3-D FSS with wide tuning range was introduced in [90]. By employing parallel-strip line resonator, a single-polarized wideband bandstop response was presented in [91]. Although it has a good filtering performance, the new 3-D FSS is very thick, which may limit its practical applications. One way to reduce the thickness was introduced in [92] using stepped-impedance resonator (SIR) and it was successful to reduce the thickness by 36%. Most of the reported 3-D FSS structures exhibit bandpass response.

Based on 3-D FSS, an AFST structure was reported in [23] and shown in Fig. 2.11. Only upper absorption band was accomplished by placing a resistor in the air path of [86]. Moreover, the insertion loss was high because the losses due to the resistor were not bypassed. A design with low insertion loss was reported in [24]. A short-circuited half-wavelength transmission line was connected in shunt with the resistor, hence bypassing the losses within the transmission window. Two lossy resonators were employed in [93] to achieve two-sided absorption bands. One of these lossy resonators was responsible for the lower and upper absorption bands, while the other lossy resonator was utilized to extend the upper absorption band. The lower absorption band was narrow, while the upper absorption band was very wide. By employing magnetic material within 3-D FSS, AFST structure with a wide upper band was achieved

Figure 2.11: Structure and detailed unit cell view of AFST structure presented in [23].

in [94]. However, the bandwidth of the upper band is narrower compared to the design reported in [93]. In the end, a comparison between 2-D and 3-D FSSs is shown in Table 2.1. This comparison contains performance characteristics and fabrication complexity for first- and high-order bandpass and bandstop FSSs.

2.8 Summary

Several techniques presented in the literature to overcome the limitations of 2-D frequency-selective surfaces including stability of the performance under oblique incidence and selectivity sharpness have been reviewed in this chapter.

Table 2.1 Comparison between traditional 2-D and 3-D FSSs.

	Polarization sensitivity	Selectivity sharpness	Performance under oblique incidence	Fabrication complexity
Bandpass bandstop first-order 2-D FSS.	Work for both TE and TM polarizations.	Exhibits Butterworth response (Not sharp).	Stable under both polarizations for the loop-type resonator.	- Employing PCB technology. - Simple.
Bandpass high-order 2-D FSS.	Work for both TE and TM polarizations.	Exhibits Butterworth and Chebyshev responses (sharper than the first-order one).	Not stable under TM polarization	- Employing PCB technology. - Simple.
			Stability improved by adding two cover dielectric layers.	Requires dielectric constant values for the cover layers that may be not commercially available.
Bandpass first and high-order 3-D FSS.	Work for both TE and TM polarizations.	Exhibits quasi-elliptic response (sharpest roll-off among all responses).	Stable under both polarizations.	- Employing PCB technology. - Complex.
Bandstop high-order 2-D FSS.	Work for both TE and TM polarizations.	- Sharper roll-off than the first-order one. - Out-of-band performance is not good.	Not stable under both polarizations.	- Employing PCB technology. - Simple.
Bandstop high-order 3-D FSS.	Work for TE polarization only.	Exhibits quasi-elliptic response.	Stable under TE polarization.	- Employing PCB technology. - Complex.

This chapter also summarizes several techniques used in the past to achieve the multiband bandstop response.

Following the review of 2-D frequency-selective surface, microwave absorber and absorptive frequency-selective transmission structure designs have been reviewed. In addition, the new 3-D frequency-selective structure has been presented. Some of the work based on the new structure has been reviewed, while most of this work shows bandpass response. There is no general methodology for achieving multiband high-order response presented in the literature for a bandpass or bandstop responses. This thesis proposes a general methodology to achieve multiband high-order bandstop 3-D frequency-selective structure in Chapter 3. Then Chapter 4 proposes a new resonator inside the unit cell of 3-D bandstop frequency-selective structure aiming to reduce its thickness. A design method based on folded-substrate is employed in Chapter 5 to reduce the thickness of bandpass 3-D frequency-selective structure. This review also highlights a dual-polarization problem with the bandstop 3-D frequency-selective structure. In Chapter 6, a double-sided parallel-strip line resonator is employed inside the unit cell to solve this problem.

It has been shown in the literature review that designing an absorptive frequency-selective transmission structure with two-sided and wide absorption bands is challenging. In Chapter 7, a tailoring methodology is proposed to achieve an arbitrary frequency response in a wide absorption band. This proposed methodology helps to design absorptive frequency-selective transmission structure with two-sided and wide absorption bands.

CHAPTER 3

MULTIBAND HIGH-ORDER BANDSTOP 3-D FREQUENCY-SELECTIVE STRUCTURES

3.1 Introduction

As mentioned in the previous chapters, most of the designs were based on the traditional 2-D frequency-selective surface (FSS), and little has been reported on the 3-D bandstop structure. In this chapter, a general methodology to design a multiband high-order 3-D bandstop frequency-selective structures (FSSs) is proposed. The methodology is based on employing multiple resonators inside a unit cell to obtain a multiband response and by controlling the coupling between identical resonators a high-order filtering performance is achieved. This provides more transmission zeros in the operating bands, which leads to an improved bandwidth and selectivity. Detailed analysis of the coupling between identical resonators is provided as well as equivalent circuit models to explain the operating principle. Two structures are designed, fabricated, and measured to verify the proposed methodology. The first structure is a tri-band second-order FSS with center frequencies at 9.6, 13.4 and 17.5 GHz and their -3 dB fractional bandwidths of 13.5%, 8.5%, and 12.6%, respectively. The second structure is designed to have a dual-band third-order response with center frequencies at 13.2 and 19.6 GHz and their -3 dB fractional bandwidths of 16% and 13.3%, respectively.

3.2 General Methodology

The 3-D FSS presented in [20] depends on the coupling between two quasi-TEM modes, respectively, concentrated in air and substrate regions. Its unit cell produces a single-band first-order response with one transmission zero occurring when the length of the microstrip line approximately equals to $0.5 \lambda_{\text{eff}}$

at which microstrip line resonates. At the same time, two reflection zeros are obtained when the two paths (air-path and substrate-path) have an electrical length equal to π at which the input impedance of the propagating path is matched to the free space impedance for minimum reflection. As these two modes propagate in different media at different velocities, they then have two different propagation constants meaning that they will achieve an electrical length equal to π at two different frequencies.

Based on the coupled-mode theory [95], Fig. 3.1(a) shows a general equivalent block diagram of a multiband high-order 3-D FSS. Constructed from N different groups of M identical resonators. The N different resonators produce N -band response and M identical resonators along the same transmission path produce M^{th} -order filtering response. By controlling the coupling between the identical resonators, one can achieve more in-band zeros. The structure presented in [20] has a single resonator; therefore it has a single-band first-order response with the equivalent block diagram shown in Fig. 3.1(b), which is the simplest case of the general block diagram when $N = 1$ and $M = 1$.

To verify this general methodology, three designs are considered here. The first design has a single-band second-order response with one group of two identical resonators ($N = 1$ and $M = 2$); the second has a tri-band second-order response, which consists of three groups of two identical resonators ($N = 3$ and $M = 2$); the third design has a dual-band third-order response with two groups of three identical resonators ($N = 2$ and $M = 3$). The flexibility of this methodology appears versatile when a multiband response with different orders is desired. For example, to achieve a dual-band with a second-order and third-order response for the first and second bands, respectively, we then need two groups of different resonators: the first group has two identical resonators and the second group has three identical resonators. By adjusting the coupling between the identical resonators in each group, the required response can be finally achieved.

Figure 3.1: Equivalent block diagrams showing the coupling between modes and resonators. (a) Block diagram for the proposed methodology. (b) Block diagram of the presented structure [20]. C_{1a} : coupling between air-mode and free space, C_{2a} : coupling between substrate-mode and free space, C_{12} : coupling between the two modes, and CR_{MN} : coupling between identical resonators.

3.3 Single-Band Second-Order Bandstop FSS

3.3.1 Description of the Structure

Fig. 3.2(a) shows the 3-D view and unit cell details of a single-band second-order bandstop FSS. It is similar to the structure presented in [20] except a gap at the middle of the microstrip line that makes it divided into two equal parts and two via holes at the end of each part that make them resonate at a quarter-wavelength instead of a half-wavelength for the one without via holes.

Fig. 3.2(b) shows the simulated reflection and transmission coefficients of the proposed structure using HFSS. As shown, two transmission zeros are observed. Detailed even-odd mode analysis of the structure will be provided in the following subsection. In HFSS simulations, a Frequency-Domain Solver is used with Master and Slave boundary conditions to represent the periodic boundary conditions of the unit cell and Floquet ports are used as the excitation ports.

3.3.2 Even-Odd Mode Analysis

Consider the unit cell proposed in Fig. 3.2(a) without via holes first. A microstrip line with a gap at the middle can be analyzed using the even-odd mode method [96]. Figs. 3.3(a), (b) and (c) show the overall, odd- and even-mode equivalent circuits, respectively. Odd-mode equivalent circuit consists of two transmission lines (TLs) each connected at its end to ground through the odd capacitor ($C_{odd} = 2C_g + C_p$), each TL resonates at a frequency lower than its fundamental frequency. The even-mode equivalent circuit consists of two TLs each connected at the middle to ground through a capacitor ($C_{even} = C_p$). The even-mode TL resonates at a frequency lower than its first harmonic of the odd-mode resonance. For $L = 10.5$ mm and $\epsilon_r = 3$, $f_{even} < 20.2$ GHz and $f_{odd} \ll 20.2$ GHz. Because C_{odd} is larger than C_{even} , the odd-mode resonant frequency is lower than the even-mode one.

Figure 3.2: Structure of the proposed single-band second-order FSS. (a) 3-D structure view with unit cell details. (b) Simulated reflection and transmission coefficients ($b = 5$ mm, $h = 3.524$ mm, $L = 10.5$ mm, $d = 1.524$ mm, $w = 1$ mm, $g = 0.1$ mm, $r_1 = 0.25$ mm, $r_2 = 0.25$ mm, $\epsilon_r = 3$).

The full-wave simulated reflection and transmission coefficients are shown in Fig. 3.4(a) for the structure illustrated in Fig. 3.2(a) without via holes. It is

Figure 3.3: Equivalent circuits for a microstrip with gap capacitor.

Figure 3.4: HFSS simulations of the structure shown in Fig. 3.2 without via holes. (a) Simulated reflection and transmission coefficients. Simulated vector surface current and electric field for (b) odd mode and (c) even mode.

seen that we have two transmission zeros (f_{odd} and f_{even}). The substrate reflection zero ($R_{zs} \approx 10$ GHz) occurs when the substrate path has an electrical length equal to π . Figs. 3.4(b) and (c) depict the simulated vector currents on the TL and vector electric field inside the substrate at the lower and upper frequencies, respectively. It is seen that the full-wave simulation is consistent with even-odd mode analysis.

If the gap size (g) changes, C_g will be affected more than C_p . Therefore, C_{odd} is affected more than C_{even} , which indicates a more significant change in the lower resonant frequency. If the gap size increases, then C_g decreases, which causes the odd-mode resonant frequency to increase. Fig. 3.5(a) shows the simulated transmission coefficient variation with different gap sizes. As shown, the lower resonance increases as g increases. Fig. 3.5(b) shows the changes of the two resonances with the line width because the line width affects C_p as well as C_g .

The microstrip line with a gap can be considered as two identical half-wavelength resonators resonating at the same frequency. According to coupled-mode theory [95], if two identical resonators are brought near each other, their single resonance is split into two separated resonances. This separation is controlled by their coupling coefficient. In our case, this coupling is due to the spacing between the resonators that is controlled by the gap size (g). Usually, the coupling coefficient (k) is used to represent the coupling strength [97, 98], which can be calculated using the frequency splitting formula [98].

$$k = \frac{f_{++}^2 - f_{+-}^2}{f_{++}^2 + f_{+-}^2} \quad (1)$$

The symmetric and anti-symmetric frequencies (f_{++} and f_{+-}) are calculated using HFSS Eigenmode Solver. The symmetric (anti-symmetric) frequency is the resonant frequency when the symmetry plane is a magnetic (electric) wall. Fig. 3.6 shows the variation of the coupling coefficient (k), symmetric and anti-symmetric frequencies with respect to the spacing between the two resonators (g) shown in Fig. 3.4. From (1), when the coupling is stronger, the separation

(a)

(b)

Figure 3.5: Simulated transmission coefficient of the structure shown in Fig. 3.2 (a) without via hole, with varying: (a) gap size and (b) line width.

between their resonant frequencies becomes bigger. Fig. 3.6 is consistent with Fig. 3.5(a). The larger the separation (g) between the resonators is, the weaker is the coupling between them and the narrower is the separation between these resonant frequencies. The simulation setup used in Fig. 3.2 is utilized to generate the results presented in Fig. 3.5, while the results presented in Fig. 3.6 are extracted using HFSS Eigenmode Solver, in which a single unit cell is simulated without any excitation. The difference in the resonant frequency values between Figs. 3.6 and 3.5(a) is due to the effect of the coupling between

Figure 3.6: Simulated coupling coefficient, symmetric and anti-symmetric frequencies with varying gap size.

adjacent unit cells in Frequency-Domain Solver, which is not presented in the Eigenmode Solver.

As a via hole is included at the end of each half, it makes them resonate approximately at half of the resonant frequencies for the structure without the via hole, as shown in Fig. 3.2(b). Also, this structure can be considered as one with two identical resonators coupled in space, which results in splitting in their original resonant frequency. Fig. 3.7(a) shows the effect of changing the via hole radius (r) when via hole1 radius (r_1) is the same as via hole2 radius (r_2). As expected, via hole radius affects the two resonant frequencies. We can add another degree of freedom to control the separation between the two transmission zeros besides varying the gap size, by making the two via holes radii different. Fig. 3.7(b) shows the effect of changing via hole1 radius while maintaining via hole2 radius constant. It is found that the change of r_1 affects the even-mode resonant frequency and as r_1 decreases, the increase in its inductance leads to a decrease in the even-mode resonant frequency.

Figure 3.7: Simulated transmission coefficient of the structure shown in Fig. 3.2 with varying via hole radius: (a) $r_1 = r_2 = r$ and (b) varying $r_1, r_2 = 0.25$ mm.

3.3.3 Equivalent Circuit Model

In order to construct the equivalent circuit model of the unit cell presented in Fig. 3.2, we consider three unit cells (UC) as follows: (1) UC1 shown in Fig. 2.9, which is similar to the one shown in Fig. 3.2(a) without gap and via holes, (2) UC2 shown in Fig. 3.8(a), which is like the one shown in Fig. 3.2(a) without the via holes and (3) UC3 shown in Fig. 3.9(a), the one shown in Fig. 3.2(a).

For UC2 (UC1 with a gap in the middle of the microstrip line), the substrate TL has gap capacitances C_{gs} and C_{ps} , while the air TL has gap capacitances C_{ga}

Figure 3.8: Modeling of UC2. (a) Unit cell 2-D view and equivalent circuit model. (b) Simulated reflection and transmission coefficients using HFSS and ADS.

and C_{pa} . These capacitances can be approximately calculated using the curves in [95].

The equivalent circuit model of UC1 is shown at the end of Chapter 2. Fig. 3.8(a) shows the equivalent circuit model of UC2. Using the same physical unit cell dimensions of UC1 and a gap size $g = 0.1$ mm. For substrate TL, this gap size produces $C_{even} = 0.00225$ pF/mm, $C_{odd} = 0.09$ pF/mm. Using $C_{odd} = 2C_g + C_p$, $C_{even} = C_p$, we obtain $C_{gs} = 0.044$ pF, $C_{ps} = 0.00225$ pF. While for the air TL, this gap size produces $C_{even} = 0.0015$ pF/mm, $C_{odd} = 0.07$ pF/mm. Using $C_{odd} = 2C_g + C_p$, $C_{even} = C_p$, we obtain $C_{ga} = 0.0343$ pF, $C_{pa} = 0.0015$ pF. These values are not accurate, because [96] uses the quasi-static analysis based on solving Laplace's equation to get the gap capacitances and neglect their frequency dependence [99, 100]. Reference [100] improved the above model by adding

Figure 3.9: Modeling of UC3. (a) Unit cell 2-D view and equivalent circuit model. (b) Simulated reflection and transmission coefficients using HFSS and ADS.

frequency-dependent elements in the computer-aided model. It divides the series capacitance into three parts and uses frequency-dependent ϵ_{eff} , w_{eff} to calculate their values. The capacitance values after using the procedures pointed in [100] are $C_{gs} = 0.078$ pF, $C_{ps} = 0.003$ pF, $C_{ga} = 0.054$ pF and $C_{pa} = 0.002$ pF. Fig. 3.8(b) compares the simulated reflection and transmission coefficients for UC2 using HFSS and those obtained from the circuit model.

Finally, we consider UC3 (UC2 with via holes). Via hole can be represented by an inductor in the circuit model and the inductance (L_v) can be calculated using the formula presented in [86]. Figs. 3.9(a) and (b) show the equivalent circuit model for UC3 and the simulated reflection and transmission coefficients using Agilent ADS and HFSS, respectively. As observed, the two resonant frequencies are lowered to half of the values compared to the one shown in Fig. 3.8(b). The structure includes two types of losses: conductor loss and dielectric

loss. It has been found through simulations (not shown here) by examining the structure with and without the conductor and dielectric losses that the total losses are relatively small at around 10 GHz and their effect on the FSS performance is negligible.

3.4 Tri-Band Second-Order Bandstop FSS

The structure proposed in Section 3.3 can be extended to achieve multiband response by adding other transmission line sections of different lengths. As an example, a tri-band second-order bandstop FSS is designed in this section. It has an equivalent block diagram like the one shown in Fig. 3.1(a) with $N = 3$ and $M = 2$. Therefore, three groups of two identical resonators are needed.

3.4.1 Description of the Structure

Fig. 3.10(a) shows the 3-D view and unit cell details of the proposed structure. As seen, two additional TLs with different lengths are added to obtain a tri-band response. The length of each line (L_1 , L_2 , and L_3) is approximately equal to $\lambda_{\text{eff}} / 2$, where $\lambda_{\text{eff}} = \lambda_0 / \sqrt{\epsilon_{\text{eff}}}$ and λ_0 is the free space wavelength at the center frequency. The separations between these bands are controlled separately and readily adjusted by changing the length of the TLs. Based on the analysis provided in Section 3.3, the separation between the two transmission zeros in each band is controlled by the gap size and via hole radius.

3.4.2 Simulation and Measurement Results

In HFSS simulations, we use periodic boundary conditions (Master and Slave boundary conditions) as the boundaries of the unit cell and Floquet port as the excitation port. Fig. 3.10(b) shows the simulated transmission and reflection coefficients of the proposed structure for TE polarization under normal and oblique incidences together with the simulated surface current distribution at the normal incidence. As shown, each TL of the three TLs is responsible for its own band; therefore each band can be controlled separately. Stable frequency response is achieved due to small unit cell size compared to the operating

Figure 3.10: Structure of the proposed tri-band second-order FSS. (a) 3-D structure view with unit cell details. (b) Simulated transmission and reflection coefficients of the tri-band second-order FSS for different incident angles ($b = 6$ mm, $h = 5$ mm, $d = 1.524$, $w = S = 1$ mm, $L = L_1 = 11$ mm, $g_1 = 1.2$ mm, $L_2 = 6.6$ mm, $g_2 = 1.6$ mm, $L_3 = 4.6$ mm, $g_3 = 1.4$ mm, $r_{11} = 0.4$ mm, $r_{12} = 0.5$ mm, $r_{21} = r_{22} = r_{31} = r_{32} = 0.2$ mm, $\epsilon_r = 3$).

wavelength. The structure presented in [20] exhibited a degraded out-of-band performance at low frequencies, especially for the reflection coefficient when the incident angle exceeds 35° . This high reflection level is due to the weak coupling to the air region, which can be solved by increasing h and maintaining d . As shown in Fig. 3.10(b), the reflection coefficient of our proposed structure is below -10 dB even at 40° . The out-of-band performance in 2-D frequency-selective surfaces is improved by adding a dielectric layer to act as a matching section which leads to a thick structure. It is worth mentioning that the proposed structures in this chapter are polarization sensitive and work as multiband bandstop FSSs under TE-polarized wave, which will be extended to a dual-polarized one in Chapter 6. Under the TM-polarized wave, the proposed structure works as a reflector because the electric field component of the incident wave is parallel to the ground plane of the unit cell.

Fig. 3.11(a) shows the photograph of the fabricated prototype, which is approximately $200 \text{ mm} \times 200 \text{ mm}$ in size and consists of 40×33 unit cells. We use Rogers 4230 substrate ($\epsilon_r = 3$, $d = 1.524 \text{ mm}$) for the printed strips. Measurements are conducted using the free space method [20] with two horn antennas in an anechoic chamber. Fig. 3.11(b) shows a photograph of the measurement setup, which includes Agilent vector network analyzer N5230 and broadband horn antenna LB-20245. A comparison between measured and simulated transmission and reflection coefficients for $\theta = 0^\circ$ and 40° is shown in Figs. 3.11(c) and (d). A good agreement is observed. The measured -3 dB fractional bandwidths are 12.5%, 8.3% and 12% with center frequencies at 9.55, 13.6 and 17.7 GHz, respectively. Table 3.1 shows a comparison between simulated and measured center frequency and -3 dB fractional bandwidth for each band under the normal and oblique incident angles.

3.5 Dual-Band Third-Order Bandstop FSS

By adding a third identical resonator to the two resonators of the unit cell shown in Fig. 3.2(a) and by adjusting the coupling between them, third-order performance can be achieved. As an example, a dual-band third-order bandstop FSS is designed in this section, whose equivalent block diagram is similar to the

Figure 3.11: (a) Photograph of the fabricated bandstop 3-D FSS. (b) Photograph of the measurement setup. (c) And (d) Simulated and measured transmission and reflection coefficients for different incident angles of the tri-band second-order FSS.

Table 3.1 Simulated and measured center frequency and fractional bandwidth under normal and oblique incident angles of the tri-band second-order FSS.

		Simulated $\theta = 0^\circ$	Measured $\theta = 0^\circ$	Simulated $\theta = 40^\circ$	Simulated $\theta = 40^\circ$
1 st band	Center frequency	9.6 GHz	9.55 GHz	9.6 GHz	9.65 GHz
	-3 dB BW	13.5%	12.5%	13.5%	12 %
2 nd band	Center frequency	13.4 GHz	13.6 GHz	13.4 GHz	13.75 GHz
	-3 dB BW	8.5%	8.3%	8.5%	8.6%
3 rd band	Center frequency	17.5 GHz	17.7 GHz	17.55 GHz	17.8 GHz
	-3 dB BW	12.6%	12%	12.8%	12.3%

one shown in Fig. 3.1(a) with $N = 2$ and $M = 3$, where two groups of three identical resonators are required.

3.5.1 Description of the Structure

Fig. 3.12(a) shows the 3-D view and unit cell details of the proposed dual-band third-order bandstop FSS. It is seen that there are two TLs corresponding to dual-band response, each of which is divided into three identical resonators to achieve the third-order response.

3.5.2 Simulation and Measurement Results

Fig. 3.12(b) shows the simulated transmission and reflection coefficients of the proposed structure for TE polarization under normal and oblique incidences together with the simulated surface current distribution at the normal incidence. As seen from the simulated surface current distribution, each group of the three identical resonators is responsible for one of the two bands. Stable frequency

Figure 3.12: Structure of the proposed dual-band third-order FSS. (a) 3-D structure view with unit cell details. (b) Simulated transmission and reflection coefficients of the dual-band third-order FSS for different incident angles ($b = 6$ mm, $h = 5$ mm, $d = 1.524$ mm, $w = 1$ mm, $S = 2$ mm, $L = L_1 = 11.1$ mm, $g_1 = 0.6$ mm, $L_2 = 6$ mm, $g_2 = 1.2$ mm, $r_{11} = 0.3$ mm, $r_{12} = 0.4$ mm, $r_{13} = 0.4$ mm, $r_{21} = r_{22} = r_{23} = 0.2$ mm, $\epsilon_r = 3$).

(a)

(b)

(c)

Figure 3.13: (a) Photograph of the fabricated bandstop 3-D FSS, (b) and (c) Simulated and measured transmission and reflection coefficients for different incident angles of the dual-band third-order FSS.

response can also be observed here. The simulated results (not shown here) show that the in-band reflected phase for the presented structures is linear decreases with frequency, which ensures the operation as good reflectors. A prototype of this structure is fabricated using Rogers 4230 substrate ($\epsilon_r = 3$, $d = 1.524$ mm), as shown in Fig. 3.13(a).

Table 3.2 Simulated and measured center frequency and fractional bandwidth under normal and oblique incident angles of the dual-band third-order FSS.

		Simulated $\theta = 0^\circ$	Measured $\theta = 0^\circ$	Simulated $\theta = 40^\circ$	Simulated $\theta = 40^\circ$
1 st band	Center frequency	13.2 GHz	13.27 GHz	13.2 GHz	13.3 GHz
	-3 dB BW	16%	15%	16.2%	15.5 %
2 nd band	Center frequency	19.6 GHz	20.25 GHz	19.6 GHz	20.6 GHz
	-3 dB BW	13.3%	12%	13.5%	11.5%

The fabricated prototype dimension is approximately 200 mm $\times$ 200 mm in size and consists of 40 $\times$ 33 unit cells. A good agreement between measured and simulated transmission and reflection coefficients for $\theta = 0^\circ$ and 40° is seen from Figs. 3.13(b) and (c). This structure has a measured -3 dB fractional bandwidths of 15% and 12% with center frequencies at 13.27 and 20.25 GHz, respectively. There is a small frequency shift due to fabrication error, which is more pronounced at high frequencies.

Table 3.2 shows a comparison between simulated and measured center frequency and -3 dB fractional bandwidth for each band under the normal and oblique incident angles.

3.6 Summary

In this chapter, two structures have been presented to demonstrate the general methodology proposed for realizing multiband high-order bandstop frequency-selective structures. The first structure has been designed to exhibit a tri-band second-order response and the second has been designed to achieve a dual-band third-order response. Even-odd mode analysis has been discussed for the coupling between two identical resonators consequently controls the separation

between the two in-band transmission zeros and equivalent circuit model has been constructed. Measured results of the two designs show a very good agreement with simulated ones and stable filtering performance under various angles of incidence. It is expected that the same methodology may also apply to bandpass frequency-selective structure.

Although the 3-D frequency-selective structure shows superior performance, it is polarization sensitive and will be extended to a dual-polarized one in Chapter 6. Those structures are thick, which will be investigated in the next chapter for the bandstop response.

CHAPTER 4

THIN BANDSTOP FREQUENCY-SELECTIVE STRUCTURES BASED ON LOOP RESONATOR

4.1 Introduction

In this chapter, circular and elliptical loop resonators are proposed as the unit cell of 3-D frequency-selective structure aiming to reduce the structure thickness. The structure exhibits a pseudo-elliptic response by utilizing the coupling between the two quasi-TEM modes, which are concentrated in air and substrate regions of the unit cell, respectively. Using the elliptical loop with stepped-impedance resonator, the thickness of the structure can be reduced by 80% compared to previous structures. A comparison between the response of the structure employing the microstrip line as the resonator and the structure using the proposed loop resonators is performed. Three prototypes are designed, fabricated, and tested. The first structure is designed to have a dual-band bandstop response using two half-circular loops with end via holes; it has center frequencies at 10 and 14.3 GHz with -3 dB fractional bandwidths of 6.7% and 6.6%, respectively. A half-elliptical loop with the stepped-impedance resonator is employed in the second structure to achieve more thickness reduction. The third structure is designed to exhibit a dual-polarized bandstop response. Stable frequency response under various angles of incidence is obtained.

4.2 Single-Polarized 3-D FSS Based on Circular Loop

4.2.1 Single-Band Response

Fig. 4.1(a) shows the 3-D structure, unit cell excitation, boundary conditions and detailed dimensions of the 3-D FSS based on the proposed circular loop as the resonator. As shown in Chapter 2 or in [20] that the thickness of the structure is equal to the microstrip line length, but it now becomes equal to the diameter of the loop (D). The proposed structure has the same operating principle as the original structure presented in [20], which depends on the coupling between the two quasi-TEM modes, which are concentrated in the air and the substrate regions to produce transmission and reflection zeros. The unit cell produces two reflection zeroes and one transmission zero. The two reflection zeroes are produced when the propagating path of each mode is equal to π . The one transmission zero occurs when the length of the microstrip line is approximately equal to $0.5 \lambda_{\text{eff}}$ at which microstrip line resonates, where $\lambda_{\text{eff}} = \lambda_0 / \sqrt{(\epsilon_r + 1)/2}$ and λ_0 is the free-space wavelength at the operating frequency.

The circular loop resonates when its circumference equals to λ_{eff} , which can be expressed as:

$$\lambda_{\text{eff}} = \pi \times D \quad (1)$$

where D is the loop diameter. This means that $D = 0.32 \lambda_{\text{eff}}$. Using the circular loop as the resonator in the unit cell, the thickness of the structure is equal to $0.32 \lambda_{\text{eff}}$ instead of $0.5 \lambda_{\text{eff}}$ in the microstrip line case. Fig. 4.1(b) shows the simulated transmission and reflection coefficients for these two 3-D FSSs, one of them employing the microstrip line as the resonator and the other uses circular loop resonator, obtained by the full-wave High-Frequency Structure Simulator (HFSS). The vector surface current distribution at resonance of the proposed resonator is shown. In HFSS simulations, Master and Slave boundary conditions are used to represent the periodic boundary conditions of the unit cell and Floquet ports are used as the excitation ports. The simulated results show a 28.5% reduction in the structure thickness using the circular loop compared to the microstrip line case with the same unit cell parameters. As seen

Figure 4.1: (a) Structure of the proposed FSS using circular loop resonator. (b) Simulated transmission and reflection coefficients of the FSS with circular loop and microstrip line resonator ($b = 8$ mm, $h = 3.524$ mm, $L = D_{\text{loop}} = 7$ mm, $L_{\text{microstrip}} = 9.8$ mm, $d = 1.524$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

from Fig. 4.1(b), the performance of the two 3-D FSSs appears approximately the same when the filtering selectivity is concerned. Because the ring doesn't have open end contributing to radiation loss, it has a higher Q-factor and narrower bandwidth than the microstrip line [101].

4.2.2 Dual-Band Response

By employing different resonators in the same unit cell, a multiband response can be achieved, as proposed in chapter 1. A unit cell for a dual-band response is shown in Fig. 4.2 with the simulated transmission and reflection coefficients as well as simulated vector surface current distribution at the resonance of each band. As shown from the current distribution, each resonator is responsible for one band so that the resonant frequencies are independently controlled. The coupling between resonators is small because each resonator resonates at a different frequency.

4.3 Single-Polarized 3-D FSS Based on Half-Circular Loop

4.3.1 Single-Band Response

The thickness of the structure can be reduced by half by adding two via holes at the position where maximum current resides. Fig. 4.3(a) shows the unit cell of half-circular loop resonator with two via holes at its end. Fig. 4.3(b) shows the simulated transmission and reflection coefficients for circular loop and half-circular loop with two end via holes under TE polarization (E in the y -direction ($\varphi = 0^\circ$)) at which two quasi-TEM modes are excited in the unit cell ($b = 8$ mm, $h = 3.524$ mm, $D_{\text{full loop}} = 7$ mm, $L_{\text{half loop}} = 3.5$ mm, $r = 0.3$ mm, $d = 1.524$ mm, $w = 0.4$ mm, $\epsilon_r = 3$). As seen from the figure, the higher roll-off doesn't change while the bandwidth is larger in the case of the half-circular loop with via holes because its Q-factor is lower due to radiation at its open ends. The upper out-of-band performance is better in the case of the half-circular loop with via holes. The lower roll-off is better for the circular loop since a sharp reflection zero

Figure 4.2: A unit cell of a dual-band FSS with simulated transmission and reflection coefficients ($b = 8$ mm, $h = 3.524$ mm, $D_1 = 7$ mm, $D_2 = 5$ mm).

occurs from the substrate path. This zero is shifted to high frequency in the half-circular loop case because the substrate path becomes smaller. According to the dimensions given, this reflection zero is supposed to occur at 13.5 GHz for the circular loop case, but it becomes 28 GHz in the half-circular loop case because the thickness of the structure is halved.

One of the problems associated with FSS, in general, is the response to the cross-polarized wave. Fig. 4.3(c) shows the simulated cross-polarized transmission and reflection coefficients at different angles of incidence. Here we use $S_{Port1}^{Pol1} S_{Port2}^{Pol2}$ to represent the reflection/transmission coefficients for TE (E in the y -direction)/TM (E in the x -direction) polarizations. Fig. 4.3(d) shows the response of the 3-D FSS employing half-circular loop under normal incidence with different polarizations (φ changes). The desired response can be only achieved when the E field component of the incident wave is perpendicular to the unit cell at which two quasi-TEM modes are excited. This problem will be addressed in Section 4.5. By introducing via holes, another degree of freedom is added for tuning the resonant frequency of the structure by adjusting the via hole radius, as shown in Fig. 4.4. As the via radius decreases its inductance increases, which leads to a reduction in the resonant frequency.

Figure 4.3: (a) Unit cell view of the proposed FSS using half-circular loop resonator. (b) Simulated transmission and reflection coefficients of the FSS with full- and half-circular loop resonators. (c) Simulated cross-polarized transmission and reflection coefficients at different angles of incidence for half-circular loop resonators. (d) Simulated transmission and reflection coefficients under normal incidence with different polarizations for half-circular loop resonators.

Figure 4.4: Simulated transmission and reflection coefficients of the structure shown in Fig. 4.3 with varying via hole radius.

4.3.2 Equivalent Circuit Model

Following the description presented in [20], [86], the unit cell employing microstrip line as a resonator consists of two parallel TLs (air TL and substrate TL), each of which is corresponding to one quasi-TEM mode. For the normal incidence of TE polarization, the Floquet ports with Master and Slave (periodic) boundary conditions can be replaced by wave ports with two perfect electric conductor (PEC) at top and bottom (parallel-plate waveguide) of the unit cell and two perfect magnetic conductor (PMC) at its sides, as shown in Fig. 4.5(a). Basically, the loop consists of two half-wavelength resonators connected in parallel [101]. Due to the unit cell symmetry, it can be divided into two identical parts at AA' plane (magnetic wall), each of which can be modelled as the circuit shown in Fig. 4.5(b). By including PEC boundary conditions, the unit cell height (h) and substrate thickness (d) shown in Fig. 4.3 should be divided by 2.

Figure 4.5: (a) The unit cell of 3-D FSS using half-circular loop resonator in the parallel plate waveguide with PMC sidewalls. (b) The equivalent circuit model. (c) Simulated transmission and reflection coefficients using HFSS and ADS.

The air TL resonates ($\beta_a l = 180^\circ$) at 33 GHz and has characteristic impedance $Z_a = 147 \, \Omega$ with $C_a = 0.009$ pF, which represents the discontinuity between air TL and Port 1/Port 2. The substrate TL resonates ($\beta_s l = 180^\circ$) at 23.6 GHz and has characteristic impedance $Z_s = 80 \, \Omega$ with $C_s = 0.023$ pF representing the discontinuity between substrate TL and Port 1/Port 2. Port 1 and Port 2 have characteristic impedance $Z_p = 165.9 \, \Omega$ of parallel-plate waveguides. The via hole can be modeled as an inductor and its inductance (L_v) can be calculated using the formula [86]

$$L_v = \frac{\mu_0}{2\pi} \left\{ (d/2) \ln 2 \frac{(d/2) + \sqrt{r^2 + (d/2)^2}}{2r} + 1.5 \left(r - \sqrt{r^2 + (d/2)^2} \right) \right\} \quad (2)$$

The characteristic impedance of air and substrate TLs can be approximated as the characteristic impedance of an inverted microstrip line loaded by PEC and microstrip line with PEC cover [102], respectively. A comparison between the simulated transmission and reflection coefficients obtained from HFSS and those obtained from the circuit model using Agilent Advanced Design System (ADS) is shown in Fig. 4.5(c). The difference between them is due to the fact that the transmission line becomes more dispersive when its width becomes thin.

4.3.3 Dual-Band Response

The structure presented in Fig. 4.3 can be extended to achieve dual-band response by adding another resonator in the same unit cell. Fig. 4.6(a) shows the 3-D structure and unit cell details for the dual-band design. The simulated transmission and reflection coefficients for TE polarization under normal and oblique incidences together with the simulated surface current distribution under the normal incidence is shown in Fig. 4.6(b). As shown from the simulated surface current distribution, each resonator is responsible for its own band. A stable frequency response is obtained up to $\Theta = 60^\circ$. It is seen that the performance is acceptable regarding the reflection coefficient that is maintained below -10 dB in between the two bands.

Figure 4.6: (a) Structure of the proposed dual-band 3-D FSS. (b) Simulated transmission and reflection coefficients of the structure with simulated current distribution at each band ($b = 9$ mm, $h = 3.524$ mm, $L = 3.5$ mm, $r_1 = 2.5$ mm, $r = 0.2$ mm, $d = 1.524$ mm, $w = 0.6$ mm, $\epsilon_r = 3$).

Fig. 4.7(a) shows the photograph of the fabricated prototype, which is approximately $190 \text{ mm} \times 190 \text{ mm}$ in size and consists of 20×50 unit cells. Rogers 4230 is used for the substrate material ($\epsilon_r = 3$, $d = 1.524 \text{ mm}$). The fabricated prototype is then measured by the free space method [85] using two horn antennas in an anechoic chamber. Fig. 4.7(b) shows a photograph of the measurement set-up, which includes Agilent vector network analyzer N5230 and two identical broadband horn antennas LB-20245. Fig. 4.7(c) shows a comparison between simulated results and measured data for $\Theta = 0^\circ$ and 40° . Good agreement is observed and Table 4.1 summarizes the simulated and measured resonant frequencies and -3 dB fractional bandwidths (FBW) for the two bands. There is a small frequency shift due to fabrication errors in the loop dimension and via hole radius, which is more critical at high frequencies (small dimensions).

4.4 Single-Polarized 3-D FSS Based on Half-Elliptical Loop

By employing elliptical loop resonator, one can achieve a further reduction in the structure thickness than circular loop resonator. Similarly, the elliptical loop resonates when its circumference equals λ_{eff} . There is no closed-form formula to calculate the circumference of an ellipse. All exact solutions are in infinite series form. These solutions are known by the names Maclaurin, Gauss-Kummer, Cayley, and Euler [103]. There is an approximate expression [104]:

$$\lambda_{\text{eff}} = \pi \left\{ (r_x + r_z) + \frac{3(r_x - r_z)^2}{10(r_x + r_z) + \sqrt{r_x^2 + 14r_x r_z + r_z^2}} \right\} \quad (3)$$

where r_x is the length of the semi-major axis and r_z is the length of the semi-minor axis, as shown in Fig. 4.8(a). For further reduction, SIR can be used to increase the current path. The relation between two resonant frequencies with and without SIR can be expressed as [105]:

$$\frac{f_{0\text{SIR}}}{f_0} = \frac{\tan^{-1} \sqrt{R_z}}{0.5 \times \varphi_2} \quad (4)$$

(a)

(b)

(c)

Figure 4.7: (a) Photograph of the fabricated dual-band 3-D FSS. (b) Photograph of the measurement set-up (inset shows one horn antenna and VNA). (c) Comparison between simulated results and measured data under different incident angles.

Table 4.1 Simulated and measured results for the resonant frequency and fractional bandwidth of 3-D FSS based on half-circular loop under normal incidence.

		Simulated	Measured
1 st band	Center frequency	10 GHz	10.3 GHz
	-3 dB BW	6.7%	7.5%
2 nd band	Center frequency	14.3 GHz	15 GHz
	-3 dB BW	6.6%	6.5%

where $f_{0\text{SIR}}$ is the fundamental resonant frequency using SIR, f_0 is the fundamental resonant frequency without SIR. R_z is the impedance ratio, which equals to Z_2 / Z_1 , where Z_1 and Z_2 are the characteristic impedances of the two segments with corresponding electrical lengths φ_1 and φ_2 , respectively, as shown in Fig. 4.8(b). According to [105], if the resonator is circular then the maximum reduction is achieved when $\varphi_1 = \varphi_2$. In the elliptical resonator case, maximum reduction is achieved when φ_1 and φ_2 equal to the angles created by the intersection of the two diagonals of the rectangle that contains the ellipse. The circular resonator is a special case of the elliptical one at which a circle is contained by a square and the intersection between its diagonals results in two identical angles. Fig. 4.8(c) shows the simulated transmission and reflection coefficients of 3-D FSS employing two resonators shown in Figs. 4.8(a) and (b). It is seen that SIR with elliptical loop resonator has much effect in lowering the resonant frequency, which can be approximately predicted using (3) and (4). The Q-factor of the SIR is high [106], which leads to a decrease in the bandwidth compared to resonator without SIR, as shown in Fig. 4.8(c). The resonant frequency (f_0) of the ellipse resonator shown in Fig. 4.8(a) equals 7.77 GHz using (3) and the resonant frequency employing SIR ($f_{0\text{SIR}}$), Fig. 4.8(b)

Figure 4.8: Ellipse resonator (a) without SIR. (b) With SIR. (c) Simulated transmission and reflection coefficients of the structure employing the two resonators ($b = 13$ mm, $h = 3.524$ mm, $L = 7$ mm, $r_x = 5.5$ mm, $r_z = 3$ mm, e_1 ($r_{x1} = 5.5$ mm, $r_{z1} = 2$ mm), e_2 ($r_{x2} = 6$ mm, $r_{z2} = 3.5$ mm), $w_1 = 0.5$ mm, $w_2 = 3$ mm, $d = 1.524$ mm, $\epsilon_r = 3$).

equals 5.8 GHz using (4), which are very close to the simulated ones 7.88 GHz and 5.7 GHz, respectively.

The resonant frequency can be reduced by increasing the circumference of Ellipse1 (e_1). By employing two via holes at the middle of the resonator, the same resonant frequency can be achieved with half of the thickness of the structure. Moreover, the Q-factor of the resonator is reduced, which leads resultantly to an enhancement in bandwidth.

Fig. 4.9(a) shows the 3-D FSS structure using half-elliptical loop resonator with end via holes and SIR. Fig. 4.9(b) shows the simulated transmission and

Figure 4.9: Structure of the proposed 3-D FSS using half-elliptical loop with SIR resonator.

reflection coefficients for TE polarization under normal and oblique incidences ($b = 13$ mm, $h = 3.524$ mm, $L = 3.5$ mm, e_1 ($r_{x1} = 5.5$ mm, $r_{z1} = 3$ mm), e_2 ($r_{x2} = 6$ mm, $r_{z2} = 3.5$ mm), $w_1 = 0.5$ mm, $w_2 = 3$ mm, $r = 0.2$ mm, $d = 1.524$ mm, $\epsilon_r = 3$). It is seen that the structure exhibits excellent in-band stability regarding the resonant frequency and bandwidth. Fig. 4.9(c) shows the simulated cross-polarized transmission and reflection coefficients at different angles of incidence. As in the half-circular loop case, the desired response can be only achieved when the E field component of the incident wave is perpendicular to the unit cell.

An equivalent circuit model can be constructed by modeling the SIR as an offset step in the microstrip line width (approximated here as a non-offset step). The capacitance and inductance of the step can be expressed by [102]

$$C_{s(s/a)} = 0.00137 \frac{\sqrt{\epsilon_{r2}}}{Z_{02}} \left(1 - \frac{w_1}{w_2}\right) d \left(\frac{\epsilon_{r2} + 0.3}{\epsilon_{r2} - 0.258} \right) \left(\frac{\frac{w_2}{d} + 0.264}{\frac{w_2}{d} + 0.8} \right) \quad (5)$$

$$L_{step} = 0.000987 \left(1 - \frac{Z_{02}}{Z_{01}} \sqrt{\frac{\epsilon_{r2}}{\epsilon_{r1}}} \right)^2 \quad (6)$$

The equivalent circuit is shown in Fig. 4.10(a). There are two stepped impedance circuits: one in air TL and the other in substrate TL. A comparison between the simulated transmission and reflection coefficients using HFSS and those obtained from circuit model using ADS is shown in Fig. 4.10(b). The small frequency shift is due to the approximations made.

Fig. 4.11 compares the filtering performance for different resonators (their description is given in Table 4.2) that can be used in the unit cell of 3-D FSS excluding losses. All structures resonate at 4.5 GHz and use the same substrate with $\epsilon_r = 3$ and $d = 1.524$ mm. Table 4.2 shows a detailed comparison between these resonators regarding the structure thickness, fractional bandwidth (FBW), lower edge ratio $= f_{-3dB}/f_{-10dB}$ and upper edge ratio $= f_{-10dB}/f_{-3dB}$. Lower (upper) edge ratio determines the selectivity sharpness of the filtering performance before (after) the resonant frequency. Regarding the lower roll-off edge, the

Figure 4.10: (a) The equivalent circuit model of the unit cell using half-elliptical loop with SIR, (b) Simulated transmission and reflection coefficients using HFSS and ADS ($Z_p = 102 \, \Omega$, $C_a = 0.008 \, \text{pF}$, $C_s = 0.021 \, \text{pF}$, $L_v = 0.09 \, \text{nH}$, $Z_{aL} = 90 \, \Omega$, $\beta_{aL}l = 180^\circ$ @ 37.5 GHz, $Z_{aH} = 128 \, \Omega$, $\beta_{aH}l = 180^\circ$ @ 56.4 GHz, $Z_{sL} = 34.6 \, \Omega$, $\beta_{sL}l = 180^\circ$ @ 25.5 GHz, $Z_{sH} = 70 \, \Omega$, $\beta_{sH}l = 180^\circ$ @ 43 GHz, $C_{sa} = 0.009 \, \text{pF}$, $L_{1a} = 0.085 \, \text{nH}$, $L_{2a} = 0.197 \, \text{nH}$, $C_{ss} = 0.021 \, \text{pF}$, $L_{1s} = 0.0846 \, \text{nH}$, $L_{2s} = 0.274 \, \text{nH}$).

microstrip line with SIR [92] is the best. However, the half-elliptical loop with end via holes and SIR is the best for the upper edge sharpness and out-of-band performance. Moreover, using half-elliptical loop resonator with SIR, an 80% reduction in the structure thickness is achieved compared to microstrip resonator used in [20]. The periodicity along the x - and y -axis compared to the free-space wavelength at 4.5 GHz are also shown in the table. This small period is an advantage of the 3-D FSS, which makes it less sensitive to oblique incidence and prevents grating lobes. Other loop shapes may be used as the

Figure 4.11: Simulated transmission and reflection coefficients of the 3-D FSS using different resonators.

Table 4.2 Simulated performance comparison between different resonators used in 3-D FSS (all resonate at 4.5 GHz).

Resonator	Structure thickness (mm)	L/λ_0	FBW	Lower edge ratio	Upper edge ratio	h/λ_0	b/λ_0
Resonator I (Microstrip line) [20]	22	0.33	4.7%	1.03	1.08	0.06	0.09
Resonator II (Microstrip line with SIR) [92]	14	0.21	4.2%	1.02	1.12	0.06	0.1
Resonator III (Half-circular loop with end via holes)	8.7	0.13	5.8%	1.08	1.07	0.05	0.27
Resonator IV (Half-elliptical loop with end via and SIR)	4	0.06	5%	1.07	1.04	0.05	0.24
Half-four-legged loaded element [1] with end via and SIR	5	0.075	8.7%	1.12	1.08	0.05	0.16

resonator such as three- and four-legged loaded elements [1]. The half-four-legged element with end via hole and SIR is shown in Table 4.2 for comparison.

To validate the simulated results shown in Fig. 4.9, a prototype is fabricated and measured using the free space method mentioned above, shown in Fig. 4.12(a). The fabricated prototype is approximately $240 \text{ mm} \times 240 \text{ mm}$ in size and contains 18×70 unit cells. We use Rogers 4230 ($\epsilon_r = 3$, $d = 1.524 \text{ mm}$) as the substrate material. Fig. 4.12(b) shows a comparison between simulated results and measured data of the 3-D FSS for $\theta = 0^\circ$ and 40° . Good agreement is obtained, which verifies the advantage of using the half-elliptical loop with SIR as a resonator in the unit cell.

4.5 Dual-Polarized 3-D FSS Based on Loop Resonator

4.5.1 The Structure

The previous examples in Sections 4.3 and 4.4 operate only under TE polarization (E is along the y -direction) at which two quasi-TEM modes are excited in the structure. In this section, a dual-polarized bandstop 3-D FSS based on loop resonator is proposed. In order to achieve a dual-polarized response, the unit cell has to be symmetric ($b = h$). First, we examine the last structure (presented in Fig. 4.9) without SIR (for simplicity) under both polarizations when b and h are equal. Fig. 4.13(a) shows the 3-D unit cell employing half-elliptical resonator with two end via holes. Fig. 4.13(b) shows the simulated transmission and reflection coefficients under TE and TM polarizations. The response for TE polarization is similar to the one shown in Fig. 4.9(b) with a narrower bandwidth and lower out-of-band reflection due to a bigger air region ($h - d$), which leads to more energy coupled to it. Under TM polarization (E in the x -direction), it exhibits a bandstop response similar to the traditional 2-D array [1], [21]. A resonator is formed when the loop constructed from the upper half-ellipse, two via holes and bottom half-ellipse resonates, as illustrated in Fig. 4.13(c). By calculating the effective circumference of this loop, its resonant frequency can be estimated. Fig. 4.13(c) shows the simulated

(a)

(b)

Figure 4.12: (a) Photograph of the fabricated 3-D FSS using half-elliptical loop with end via and SIR. (b) Comparison between simulated results and measured data under different incident angles.

Figure 4.13: (a) Unit cell employing half-elliptical loop resonator. (b) Simulated transmission and reflection coefficients under TE and TM polarizations ($b = 13$ mm, $h = 13$ mm, $L = 3.5$ mm, $r_x = 5.5$ mm, $r_z = 3$ mm, $w = 0.5$ mm, $d = 1.524$ mm, $\epsilon_r = 3$). (c) 2-D top view for the unit cell with simulated vector current distribution at loop resonant frequency (8.6 GHz).

Figure 4.14: (a) Dual-polarized 3-D FSS using half-elliptical loop with SIR resonator. (b) Simulated transmission and reflection coefficients ($b = h = 15$ mm, $L = 3.5$ mm, e_1 ($r_{x1} = 5.5$ mm, $r_{z1} = 3$ mm), e_2 ($r_{x2} = 6$ mm, $r_{z2} = 3.5$ mm), $w_1 = 0.5$ mm, $w_2 = 3$ mm, $r = 0.2$ mm, $d = 1.524$ mm, $\epsilon_r = 3$).

vector current distribution at the resonant frequency under TM polarization, which is consistent with our explanation.

Including SIR and adding one more identical unit cell orthogonal to the original one (Fig. 4.9(a)) to excite the other polarization, a dual-polarized 3-D FSS is constructed and shown in Fig. 4.14(a). Fig. 4.14(b) shows the simulated transmission and reflection coefficients under TE and TM polarizations. There are two resonant frequencies occurring when TE polarization excites x -strips

and y -strips, respectively (TM polarization excites y -strips and x -strips). There are two main problems associated with this dual-polarized structure. The first one is its narrow bandwidth compared to the single-polarization response, which can be solved by increasing the substrate thickness and decreasing its permittivity. The second problem is the bandstop caused by the loop, which degrades the upper out-of-band performance. This can be solved by using a full-elliptical loop with only one via hole at its end.

Fig. 4.15(a) shows the dual-polarized bandstop 3-D FSS with the detailed unit cell dimensions. Using one via hole at the end of a full-elliptical loop has the effect of reducing the resonant frequency to half of its original value. The simulated transmission and reflection coefficients for TE and TM polarization are shown in Fig. 4.15(b) ($b = h = 25$ mm, $L = 5$ mm, $w_1 = 0.3$ mm, $w_2 = 2.2$ mm, e_1 ($r_{x1} = 6.62$ mm, $r_{z1} = 2$ mm), e_2 ($r_{x2} = 7.12$ mm, $r_{z2} = 2.5$ mm), $r = 0.3$ mm, $d = 6.36$ mm, $\epsilon_r = 2.2$). It is seen that the structure exhibits a stable frequency response for TE and TM polarization under oblique incidence. Simulated results show in-band stability up to 85° for both polarizations. Because the structure is symmetric, it exhibits the same response under different polarization angles for TE and TM polarizations. The purpose of adding air holes in the substrate shown in Fig. 4.15(a) is to reduce the effective dielectric constant of the substrate so as to reduce the out-of-band reflection to achieve a better performance. Fig. 4.16 compares the performance of the structure with and without air holes. The resonant frequency is shifted due to the change in the effective dielectric constant and the out-of-band performance is better in the case with air holes.

To construct the structure, the copper ellipse patterns with vias are printed on the substrate. Then cut into long strips with grooves in the mid-way between unit cells, as seen in Fig. 4.17(a). Because a dielectric material with 6.36 mm thickness is not commercially available, two dielectric materials (RT5880 with $\epsilon_r = 2.2$ and $d = 3.18$ mm) are combined. One of them has only a top metallic ellipse pattern and the other has only bottom metallic ellipse pattern. The combination can be done by two ways: the first one is to use a bonding material

Figure 4.15: (a) Structure of the proposed dual-polarized bandstop 3-D FSS using elliptical loop with SIR resonator. (b) Simulated transmission and reflection coefficients of the structure for TE and TM polarization.

Figure 4.16: Comparison between Simulated transmission and reflection coefficients for the structure presented in Fig. 4.13 with and without air holes.

to bond these two materials together then do via hole to connect the top ellipse to the bottom ellipse, FR-4 glue film ($\epsilon_r = 4.3$, $d = 0.1$ mm) could be used as the bonding material. The second way is to do this combination manually by drilling an air hole in these two dielectric materials then put a metal post with the same diameter of this air hole and make soldering in the top and bottom ellipses as shown in Fig. 4.17(b). The second method is used here. This results in a small air gap between these two materials, which will be discussed in Section 4.5.2. Although the first method is more accurate, the second one is cheaper and gives a good result. The strips are then interlocked together through the grooves, as illustrated in Fig. 4.17(c). All these factors are included in the simulated structure presented in Fig. 4.15.

The fabricated prototype is approximately $260 \text{ mm} \times 260 \text{ mm}$ in size and contains 10×10 unit cells. A comparison between the simulated results and measured data for both TE and TM polarizations is shown in Figs. 4.18(a) and (b), respectively. A good agreement is achieved under various incident angles. More FBW can be achieved for all structures by using a thicker substrate with low dielectric constant without affecting the structure thickness.

Figure 4.17: The assembly details and the fabricated structure.

4.5.2 Practical Considerations

Because the structure is manually assembled, some practical considerations are considered here. The first one is the existence of an air gap due to the imperfect combination of the two dielectric materials (Fig. 4.17(b)). Fig. 4.19(a) shows the effect of this air gap (g_a) with different sizes on the performance of the structure. The second consideration is the misalignment between the two elliptical patterns printed on both sides of the strips. Fig. 4.19(b) shows the simulated transmission and reflection coefficients for different values of this misalignment (m_s). The third consideration is the change in the air region of the unit cell ($b - d$) and Fig. 4.19(c) shows the effect of this factor. The last consideration is the loss due to the non-ideal dielectric material, and the effect of this factor is shown in Fig. 4.19(d). More simulations are carried out for both

(a)

(b)

Figure 4.18: (a) and (b) Comparison between simulated results and measured data for the dual-polarized 3-D FSS under different incident angles for TE and TM polarization, respectively.

Figure 4.19: Simulated transmission and reflection coefficients showing the effect of some practical considerations on the performance.

polarizations and other cases. It is concluded that these factors have little influence on the performance of the structure under both polarizations for different angles of incidence.

4.6 Summary

In this chapter, an 80% reduction in the thickness of 3-D frequency-selective structure has been achieved by employing loop resonators instead of microstrip line resonator in the unit cell. A comparison between the microstrip line resonator and the proposed resonator shows comparable performance as far as the bandwidth and selectivity are concerned. An equivalent circuit model has been provided to explain the operating principle of the proposed resonators.

Three structures have been designed and fabricated to verify the advantages of the proposed resonator. The first two structures (dual-band 3-D FSS based on half-circular loop and single-band 3-D FSS based on the half-elliptical loop with SIR) operate only under TE polarization, while the third structure (single-band dual-polarized 3-D FSS based on the elliptical loop with SIR) operates under both TE and TM polarizations. Stable frequency response has been achieved for all structures up to more than 60° oblique incidence.

This chapter has been successfully addressed the thickness problem associated with the bandstop 3-D frequency-selective structure. The following chapter is focused on reducing the thickness of a 3-D FSS with a bandpass response through a folded substrate technique.

CHAPTER 5

THIN BANDPASS 3-D FSS BASED ON FOLDED SUBSTRATE FOR CONFORMAL RADOME APPLICATION

5.1 Introduction

In this chapter, a new method based on a folded substrate is proposed to reduce the thickness of a 3-D bandpass frequency-selective structure (FSS). A 67% and 79% thickness reduction compared to the basic bandpass 3-D FSS is achieved by employing three- and five-layer folding of the substrate, respectively. Single- and dual-polarized designs are presented. By integrating one more substrate with different dielectric constant, dual-band thin structure is achieved. One structure with the five-layer folded substrate is designed, fabricated and measured using a parallel-plate waveguide measurement setup. It has a center frequency of 3.57 GHz with 26.9% transmission bandwidth. The structure thickness equals to $0.06 \lambda_0$, where λ_0 is the free-space wavelength at the center frequency of the passband. Stable frequency response is achieved under oblique incidence. A good agreement is accomplished between simulated and measured results. Moreover, a half-cylindrically shaped radome as an example of a conformal 3-D FSS is constructed based on the thin 3-D FSS and integrated with a broadband horn antenna. Besides working as a radome it shows good filtering properties. The radiation properties of the whole antenna-radome system are finally investigated. A fabricated prototype of this radome is measured in the presence of a broadband horn antenna.

5.2 3-D FSS Based On Folded Substrate

5.2.1 Principle of Operation

Fig. 5.1(a) shows the 3-D view of the unit cell for the bandpass 3-D FSS presented in [86] and Fig. 5.1(b) shows the 3-D view of the unit cell based on a folded substrate with three-layer folding. The simulated reflection and transmission coefficients for the two unit cells are shown in Figs. 5.1(c) and (d). As shown in the figures, the center frequency of the transmission band is reduced by employing the folded substrate principle with the same thickness. The simulated results are obtained by using a full-wave simulator CST Microwave Studio (CST-MWS). These two designs presented in Fig. 5.1(a) and (b) have the same substrate material and thickness. By utilizing the folded substrate, the transmission path is increased and the transmission frequency band is reduced. The center frequency of the structure shown in Fig. 5.1(a) is 7.23 GHz and its -3 dB bandwidth is 38%. While the center frequency of the structure with three-layer folded substrate shown in Fig. 5.1(b) is 2.4 GHz with -3 dB bandwidth of 31.7%. This means that a 67% thickness reduction between the original structure, Fig. 5.1(a), and the one with the folded substrate, Fig. 5.1(b), is achieved utilizing a three-layer folded substrate. Under normal incidence, the passband ripple equals to 0.2 dB for the three-layer folded substrate.

The two transmission zeros at 12 and 16 GHz in Fig. 5.1(c) are produced due to the reflection from the two paths resulting from blocking the air region with a metallic block [86]. There are no transmission zeros presented in the folded substrate case because the frequencies of the transmission zeros are beyond the second harmonic of the transmission frequency. Fig. 5.2 shows the simulated vector electric field distribution for both unit cells shown in Fig. 5.1(a) and (b) at 6.7 and 2.25 GHz, respectively. As shown in the figure, the electric field propagation path follows the folding profile.

Figure 5.1: Proposed methodology. (a) 3-D view of the unit cell for bandpass 3-D FSS [86]. (b) 3-D view of the unit cell for bandpass 3-D FSS with three-layer folded substrate. Simulated reflection and transmission coefficients for (c) 3-D FSS [86], (d) 3-D FSS with three-layer folded substrate ($b = 5$ mm, $h_a = 2$ mm, $L = 10.5$ mm, $d = 1.524$ mm, $w = 3$ mm, $L_b = 4$ mm, $t_b = 1.2$ mm, $D_{v1} = 0.5$ mm, $D_{v2} = 0.06$ mm, $\epsilon_r = 3.55$).

Figure 5.2: Simulated vector electric field distribution for both unit cells shown in Fig. 5.1(a) and (b) at 6.7 and 2.25 GHz.

5.2.2 Five-layer Folding

Increasing the number of folding layers leads to more reduction for the transmission frequency band, which indicates more thickness reduction. Fig. 5.3(a) shows the 3-D view of the 3-D FSS and unit cell with a five-layer folded substrate. The simulated reflection and transmission coefficients under the normal and oblique incidences are shown in Fig. 5.3(b). The parallel-plate waveguide with a middle via hole can be analyzed by even-odd mode analysis method [88]. This variation with respect to the incident angle is attributed to the change of the electrical length of the parallel-plate waveguide viewed from the oblique incident angle. At a large incident angle, a smaller effective electrical length is obtained, and the second transmission pole (odd-mode resonant frequency) is shifted to higher frequency resulting in an increase of the in-band reflection level. The effect of oblique incidence is stronger on the odd-mode resonant frequency because the even-mode resonant frequency is due to half-TL shorted by a via hole, which forces a certain condition at the TL end. This response is achieved under the TE-polarized wave (E in the y -direction), which will be extended to a dual-polarized structure in Section 5.3. The center

Figure 5.3: (a) 3-D view of the unit cell for bandpass 3-D FSS with five-layer folded substrate. (b) Simulated reflection and transmission coefficients under the normal and oblique incidences ($b = 6$ mm, $L = 10.5$ mm, $d = 1.524$ mm, $w = 4$ mm, $D_v = 0.16$ mm, $\epsilon_r = 3.55$).

frequency is 1.53 GHz with -3 dB bandwidth of 26%. This means that 79% thickness reduction is accomplished utilizing five-layer substrate folding. Under normal incidence, the passband ripple equals to 0.11 dB. It is worth mentioning that the air region and the metallic block have been removed in the structure shown in Fig. 5.3 because they have no effect on the passband and their removal can make the structure simpler. They were employed to produce two

transmission zeros in the design presented in [86]. The transmission loss in three situations (no-folded substrate, three-layer folded substrate, and five-layer folded substrate) is investigated. It is found that the average transmission loss for no-folded substrate is 0.19 dB, three-layer folded substrate is 0.31 dB, and five-layer folded substrate is 0.34 dB. The difference is not that big because as the number of folds increases, the working frequency decreases, at which the dielectric loss decreases as well.

5.2.3 Implementation and Measurement

In this study, the parallel-plate waveguide measurement setup [69] is employed to conduct the measurement of reflection and transmission coefficients of the folded structure, as shown in Fig. 5.4(c). A conical metallic disk is utilized to achieve the transition from a coaxial cable to parallel-plate waveguide [69]. In order to excite the dominant TEM mode inside the parallel-plate waveguide, the height of the cone is selected to be 10 mm. The purpose of the tapering section is to gradually transform the separation from 10 mm at the source to 24 mm at the FSS under test.

Because the measurement setup is suitable for measurement from 1.8 GHz to 15 GHz, the thickness (L) of the FSS presented in Fig. 5.3 is changed ($L = 5$ mm) to accommodate this requirement. To build the structure, the multi-layered technique is employed. The two sidewalls of substrates can be fabricated with either via holes or deposit copper on the side of substrates. The second method is used in our fabrication. The size of the measurement setup at the side of the structure under test is 24 mm $\times$ 144 mm (height $\times$ width) and the size of the unit cell presented in Fig. 5.3 is 7.8 mm $\times$ 6 mm. Therefore, 3 rows and 25 columns are required to construct the 3-D FSS to be measured. Fig. 5.4(a) shows the top and bottom views of the fabricated fifteen substrates required to build the FSS. These substrates are combined together with Nylon screws and nuts with M4 size then soldering is performed at the two sides of the combined substrates to construct sidewalls, as shown in Fig. 5.4(b). A comparison between the measured and simulated reflection and transmission coefficients is shown in Fig. 5.4(d). A good agreement is achieved between the simulated and measured

Figure 5.4: Experimental verification. (a) Photograph of the top and bottom view of the fabricated substrates. (b) Photograph of 3-D FSS constructed by combined substrates. (c) Photograph of the measurement setup (the upper plate is removed to clearly see inside). (d) A comparison between measured and simulated reflection and transmission coefficients ($b = 6$ mm, $L = 5$ mm, $d = 1.524$ mm, $w = 4$ mm, $D_v = 0.5$ mm, $\epsilon_r = 3$).

transmission and reflection coefficients. The difference between them is due to the tolerance in the fabrication and assembly. More transmission loss is noticed in the measurement results, which is due to the assembly method and soldering.

5.3 Dual-Polarized and Dual-Band FSSs

5.3.1 Dual-Polarized Structure

The 3-D FSS proposed in the previous section can be easily extended to dual-polarized FSS, as shown in Fig. 5.5. The unit cell is divided into four sub-cells and arranged in rotating fashion. Here, the three-layer folded substrate is utilized to show the possibility to extend the single-polarized FSS to dual-polarized one. The simulated reflection and transmission coefficients under the normal and oblique incidences for TE and TM polarizations are shown in Fig. 5.5(b). It is observed that the structure exhibits a stable frequency response under oblique incidence for both TE and TM polarizations. As discussed in Section 5.2.2 under TE polarization, the high-frequency transmission pole moves to a higher frequency as the angle of incidence increases, which leads to an increase in the in-band reflection level. Under TM polarization, the coupling between the two resonances decreases leading to a decrease in the in-band reflection level. It has a center frequency at 2.8 GHz with -3 dB bandwidth of 18.6%. Under normal incidence, the passband ripple equals to 0.14 dB.

5.3.2 Dual-Band Structure

Dual-band response can also be accomplished by adding one more substrate with a different dielectric constant to act as another passband path at another frequency. Fig. 5.6 shows the unit cell and simulated reflection and transmission coefficients under the normal and oblique incidences. As shown, a dual-band bandpass response is achieved. This structure can be fabricated and measured utilizing the same method presented in the previous section. It exhibits center frequencies at 3.3 and 6.2 GHz with -3 dB bandwidths of 35% and 13.5%, respectively. The bandwidths of the two bands can be controlled by

Figure 5.5: Dual-polarized 3-D FSS based on folded substrate. (a) 3-D view of dual-polarized unit cell. (b) Simulated reflection and transmission coefficients under oblique incidence for TE and TM polarizations ($b = 4.68$ mm, $L = 10.5$ mm, $d = 1.524$ mm, $w = 3$ mm, $D_v = 1.4$ mm, $\epsilon_r = 3.55$).

Figure 5.6: Dual-band 3-D FSS. (a) 3-D view of the unit cell. (b) Simulated reflection and transmission coefficients under the normal and oblique incidences ($b = 6$ mm, $L = 8$ mm, $d_1 = 1.524$ mm, $d_2 = 1.27$ mm, $w = 5$ mm, $D_{v1} = 0.25$ mm, $D_{v2} = 2.2$ mm, $\epsilon_{r1} = 3$, $\epsilon_{r2} = 11.2$).

adjusting the diameter of the via in each path, which controls the coupling between resonators. Under normal incidence, the passband ripple equals to 0.17 dB for the lower band, while it is equals to 0.05 dB for the upper band.

5.4 Conformal Radome Based on the 3-D FSS

5.4.1 Description of the Structure

As shown in the previous sections, thin 3-D bandpass FSS is achievable employing the folded substrate technique, which makes it suitable for practical

applications. As an example, the 3-D bandpass FSS shown in Fig. 5.4 can be employed as a conformal radome with excellent filtering performance. For simplicity, a half-cylindrically shaped radome is designed. The 3-D and top views of half-cylindrically shaped radome based on folded substrate integrated with a horn antenna are shown in Fig. 5.7. The radome radius is chosen 100 mm to be suitable for the horn antenna available in our lab. The broadband horn antenna is A-INFO LB-20245 with an aperture size of 84 mm $\times$ 64 mm. The horn antenna illustrated in Fig. 5.7 consists of two parts. The first part (*Part I*) is a pyramidal horn antenna with aperture size equal to the broadband horn antenna in the lab. The second part (*Part II*) is an extension to make the beamwidth of the horn antenna narrower, which lowers the side lobe level of the whole antenna-radome system as detailed in the next section.

5.4.2 Simulated Results

The simulated reflection coefficient for the horn antenna alone, shown in Fig. 5.7, and the whole antenna-radome system is shown in Fig. 5.8(a). The whole system shows good matching in the transmission band of the FSS shown in Fig. 5.4. The two reflection dips shown in Fig. 5.8(a) are corresponding to the two transmission poles of the radome. Fig. 5.8(b) shows the realized gain at $\theta = 0^\circ$ direction with and without the radome. As shown, the whole system demonstrates a sharp filtering property. To validate the simulated result, the simulated normalized gain, which is the difference between the two curves shown in Fig. 5.8(b), is compared with the simulated transmission coefficient of the infinitely large planar 3-D FSS illustrated in Fig. 5.4, which indicates an excellent agreement.

The simulated far-field radiation patterns for the horn antenna with and without radome at four different frequencies are shown in Fig. 5.9. The results are shown in two different cut planes: the *E*-plane (*yz* plane) and the *H*-plane (*xz* plane). The results show that the two systems have very close far-field radiation patterns for both *E*- and *H*-planes. The side-lobe level in the *E*-plane for the antenna-radome system is little higher than the horn antenna alone, which is caused by the edges of the radome along the *y*-axis as illustrated in the next

Figure 5.7: Half-cylindrical radome integrated with a broadband horn antenna. (a) 3-D view. (b) Top view ($b_1 = 42$ mm, $b_2 = 66$ mm, $b_3 = 100$ mm, $L = 5$ mm, $L_1 = 50$ mm, $L_2 = 79$ mm, $L_3 = 21$ mm, $a_1 = 62$ mm, $a_2 = 86$ mm, $a_3 = 100$ mm, $r_{\text{radome}} = 100$ mm, $d = 1.524$ mm, $D_v = 0.3$ mm, $\epsilon_r = 3$).

(a)

(b)

(c)

Figure 5.8: Simulated results of the design shown in Fig. 5.7. (a) Reflection coefficient for the horn antenna with and without radome. (b) Realized gain at $\Theta = 0^\circ$ direction for the horn antenna with and without radome. (c) A comparison between difference in realized gain shown in (b) and transmission coefficient of infinite planar structure excited by plane wave.

Figure 5.9: Simulated far-field radiation patterns for the horn antenna with (solid-gray) and without (dash-black) radome in E -cut plane (left) and H -cut plane (right) at, (a) 3.1 GHz. (b) 3.3 GHz. (c) 3.5 GHz. (d) 3.7 GHz.

Figure 5.10: Three different horn antenna. (a) Open-ended waveguide. (b)

Pyramidal horn antenna with open end equal to the one in our lab. (c)

Pyramidal horn antenna in (b) with an extension with bigger open end. ($b_1 = 42$ mm, $b_2 = 66$ mm, $b_3 = 100$ mm, $L_1 = 50$ mm, $L_2 = 79$ mm, $L_3 = 21$ mm, $a_1 = 62$ mm, $a_2 = 86$ mm, $a_3 = 100$ mm).

section. There is an asymmetry in the radiation pattern of the horn antenna with the radome in the E -plane radiation pattern because the two edges of the radome along the y -axis are not fully identical, as seen from the enlarged pictures for these two edges shown in Fig. 5.7(a). The difference between the two edges along the y -axis affects the E -plane radiation pattern, but not the H -plane radiation pattern.

5.5 Horn Antenna

In this section, the effect of the horn antenna's radiation properties on the performance of the antenna-radome system is investigated. More specifically, the side-lobe level of the whole system employing three antenna types is illustrated. The antennas, shown in Fig. 5.10, are an open-ended waveguide,

pyramidal horn I, and pyramidal horn II. Pyramidal horn I is a horn antenna with an aperture size equal to the one in our lab A-INFO LB-20245. Pyramidal horn II is pyramidal horn I with an extension to make the beamwidth narrower.

A comparison between simulated far-field radiation patterns for these three antennas integrated with radome (Fig. 5.7) at three different frequencies is shown in Fig. 5.11. Two different cut planes are shown and labeled as E - and H -planes, respectively. The results show that as the aperture of the antenna increases, the lower is the side-lobe level. This is due to two main reasons: (i) the side-lobe level from the antenna itself; (ii) the reflection and diffraction from the ends of the radome along the y -axis. As the aperture of the antenna is increased, the narrower is the 3-dB beamwidth of the antenna, which lowers the influence of the edges of the radome. Fig. 5.12 shows the simulated electric field distribution at 3.5 GHz for the system shown in Fig. 5.7 employing three types of antennas illustrated in Fig. 5.10. This provides an evidence of the previously interpreted reasons about the impact of the side-lobe level of the antenna and the edges of the radome on the side-lobe level of the whole antenna-radome system. Because the electric field distribution is simulated at a frequency in the passband of the radome, the incident electric field passes through the radome in the three cases. Fig. 5.13 shows the simulated electric field distributions at 2.8 and 5 GHz. At these frequencies, the radome works as a reflector.

5.6 Radome Fabrication and Measurement

The conformal 3-D FSS radome can be fabricated using multi-layer technology, as shown in Fig. 5.4. The only difference is the shape of slices. As illustrated in Fig. 5.4(a), slices are planar because the structure is planar. To fabricate this half-cylindrically shaped radome, the shape of the slices is semi-circular, as shown in Fig. 5.14(a). The length of the radome is 319.5 mm and the slice thickness is 1.524 mm. Therefore, 213 semi-circular slices are needed. There is one drawback of the proposed design, which is the consumption of substrate material because of its circular shape. Fig. 5.14(b) illustrates one-set of 5 slices, which are then combined together to form one vertical unit cell. The whole

Figure 5.11: Simulated far-field radiation patterns for the three horn antennas shown in Fig. 5.10(a) blue curve with triangle symbol, Fig. 5.10(b) orange curve with circular symbol, and Fig. 5.10(c) black curve with square symbol in E -cut plane (left) and H -cut plane (right) at, (a) 3.3 GHz. (b) 3.5 GHz. (c) 3.7 GHz.

Figure 5.12: Simulated electric field distribution at 3.5 GHz for the system shown in Fig. 5.7 employing three types of antennas shown in Fig. 5.10.

Figure 5.13: Simulated electric field distribution at 2.8 and 5 GHz utilizing the antenna shown in Fig. 5.10(c).

Figure 5.14: Fabricated prototype of the cylindrical radome. (a) Top and bottom view of semi-circular slices. (b) 3-D view of one-set of 5-slices. (c) Half cylindrical radome integrated with the horn antenna.

Figure 5.15: Measured results of the prototype shown in Fig. 5.14. (a) Reflection coefficient for the horn antenna with and without radome. (b) A comparison between normalized gain of the horn antenna with and without radome, and transmission coefficient of infinite planar structure excited by plane wave.

radome is constructed from 41 sets, as shown in Fig. 5.14(c). This radome is integrated with a wideband horn antenna with an extension part.

The measured reflection coefficient results of the horn antenna with and without the fabricated radome are shown in Fig. 5.15(a). As predicted by the simulation, the whole system shows a good matching in the transmission band of the radome. Two reflection zeros in the bandwidth of the radome are

corresponding to the two reflection zeros of the 3-D FSS shown in Fig. 5.4. Fig. 5.15(b) shows the measured normalized gain, which is the difference between the realized gain of the horn antenna with and without the radome at the $\Theta = 0^\circ$ direction. As shown, the whole system demonstrates the filtering property of the radome. The differences between these two curves are due to the fabrication tolerance, assembly errors, and misalignment between the center of the horn and the center of the radome in the measurement.

The measured far-field radiation patterns of the horn antenna with and without the curved radome at three different frequencies are shown in Fig. 5.16. The results are shown in two different cut planes: the E -plane (yz plane) and the H -plane (xz plane). The results show that these two systems have very close far-field radiation patterns for both E - and H -planes. The side-lobe level in the E -plane for the antenna-radome system is a little higher than the horn antenna alone, which is caused by the edges of the radome along the y -axis, as predicted by the simulation. For both planes, the high back lobe level is due to the factors related to the fabrication and measurement imperfections.

5.7 Summary

In this chapter, thin bandpass 3-D frequency-selective structure has been presented based on a folded substrate. It has been demonstrated that a 79% thickness reduction is possible by employing five-layer folding compared to the basic structure. By increasing the folding layers, more thickness reduction can be achieved. The fabricated prototype has been measured with the parallel-plate measurement setup and measured data show good agreement with the simulated results. Dual-polarized and dual-band versions have been introduced and they demonstrate a stable performance under oblique incidence. A half-cylindrically shaped radome with 3-D frequency-selective structure has been integrated with a broadband horn antenna. The effect of the beamwidth of the horn antenna has been investigated. A fabricated prototype has been tested to validate the simulated results.

Figure 5.16: Measured far field radiation patterns for the horn antenna with (solid-gray) and without (dash-black) radome in E-cut plane (left) and H-cut plane (right) at, (a) 3.4 GHz. (b) 3.6 GHz. (c) 3.8 GHz.

As demonstrated in the previous two chapters, the thickness problem of the 3-D frequency-selective structure has been solved by employing loop resonator and folded substrate technique for the bandstop and bandpass responses, respectively. The next chapter is to physically investigate the second problem associated with bandstop and absorber based on 3-D frequency-selective structure, which is their operation under single polarization.

CHAPTER 6

DOUBLE-SIDED PARALLEL-STRIP LINE RESONATOR FOR DUAL-POLARIZED 3-D FREQUENCY-SELECTIVE STRUCTURE AND ABSORBER

6.1 Introduction

A double-sided parallel-strip line (DSPSL) resonator is employed in the unit cell of a 3-D frequency-selective structure in order to exhibit the desired response (bandstop, bandpass and absorption) under an incident wave of arbitrary polarization. It is found that this unit cell supports the propagation of three quasi-TEM modes. The first and third modes are concentrated in the substrate and air regions of the unit cell, respectively. While the second mode is concentrated in the interface between the air and substrate of the unit cell. The polarization of the second mode is perpendicular to the polarization of the first and third modes. Detailed analysis of these propagating modes is provided with equivalent circuit models to explain the operating principle. A dual-polarized bandstop 3-D frequency-selective structure is then proposed. It exhibits a stable frequency response with a constant out-of-band bandwidth under a large range of oblique incidence angle. A dual-polarized absorber is also realized employing a unit cell with double-sided parallel-strip line resonator. By properly exciting and absorbing the second mode of the modified unit cell, an ultra-wideband absorber is constructed, which has -10 dB fractional bandwidth of 148% from 1.51 GHz to 10.1 GHz with an absorber thickness of $0.1 \lambda_0$ at the lowest absorption frequency.

6.2 Microstrip Line Unit Cell

6.2.1 The Operating Principle

Although the microstrip line unit cell was described in [20], an extended explanation is provided regarding its response for TM-polarized wave and the first high-order mode propagating in the unit cell. Fig. 6.1(a) shows the single-polarized unit cell employing microstrip line as the resonator with its excitation and boundary conditions included. It consists of a vertically placed microstrip line with b and h to represent the period along the x - and y -axis, respectively. The simulated reflection and transmission coefficients under the normal incidence ($\Theta = 0^\circ$) for TE and TM polarizations are shown in Fig. 6.1(b) and they are obtained by using the full-wave simulator CST Microwave Studio (CST-MWS). When the electric field component of the incident wave is perpendicular to the microstrip line [TE polarization, E in the y -direction ($\varphi = 90^\circ$)], two quasi-TEM modes are excited in the unit cell. One is concentrated in the substrate region and the other is concentrated in the air region [22], as shown in Figs. 6.2(a) and (b), respectively. The dispersion diagram of the unit cell presented in Fig. 6.1(a) is shown in Fig. 6.2(d) obtained by CST-MWS eigenmode solver. As shown in Fig. 6.1(b), this unit cell exhibits two reflection zeroes and one transmission zero. The two reflection zeroes are produced when the propagating path of each quasi-TEM mode is equal to π [points $A1$ and $B1$ in Figs. 6.1(b) and 6.2(d)]. The transmission zero occurs when the length of the microstrip line is approximately equal to $0.5 \lambda_{\text{eff}}$ at which the microstrip line section resonates as detailed in [20], [86], where $\lambda_{\text{eff}} = \lambda_0 / ((\epsilon_r + 1)/2)^{1/2}$ and λ_0 is the free-space wavelength at the operating frequency.

Under TM polarization [E in the x -direction ($\varphi = 0^\circ$)], this unit cell prevents the TM-polarized wave from propagating up to a certain frequency because its plates are parallel to the electric field component of the incident wave. The electric field distribution of the first high-order mode is shown in Fig. 6.2(c). It is similar to the electric field distribution of TE₁ mode of a parallel-plate waveguide and its cut-off frequency can be calculated by [107]

Figure 6.1: (a) 3-D View of the unit cell with microstrip line as the resonator. (b) Simulated reflection and transmission coefficients of the FSS under TE and TM polarizations ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

Figure 6.2: Electric field patterns of the first three modes in the microstrip line unit cell. (a) First mode; (b) Second mode; (c) Third mode. (d) Dispersion diagram of the first three modes supported by the unit cell ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

$$f_{cTE1} = \frac{c}{2h\sqrt{\epsilon_{\text{eff}}}} \quad (1)$$

where c is the speed of light in free space. Its reflection zero occurs when its propagating path is equal to π [point C1 in Figs. 6.1(b) and 6.2(d)]. The difference in frequency values between Fig. 6.1(b) and Fig. 6.2(d) is due to the effect of the source loading in the frequency-domain solver. This unit cell can exhibit the desired frequency response only under single-polarization, at which

the electric field component of the incident wave is perpendicular to the microstrip line. In order to achieve a dual-polarized response, the unit cell has to be symmetric along both x - and y -axis. This means that one has to add one more identical unit cell orthogonal to the original one to excite the other polarization. Unfortunately, this symmetric arrangement leads to a square waveguide that has cut-off frequencies for modes of both polarizations due to the ground plane of the microstrip lines. By replacing these ground planes with strips, one can solve this problem and excite quasi-TEM waves under both polarizations. More detailed analysis of this modified unit cell is given in Section 6.3.

6.2.2 Equivalent Circuit Model

Following the description provided above and studies in [20], [86] the unit cell employing microstrip line as a resonator can be modeled as two parallel TLs (air TL and substrate TL), each of which is corresponding to one quasi-TEM mode. Under the normal incidence of TE polarization, the periodic boundary conditions can be replaced by two perfect electric conductors (PEC) along the y -direction (parallel-plate waveguide) and two perfect magnetic conductors (PMC) along the x -direction of the unit cell. The circuit model of the microstrip unit cell associated with simulated reflection and transmission coefficients using CST and Agilent Advanced Design System (ADS) are shown in Fig. 6.3. The air TL resonates ($\beta_a l = 180^\circ$) at 13.9 GHz and has characteristic impedance $Z_a = 233.7 \, \Omega$ with $C_a = 0.0053 \, \text{pF}$ that represents the discontinuity between Port 1/Port 2 and air TL. The substrate TL resonates ($\beta_s l = 180^\circ$) at 9.6 GHz and has characteristic impedance $Z_s = 81.6 \, \Omega$ with $C_s = 0.017 \, \text{pF}$ representing the discontinuity between Port 1/Port 2 and substrate TL. The values of C_a and C_s are calculated following the procedures detailed in [86]. Port 1 and Port 2 are two parallel-plate waveguides with characteristic impedance $Z_p = 377 \, \Omega$. The characteristic impedance of air and substrate TLs are extracted from CST [22].

Figure 6.3: (a) Equivalent circuit model of the unit cell presented in Fig. 6.1 for TE-polarized wave. (b) Simulated reflection and transmission coefficients using CST and ADS.

6.3 Double-Sided Parallel-Strip Line Unit Cell

6.3.1 The Operating Principle

Fig. 6.4 shows the modified unit cell (the ground planes are replaced in the microstrip line unit cell with narrow strips) with simulated reflection and transmission coefficients under the normal incidence for TE and TM polarizations. Here, the unit cell is labeled as y-oriented unit cell because it exhibits the desired response when the electric field component of the incident

Figure 6.4: (a) 3-D View of the unit cell with double-sided parallel-strip lines as the resonator. (b) Simulated reflection and transmission coefficients of the FSS under TE and TM polarizations. ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

wave is in the y -direction. As shown in the figure, the TM polarization can propagate through the modified unit cell. The dispersion diagram of the modified unit cell is shown in Fig. 6.5 associated with the electric field

Figure 6.5: Electric field patterns of the first three modes in the double-sided parallel-strip line unit cell. (a) First mode; (b) Second mode; (c) Third mode. (d) Dispersion diagram of the first four modes supported by the unit cell ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

distribution of the first three modes. There are three quasi-TEM modes supported by this unit cell. The first and third modes are similar to the two quasi-TEM modes excited in the microstrip line unit cell (y-polarized, one concentrated in the substrate and the other concentrated in air region of the unit cell) as shown in Figs. 6.5(a) and (c). These two modes naturally couple the incident TE-polarized wave to the unit cell. The second mode is propagating between strips in adjacent unit cells (x-polarized, concentrated in the interface between the substrate and air regions of the unit cell), as shown in Fig. 6.5(b). It

couples the incident TM-polarized wave to the unit cell, as shown in Fig. 6.4(b). There are three reflection zeroes, two are due to TE polarization and one due to TM polarization, produced when the propagating path of each mode is equal to π [points A2, B2 and, C2 in Figs. 6.4(b) and 6.5(d)].

As shown in Fig. 6.4(b), the reflection level under TM polarization is higher than the reflection level under TE polarization, which leads to a problem in the dual-polarized design, as will be explained in Section 6.3.3.

6.3.2 Equivalent Circuit Model

Under TE polarization, the circuit model of the unit cell shown in Fig. 6.4 is similar to the one shown in Fig. 6.3 because it supports two quasi-TEM modes concentrated in the air and substrate regions. There are small differences regarding the characteristic impedance and resonant frequency because the two unit cells are different: one is based on a microstrip line and the other one is based on DSPSL. A comparison between simulated results from CST and ADS is shown in Fig. 6.6(a). The air TL resonates ($\beta_a l = 180^\circ$) at 13.5 GHz and has characteristic impedance $Z_a = 114.6 \Omega$ with $C_a = 0.0098$ pF. The substrate TL resonates ($\beta_s l = 180^\circ$) at 9.7 GHz and has characteristic impedance $Z_s = 37.8 \Omega$ with $C_s = 0.025$ pF. Port 1 and Port 2 have characteristic impedance $Z_p = 188.5 \Omega$. Under TM polarization, the periodic boundary conditions can be replaced by two PECs along the x -direction (parallel-plate waveguide) and two PMCs along the y -direction of the unit cell. One quasi-TEM mode is excited, which can be modelled as one TL shown in Fig. 6.6(b). The characteristic impedance of this TL is equal to 502Ω (extracted from CST) and it resonates at 11.1 GHz.

6.3.3 Dual-Polarized Bandstop 3-D FSS

This modified unit cell can be extended to operate under the dual-polarized electromagnetic wave. Fig. 6.7(a) shows the 3-D view of the dual-polarized unit cell with an alternative model to be used to construct the equivalent circuit. The simulated reflection and transmission coefficients under the normal incidence for TE and TM polarizations are shown in Fig. 6.7(b). It is seen that the new

Figure 6.6: Equivalent circuit model with the simulated reflection and transmission coefficients using CST and ADS. (a) Under TE polarization. (b) Under TM polarization.

unit cell exhibits the desired response under TE and TM polarizations. Under TE polarization, the circuit model consists of three different TLs corresponding to the three quasi-TEM modes supported by this unit cell. Two of these modes are excited in the y -oriented unit cell and one is excited in the x -oriented unit cell. Fig. 6.8(a) shows the equivalent circuit model of the dual-polarized unit cell. A comparison between the simulated reflection and transmission coefficients using CST and ADS is shown in Fig. 6.8(b) and they show a very good agreement. The small difference between these results comes from the inaccurate calculation of the circuit model components, especially the air and substrate TLs characteristic impedance and edge capacitors.

A problem associated with this structure is the high out-of-band reflection level. This is mainly due to the reflection from the x/y -oriented unit cell under the TE/TM-polarized wave, respectively, as stated at the end of Section 6.3.1. It can be reduced by utilizing a substrate with low dielectric constant. Because the

Figure 6.7: (a) Dual-polarized unit cell employing double-sided parallel-strip lines as the resonator. (b) Simulated reflection and transmission coefficients of the FSS under TE and TM polarization ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 0.4$ mm, $\epsilon_r = 3$).

Figure 6.8: (a) Equivalent circuit model of the unit cell shown in Fig. 6.7. (b) Simulated reflection and transmission coefficients using CST and ADS ($Z_p = 377 \Omega$, $C_a = 0.0092$ pF, $C_s = 0.036$ pF, $Z_a = 247.6 \Omega$, $\beta_a l = 180^\circ$ @ 13.9 GHz, $Z_s = 81.6 \Omega$, $\beta_s l = 180^\circ$ @ 9.6 GHz, $C_x = 0.0005$ pF, $Z_x = 537.2 \Omega$, $\beta_x l = 180^\circ$ @ 11.5 GHz).

lowest dielectric constant we have is 2.2, air holes can be drilled in the substrate to reduce the equivalent dielectric constant. The structure thickness can be halved by employing via hole at the end of the strip. Fig. 6.9(a) shows the 3-D structure and detailed dimensions of the dual-polarized unit cell employing air holes to improve the out-of-band reflection level and via hole to reduce the

Figure 6.9: (a) Structure of the proposed dual-polarized 3-D FSS. (b) Simulated reflection and transmission coefficients of the FSS under normal and oblique incidences for TE and TM polarization ($b = 8$ mm, $h = 8$ mm, $L = 10$ mm, $d = 3$ mm, $w = 1$ mm, $\epsilon_r = 2.2$).

structure thickness by half. Fig. 6.9(b) shows the simulated reflection and transmission coefficients under normal and oblique incidences for TE and TM polarizations ($\Phi = 0^\circ$ and Θ changes in the x - z plane). A very stable frequency response is achieved due to small unit cell size compared to the operating wavelength ($0.16 \lambda_0$). The simulated results (not shown here) show the stability of resonant frequency and fractional bandwidth under oblique incidence up to 85° . It is worthwhile mentioning that the first harmonic is fixed at its location under a large range of oblique incidence angle, which makes the upper out-of-band transmission bandwidth unchanged under oblique incidence.

6.4 Dual-Polarized Dual-Band Second-Order Bandstop 3-D FSS

6.4.1 The Structure

As mentioned in Chapter 3, in order to design a dual-band second-order bandstop FSS, one needs $N = 2$ and $M = 2$ for the equivalent block diagram shown in Fig. 3.1(a). This indicates that two groups of two identical resonators are required. Fig. 6.10(a) shows the 3-D view and unit cell details of the proposed dual-band second-order bandstop FSS. There are two TLs corresponding to the dual-band response, each of which is divided into two identical resonators to achieve a second-order response. The purpose of the air holes shown in Fig. 6.10 is to reduce the effective dielectric constant of the unit cell, which results in a reduction in the out-of-band reflection level.

6.4.2 Simulation and Measurement Results

Fig. 6.10(b) shows the simulated transmission and reflection coefficients of the proposed structure for TE and TM polarizations under normal and oblique incidences. Stable frequency response can be observed. To build the structure, the copper strips with vias are printed on the substrate then cut into long strips with grooves in the mid-way between unit cells, as seen in Fig. 6.11(a). The strips are then interlocked together through the grooves, as illustrated in Fig.

Figure 6.10: (a) Structure of the proposed dual-polarized FSS and (b) Simulated transmission and reflection coefficients for TE and TM polarizations under different incident angles ($b = h = 8$ mm, $d = 1.524$ mm, $w = 1$ mm, $L = 11$ mm, $g_1 = 1.4$ mm, $L_2 = 8$ mm, $g_2 = 1$ mm, $r_{11} = 0.4$ mm, $r_{12} = 0.5$ mm, $r_{21} = r_{22} = 0.3$ mm, $r_h = 0.5$ mm, $\epsilon_r = 3$).

Figure 6.11: (a), (b) Photograph of the fabricated bandstop 3-D FSS, (c) and (d) Simulated and measured transmission and reflection coefficients under TE and TM polarization, respectively.

6.11(b). A prototype of this structure is fabricated using Rogers 4230 substrate ($\epsilon_r = 3$, $d = 1.524$ mm). The fabricated prototype dimension is approximately $180 \text{ mm} \times 180 \text{ mm}$ in size and consists of 21×21 unit cells. A good agreement between measured and simulated transmission and reflection coefficients for $\theta = 0^\circ$ and 40° under TE and TM polarizations is seen from Figs. 6.11(c) and (d).

Table 6.1 Simulated and measured center frequency and fractional bandwidth of the dual-band FSS under normal incidence.

		Simulated	Measured
1 st band	Center frequency	11.2 GHz	11.24 GHz
	-3 dB BW	5.5%	6%
2 nd band	Center frequency	13.8 GHz	13.95 GHz
	-3 dB BW	4%	4.4%

This structure has a measured -3 dB fractional bandwidths of 6% and 3.9% with center frequencies at 11.2 and 13.9 GHz, respectively. Table 6.1 shows a comparison between simulated and measured center frequency and -3 dB fractional bandwidth for each band under the normal incidence. The bandwidth may be widened by using a thicker substrate of lower dielectric constant.

6.5 Ultra-Wideband Dual-Polarized Absorber

6.5.1 The Proposed Absorber

A wideband absorber was presented in [22] and a broadband impedance matching to the air transmission line was achieved using a series R-C network and a resistor in the third dimension to absorb the power in the substrate transmission line (absorption for the first and second modes presented in Figs. 6.2(a) and (b) for the microstrip line case). Polarization sensitivity is the main limitation of this absorber. According to the description provided in Section 6.3, the absorber design presented in [22] can be extended to operate as a polarization-insensitive absorber. The dual-polarized unit cell consists of two symmetric singly polarized unit cells orthogonal to each other. One is responsible to absorb the y -polarized electric field, while the other is responsible to absorb the x -polarized electric field [absorption for the first and third modes presented in Figs. 6.5(a) and (c) for the DSPSL case].

Figure 6.12: (a) Structure of the proposed ultra-wideband dual-polarized absorber. (b) Simulated reflection coefficient of the absorber under normal and oblique incidences for TE and TM polarizations.

By utilizing the absorption of the second mode presented in Fig. 6.5(b), an ultra-wideband dual-polarized absorber can be realized. Fig. 6.12(a) shows the 3-D structure and unit cell details of the proposed absorber. The simulated reflection coefficient under normal and oblique incidences ($\Phi = 0^\circ$ and Θ changes in the x - z plane) for TE and TM polarizations is shown in Fig. 6.12(b) ($b = h = 24$ mm, $L = 20$ mm, $d = 3.048$ mm, $d_1 = 0.1$ mm, $d_2 = 3$ mm, $\epsilon_r = 3$, $w = 1.2$ mm, $w_1 = 1$ mm, $w_2 = 0.8$ mm, $R_a = 330 \Omega$, $R_s = R_h = 200 \Omega$, $C_a = 1.2$ pF). For the incident wave with electric field component polarized along the y -direction (TE polarization), the first and third modes [Figs. 6.5(a) and (c)] are excited in the y -oriented unit cell and can be absorbed using R_a , C_a and R_s as detailed in [22]. Moreover, the second mode [Fig. 6.5(b)] is also excited in the x -oriented unit cell and can be absorbed using y -oriented strips with R_h , which widens the absorption bandwidth of the absorber presented in [22]. A comparison with other designs in the literature is shown in Table 6.2. It can be concluded that our design exhibits the largest absorption bandwidth with a relatively small thickness compared to 2-D single-layer designs and three-layer design employing $\epsilon_r = 2$ and 3 spacers [64]. At the same time, it is even comparable with the three-layer design employing air spacers [64].

The theoretical limit of the minimum thickness for a non-magnetic absorber can be calculated using [110]

$$L_{min} > \frac{\left| \int_{-\infty}^{\infty} \ln|R(\lambda)| d\lambda \right|}{2\pi^2} \quad (2)$$

where L_{min} is the minimum theoretical absorber thickness, R is the reflection coefficient and λ is the wavelength. The infinite integration can be approximated by a finite one [69] from λ_{min} to λ_{max} (the minimum and maximum wavelengths corresponding to the frequency band for a certain reflection level). For normal incidence, it is found that the area under -10 dB curve from 1.51 to 10.1 GHz is equal to 0.25 mm^2 and the minimum theoretical absorber thickness equals to 12.67 mm. By dividing L_{min} by $L_{proposed}$, one can judge how good an absorber design is. This ratio equals to 63.3% for the

Table 6.2 Comparison between the proposed absorber and absorbers available in the literature.

Reference	FBW (%)	Thickness ($\lambda_{\max}$)	Unit Cell Structure
[64]*	98.5	0.112	2-D single-layer resistive patch
	137	0.151	2-D three-layer resistive patch with ϵ_r = 2 and 3 spacers
	158	0.151	2-D three-layer resistive patch with air spacers
[68]	92.3	0.071	2-D single-layer resistive triple- square loops
[108]	112	0.104	2-D single-layer resistive single- square loop
[109]	70.7	0.077	2-D single-layer metal crossed dipoles with resistors
[69]	128	0.088	2-D single-layer metal double-square loops with resistors
[22]	112	0.1	3-D FSS (single-polarized)
This work	148	0.101	3-D FSS (dual-polarized)

* These absorbers were designed to achieve -20 dB reflection coefficient (no fabrication provided).

proposed design. Using the total area under the reflection curve from 0.5 to 11 GHz, it is found that this ratio equals to 86.1%.

6.5.2 Fabricated Prototype

There are two representations for the unit cell of the proposed absorber. One of them is asymmetric and shown in Fig. 6.12(a), where one can clearly see the details inside the unit cell. The other unit cell is symmetric, as shown in Fig. 6.13(a), and used for fabrication and test through the parallel-plate waveguide

Figure 6.13: Assembly details and measurement of the proposed absorber. (a) Symmetric representation of the proposed unit cell shown in Fig. 6.12. (b), (c) Photograph of the fabricated absorber. (d) Photograph of the fabricated absorber within the measurement setup (the upper plate of the parallel-plate waveguide is removed to clearly see the metallic cone and the absorber under test). (e) Comparison between simulated and measured data under the normal incidence.

measurement setup. The unit cell has to be symmetric in order to mimic the periodic structure with a parallel-plate waveguide setup. To build the structure, the copper strips are printed on the substrate and they are then cut into long strips with grooves in the mid-way between unit cells, as shown in Fig. 6.13(b). The strips are then interlocked together through the grooves and backed by the conducting plate. The RC network is soldered as shown in Fig. 6.13(c). To achieve good agreement between the measured data and simulated results, high-frequency resistors and microwave capacitors are used. Ordinary SMD resistors can also be used to reduce the cost of the fabricated structure, but this affects the accuracy of the measured data. The used resistors' and capacitors' size is 0402 package with a part number of 71-FC0402-xxxDTWS series, AVX 04021J1R2PBSTR\500 series, respectively. Fig. 6.13(d) shows the photograph of the fabricated prototype inside the measurement setup. The wideband measurement setup consists of a transition section and a linear taper [69]. The transition from coaxial cable to the parallel-plate waveguide is achieved using a conical metallic disk. The height of the cone is selected to be 10 mm to excite the dominant TEM mode in the parallel-plate waveguide. The purpose of the tapering is to gradually transform the separation from 10 mm at the source to 24 mm at the absorber under test. The resistors and capacitors (C_a and R_a) are printed on the RT5880 substrate ($\epsilon_r = 2.2$, $d = 0.254$ mm), whose effect is already included in the simulated result shown in Fig. 6.12. The fabricated absorber is approximately $144 \text{ mm} \times 24 \text{ mm}$ in size and consists of 6×1 unit cells. Rogers 4230 is used as the substrate material ($\epsilon_r = 3$, $d = 1.524$ mm) for the printed strips. A comparison between measured and simulated reflection coefficient for different BCs along y -axis under the normal incidence is shown in Fig. 6.13(e), and a good agreement is observed.

6.5.3 Practical Considerations

Because the structure is manually assembled and uses lumped capacitors and resistors, some factors can affect the measured reflection coefficient. The most important factors are the change in the air gap between two neighbor unit cells (g_a), air region of the unit cell ($b - d$), and tolerances in d_1 dimension and in

Figure 6.14: Sensitivity studies of the proposed UWB absorber under the normal incidence, (a) Air gap between two neighbor unit cells, (b) Air region of the unit cell, (c) Tolerance in d_1 dimension, (d) Tolerance in capacitor values, (e) Tolerance in resistor values.

capacitor and resistor values. Fig. 6.14 shows the effect of these factors on the absorption performance. Some other simulations are made for various resistor and capacitor values. It is concluded that these factors have a little influence on

the performance of the structure. The main drawback of the proposed absorber might be a large amount of lumped elements used. Regarding the fabrication cost, microwave capacitors are cheap and readily available, while RF resistors are expensive, which may be replaced with ordinary surface mounted device (SMD) resistors for cost reduction at the expense of performance degradation.

6.6 Summary

By employing double-sided parallel-strip lines as a resonator in the unit cell of a 3-D frequency-selective structure, dual-polarized bandstop performance has been achieved. Three quasi-TEM modes are supported in the unit cell. Based on the theoretical explanation provided, equivalent circuit models have been constructed to explain the operating principle of the proposed structures. Two applications have been presented; one is bandstop frequency-selective structure and the other is ultra-wideband absorber. The detailed explanation for the absorber's absorption mechanism has been provided. A prototype has been designed, fabricated and measured to validate the simulated results. The measured data agree well with the simulated results. The proposed design has been fabricated using printed circuit board technology, which highlights the simplicity of the fabrication for such structures. A comparison with other absorber designs in the literature has been performed and it shows that our proposed absorber exhibits the widest absorption bandwidth. Some practical considerations have been discussed to show the insensitivity of the proposed absorber's response to the fabrication tolerance.

As demonstrated in this chapter, the single-polarization problem of the 3-D frequency-selective structure has been physically understood. The next chapter is concerned about a practical application proposed in this thesis, which is the absorptive frequency-selective transmission structure. In the next chapter, a generalized method is proposed to accomplish arbitrarily frequency response within a wide absorption band.

CHAPTER 7

ABSORPTIVE FREQUENCY-SELECTIVE REFLECTION AND TRANSMISSION STRUCTURES

7.1 Introduction

Tailoring methodology is proposed in this chapter to achieve arbitrary frequency response within a wide absorption band. An ultra-wideband absorber utilizing multiple resonances in the unit cell of 3-D frequency-selective structure is utilized as the textile for the proposed designs. Two types are proposed; one is absorptive frequency-selective reflection structure and the other is absorptive frequency-selective transmission structure. These structures have reflection/transmission band within a wide absorption band. Two prototypes are designed, fabricated, and measured to validate the proposed methodology. The first one is an ultra-wideband absorber with 155.7% absorption bandwidth from 1.8 GHz to 14.445 GHz with structure's thickness of $0.12 \lambda_0$ at the lowest absorption frequency. The second one is an absorptive frequency-selective transmission structure with wide absorption bands at both sides of its transmission window. It exhibits 54% and 76% for the lower and upper absorption bands, respectively, with 1 dB insertion loss in the transmission window. A good agreement is achieved between simulated and measured results.

7.2 The Tailoring Methodology

7.2.1 Block Diagram Representation

Fig. 7.1 shows the tailoring methodology illustrated with simple block diagrams and its predicted reflection and transmission coefficients responses. A UWB absorber is constructed at the beginning, which is the basic building block for all the designs. Multiple lossy resonators are employed in the unit cell of 3-D FSS to achieve this goal as illustrated in Fig. 7.1(a) and its design is detailed in Section 7.3. Employing a series resonator in shunt with UWB absorber, a reflection band is inserted within a wide absorption band to create absorption-reflection-absorption frequency response. By tuning the capacitance in this network, tunable AFSR is constructed with the response shown in Fig. 7.1(b). Integrating a transmission path with the UWB absorber, one can achieve absorption-transmission-absorption frequency response shown in Fig. 7.1(c). More details are given in Section 7.2.2 and illustrated with an equivalent circuit model. Introducing a second transmission path, dual-band bandpass response within a UWB absorption is achieved and shown in Fig. 7.1(d).

7.2.2 Circuit Model Representation

Fig. 7.2 shows the equivalent circuit models for the proposed methodology associated with simulated reflection and transmission coefficients using Agilent Advanced Design System (ADS). It may be mentioned that any circuit model for UWB absorber available could be used as the building block for all designs. Here, the circuit model for the absorber presented in [69] is employed. AFSR is constructed by the shunt connection of a UWB absorber with series L-C resonator (bandstop FSS), as shown in Fig. 7.2(a). The reflection band becomes flatter by utilizing a second-order bandstop FSS, which is achieved by cascading double-layer FSS with a dielectric spacer and is modeled as two series L-C resonators with a transmission line in between, as shown in Fig. 7.2(b). By introducing a transmission pass, AFST is constructed. However, this ends with a transmission window with high insertion loss as shown in Fig.

Figure 7.1: Tailoring methodology.

Figure 7.2: Equivalent circuit model of the proposed methodology.

Table 7.1 Equivalent circuit element values of Fig. 7.2.

circuit	Element Values
UWB absorber [69]	$Z_p = 377 \Omega$, $R_1 = 270 \Omega$, $L_1 = 13 \text{ nH}$, $C_1 = 0.222 \text{ pF}$, $R_2 = 280 \Omega$, $L_2 = 8.25 \text{ nH}$, $C_2 = 0.065 \text{ pF}$, $Z_2 = 254.2 \Omega$, $\beta l_2 = 180^\circ$ @ 200 GHz, $Z_1 = 377 \Omega$, $\beta l_1 = 180^\circ$ @ 10.71 GHz.
(a)	$L_s = 29.2 \text{ nH}$, $C_s = 0.034 \text{ pF}$.
(b)	$L_s = 29.2 \text{ nH}$, $C_s = 0.034 \text{ pF}$, $Z = 377 \Omega$, $\beta l = 180^\circ$ @ 10.4 GHz.
(c)	$L_v = 0.01 \text{ nH}$, $Z_s = 10 \Omega$, $\beta_s l = 180^\circ$ @ 10.56 GHz.
(d)	$L_s = 112 \text{ nH}$, $C_s = 0.0081 \text{ pF}$, $L_v = 0.01 \text{ nH}$, $Z_s = 10 \Omega$, $\beta_s l = 180^\circ$ @ 10.78 GHz.
(e)	$L_{s1} = 189.5 \text{ nH}$, $C_{s1} = 0.0087 \text{ pF}$, $L_{v1} = 0.014 \text{ nH}$, $Z_{s1} = 10 \Omega$, $\beta_{s1} l = 180^\circ$ @ 8 GHz, $L_{s2} = 73 \text{ nH}$, $C_{s2} = 0.01 \text{ pF}$, $L_{v2} = 0.009 \text{ nH}$, $Z_{s2} = 10 \Omega$, $\beta_{s2} l = 180^\circ$ @ 12 GHz.

7.2(c). By bypassing the losses associated with the absorber in the passband, the insertion loss is greatly improved. Bypassing is achieved by short-circuiting the absorber using a series L - C resonance network, as shown in Fig. 7.2(d). Utilizing two transmission paths, a dual-band AFST is achieved, as shown in Fig. 7.2(e). The losses in each band are bypassed using two L - C resonators. Table 7.1 summarizes the element values used for the equivalent circuit models of Fig. 7.2.

7.3 Ultra-Wideband Absorber

7.3.1 The structure

By employing a new microstrip line resonator to match and absorb waves at high frequencies, the absorption bandwidth is widened. Fig. 7.3(a) shows the 3-D view with the detailed unit cell dimensions of the proposed absorber employing two microstrip line resonators. The simulated reflection coefficient results under normal and oblique incidences for TE polarization (E in the y -direction, $\varphi = 90^\circ$, $\Phi = 0^\circ$ and Θ in the x - z plane) obtained by CST Microwave

Figure 7.3: (a) 3-D Structure and detailed unit cell of the proposed ultra-wide-band absorber. (b) Simulated reflection coefficient under normal and oblique incidence for TE polarization (E in the y -direction, $\varphi = 90^\circ$, $\Phi = 0^\circ$ and Θ in the x - z plane) ($b = 10$ mm, $h = 11.524$ mm, $L_l = L = 20$ mm, $L_h = 13$ mm, $d = 1.524$ mm, $\epsilon_r = 3$, $w = 2$ mm, $w_1 = 0.4$ mm, $d_l = 0.5$ mm, $d_h = 3.5$ mm, $C_{al} = 0.1$ pF, $R_{al} = 450$ Ω , $R_{sl} = 150$ Ω , $C_{ah} = 0.6$ pF, $R_{ah} = 200$ Ω , $R_{sh} = 150$ Ω).

Table 7.2 Comparison between the proposed absorber and absorbers available in the literature.

Reference	FBW (%)	Thickness ($\lambda_{\max}$)	UC Structure
[64]*	98.5	0.112	2-D single-layer resistive patch
	137	0.151	2-D three-layer resistive patch with $\epsilon_r = 2$ and 3 spacers
	158	0.151	2-D three-layer resistive patch with air spacers
[68]	92.3	0.071	2-D single-layer resistive triple-square loops
[108]	112	0.104	2-D single-layer resistive single-square loop
[109]	70.7	0.077	2-D single-layer metal crossed dipoles with resistors
[69]	128	0.088	2-D single-layer metal double-square loops with resistors
[22]	112	0.1	3-D FSS (single-polarized)
This work	155.7	0.12	3-D FSS (single-polarized)

* These absorbers were designed to achieve -20 dB reflection coefficient (no fabrication provided).

Studio are shown in Fig. 7.3(b). The periodic boundary conditions are used along x - and y -directions of the unit cell to simulate an infinite periodic structure. Under the normal incidence, a fractional bandwidth of 155.7% is achieved from 1.8 GHz to 14.445 GHz with a structure thickness of $0.12 \lambda_0$ at the lowest absorption frequency. The performance of the absorber changes under oblique incidence, especially at the high-frequency band. It is observed that the reflection coefficient in the frequency range 6 - 10 GHz is very stable (approximately constant) under a large range of oblique incidence. It is worthwhile mentioning that it exhibits -10 dB of reflection coefficient with a

fractional bandwidth of 133.3% under the 45° oblique incidence. Table 7.2 shows a comparison between the proposed absorber and other designs in the literature. It is concluded that the proposed design exhibits the largest absorption bandwidth with a relatively small structure thickness.

7.3.2 Experimental Measurement

To build the structure, the copper strips are printed on the substrate and they are then cut into long strips with grooves at their ends, as shown in Fig. 7.4(a). The RO3003 material is used as the substrate material ($\epsilon_r = 3$, $d = 1.524$ mm) for the printed strips. The strips are then interlocked with the supporting pieces through the grooves at their ends and backed by the conducting plate. The front and middle R - C strips are soldered at their locations, as shown in Fig. 7.4(b). The resistors and capacitors (C_{al} , R_{al} , C_{ah} and R_{ah}) are printed on RT5880 material ($\epsilon_r = 2.2$, $d = 0.254$ mm, width = 1 mm). Fig. 7.4(c) shows the photograph of the fabricated prototype in the measurement setup detailed in [69]. The measurement setup consists of two parts, a transition, and a linear taper. The transition from a coaxial cable to the parallel-plate waveguide is achieved using a conical metallic disk. The height of the conical disk is selected to be 10 mm to excite the dominant TEM mode in the parallel-plate waveguide from 1.5 to 14.5 GHz. The purpose of the tapering is to gradually transform the separation from 10 mm at the source to 24 mm at the absorber under test. The fabricated absorber is approximately 144 mm $\times$ 24 mm in size, which consists of 14 $\times$ 2 unit cells.

Along the x -axis, 14 unit cells are used to mimic periodic boundary condition (PBC). However, along the y -axis, the unit cell should be symmetric in order to simulate the PBC with a perfect electric conductor (PEC). Although this unit cell is asymmetric along the y -axis, the simulated results using PBC and PEC BCs along the y -axis are similar, as shown in Fig. 7.4(d). The reason behind this similarity is the fact that this unit cell is shielded along the y -axis. A comparison between the measured and simulated reflection coefficient results under the normal incidence using different BCs along the y and x -axis are shown in Fig. 7.4(d). There are two main reasons behind the difference between

Figure 7.4: (a), (b) Assembly details and the fabricated structure. (c) Fabricated absorber within the measurement setup. (d) Comparison between simulated results with different boundary conditions and measured data under normal incidence.

simulated results and measured data; the first is the assembly tolerance, as will be detailed in Section 7.5, and the second is the use of ordinary SMD resistors. This kind of resistor is used to reduce the cost of the fabricated structure. For better measurement accuracy, high-frequency resistors may be used at the expense of high fabrication cost.

7.4 Absorptive Frequency-Selective Reflection Structure

By short-circuiting the proposed absorber at a certain frequency band, a reflection band is created. Fig. 7.5(a) shows the proposed structure for AFSR utilizing a series resonant circuit to achieve the performance shown in Fig. 7.5(b). The lumped inductor and capacitor (C_s and L_s) value determines the center frequency of the reflection band as well as its fractional bandwidth. By tuning the capacitor value, a tunable AFSR can be constructed. This structure may be useful if it used as a ground plane or a reflector for an antenna to achieve low out-of-band radar cross section (RCS) [111]. The tunable version may be used to construct a tunable antenna system with low RCS. A problem associated with this design is the availability of the capacitor value, which may require some modifications in the design. A flatter reflection band may be achieved by cascading two series resonators in front of the absorber to work as second-order bandstop FSS, as demonstrated in Fig. 7.2(b).

7.5 Absorptive Frequency-Selective Transmission Structure

7.5.1 Initial Design

Fig. 7.6(a) shows the unit cell details of the proposed AFST, which consists of a UWB absorber proposed in Section 7.3 integrated with parallel-plate waveguide with a middle via hole to act as second-order passband path. The simulated reflection and transmission coefficients under the normal incidence for TE polarization (E in the y -direction, $\varphi = 90^\circ$, $\Phi = 0^\circ$ and Θ in the x - z plane) are shown in Fig. 7.6(b) ($b = 10$ mm, $h = 13.048$ mm, $L = 20$ mm, $L_h = 12$ mm,

Figure 7.5: (a) Detailed unit cell of the proposed AFSR. (b) Simulated reflection coefficient under normal incidence for TE polarization (E in the y -direction) with different capacitor values ($b = 10$ mm, $h = 12$ mm, $L = 20$ mm, $L_h = 12$ mm, $d = 1.524$ mm, $\epsilon_r = 3$, $w = 1$ mm, $w_1 = 0.4$ mm, $d_1 = 3$ mm, $C_{al} = 0.1$ pF, $R_{al} = 500 \Omega$, $R_{sl} = 200 \Omega$, $C_{ah} = 0.8$ pF, $R_{ah} = 300 \Omega$, $R_{sh} = 150 \Omega$, $L_s = 5$ nH).

$d = 1.524$ mm, $\epsilon_r = 3$, $w = 1$ mm, $w_1 = 0.4$ mm, $d_1 = 3$ mm, $C_{al} = 0.1$ pF, $R_{al} = 500 \Omega$, $R_{sl} = 200 \Omega$, $C_{ah} = 0.4$ pF, $R_{ah} = 300 \Omega$, $R_{sh} = 150 \Omega$, $d_p = 1.524$ mm, $\epsilon_{rp} = 2.2$, $w_{plate} = 10$ mm, $r = 1.2$ mm). A transmission window with two-sided wide absorption bands is achieved (90% absorption at both sides). The absorption performance of the proposed AFST is shown in Fig. 7.6(c). An absorption rate of 90% is obtained from 2.1 to 4.2 GHz and from 5.5 to 12.3

Figure 7.6: (a) 3-D View and detailed unit cell of the proposed AFST. (b) Simulated reflection and transmission coefficients under normal incidence for TE polarization (E in the y -direction).

GHz, which represent the lower and upper absorption bands, respectively. The main drawback of this design is the high-insertion loss in the passband, which is equal to 2.5 dB. By short-circuiting the source of losses in this band, low insertion loss is achieved and will be discussed in Section 7.5.2.

7.5.2 Improved Design

The unit cell details of the improved design for AFST are shown in Fig. 7.7(a). Fig. 7.7(b) shows the simulated reflection and transmission coefficients of the proposed design shown in Fig. 7.7(a). As shown, low insertion loss is achieved by bypassing the effective resistor using a series resonant circuit. The average insertion loss is improved from 2.7 dB (Fig. 7.6) to 1 dB (Fig. 7.7). By properly choosing the inductor and capacitor values, the resonant frequency, as well as the bandwidth, are adjusted according to the following equations

$$f_{\text{resonant}} = \frac{1}{2\pi\sqrt{L_s \cdot C_s}} \text{ and } BW \propto \frac{C_s}{L_s} \quad (2)$$

A narrow bandwidth is not recommended because the resonant frequency becomes very sensitive to the element value and the transmission window becomes non-flat. A wide bandwidth is also not recommended as it leads to increased out-of-band reflection at the boundaries of the transmission window. The most important factor is the commercial availability of the lumped elements at high frequencies. Especially, the inductor with high self-resonant frequency (SRF) is preferred, which means low parasitic capacitance at the operating frequency band. The required series capacitance is equal to 0.025 pF, but it is not commercially available so that two 0.05 pF capacitors in series are used.

The absorption performance of the improved design of AFST is shown in Fig. 7.7(c). The simulated reflection and transmission coefficients under oblique incidence in the x - z plane for TE polarization are shown in Fig. 7.8(a). A shift in the frequency response is observed, which is due to the change in the electrical length of the passband path with θ . By decreasing the plate's width (w_{plate}) of the passband path, a stable response is achieved and shown in Fig.

Figure 7.7: (a) Detailed unit cell of the proposed AFST with improved insertion loss. (b) Simulated reflection and transmission coefficients under normal incidence for TE polarization (E in the y -direction) ($b = 10$ mm, $h = 12.084$ mm, $L = 20$ mm, $L_h = 12$ mm, $d = 1.524$ mm, $\epsilon_r = 3$, $w = 1$ mm, $w_1 = 0.4$ mm, $d_1 = 3$ mm, $C_{al} = 0.1$ pF, $R_{al} = 700$ Ω , $R_{sl} = 500$ Ω , $C_{ah} = 0.8$ pF, $R_{ah} = 300$ Ω , $R_{sh} = 250$ Ω , $d_p = 1.57$ mm, $\epsilon_{rp} = 2.2$, $w_{plate} = 10$ mm, $r = 1.2$ mm, $L_s = 15$ nH, $C_s = 0.05$ pF).

Figure 7.8: Simulated reflection and transmission coefficients of the structure shown in Fig. 7.7 under normal and oblique incidence for TE polarization (E in the y -direction) when: (a) $w_{\text{plate}} = b$, (b) $w_{\text{plate}} = 9$ mm, $L_s = 15$ nH, $C_s = 0.04$ pF, $R_{at} = 800 \Omega$.

7.8(b). Under a large angle of incidence, the in-band insertion loss increases because of the misalignment between the transmission band and the resonant frequency of the series resonant circuit. The transmission window can be tuned by utilizing different substrate material for passband path. Fig. 7.9 shows the optimized performance of three AFSTs employing three different passband substrates with $\epsilon_r = 1, 2.2$ and 3 . All of them have two-sided wide absorption bands.

An alternative way to enhance the insertion loss of the improved design presented in Fig. 7.7 is to short-circuiting the whole absorber. Fig. 7.10 shows the alternative configuration with the simulated reflection and transmission

Figure 7.9: Simulated reflection and transmission coefficients of the design shown in Fig. 7.7 with different pass-band substrates under normal incidence for TE polarization (E in the y -direction).

coefficients. Although the insertion loss is better than the configuration shown in Fig. 7.7, its component values are not commercially available at microwave frequencies. The first configuration is fabricated, assembled and measured. A comparison with designs in the literature is shown in Table 7.3. It is concluded that the proposed design has the widest lower and upper absorption bands with a small insertion loss and comparable structure thickness at the lowest working frequency.

In order to use the measurement setup shown in Fig. 7.4, the unit cell has to be symmetric along the y -axis. The unit cell, shown in Fig. 7.11(a), is used for fabrication and test. A photograph of the fabricated prototype is shown in Fig. 7.11(b) and (c) is approximately $144 \text{ mm} \times 24.2 \text{ mm}$ in size, which consists of 14×2 unit cells. RO3003 material is used as the substrate material ($\epsilon_r = 3$, $d = 1.524 \text{ mm}$) for the absorber part and RT5880 ($\epsilon_{rp} = 2.2$, $d_p = 1.57 \text{ mm}$) for substrate of the passband part. A comparison between the measured and simulated reflection and transmission coefficients under the normal incidence is shown in Fig. 7.11(d) for different BCs. The variance, although it is not large, between the simulated and measured results is due to the assembly and fabrication errors, which may be considered as a drawback of the proposed structure. The level of the transmitted wave at lower and upper absorption

Figure 7.10: (a) 3-D View of unit cell of the alternative design for AFST with improved insertion loss. (b) Simulated reflection and transmission coefficients under normal incidence for TE polarization (E in the y -direction) ($b = 10$ mm, $h = 12.094$ mm, $L = 20$ mm, $L_h = 12$ mm, $d = 1.524$ mm, $\epsilon_r = 3$, $w = 1$ mm, $w_1 = 0.4$ mm, $d_1 = 3$ mm, $C_{al} = 0.1$ pF, $R_{al} = 600 \Omega$, $R_{sl} = 300 \Omega$, $C_{ah} = 0.8$ pF, $R_{ah} = 200 \Omega$, $R_{sh} = 250 \Omega$, $d_p = 1.57$ mm, $\epsilon_{rp} = 2.2$, $w_{plate} = 10$ mm, $r = 1.2$ mm, $L_s = 13.5$ nH, $C_s = 4.5$ fF).

Table 7.3 Simulated comparison between the proposed AFST and others presented in the literature.

Reference	Lower absorption bandwidth (%)		Upper absorption bandwidth (%)	Insertion loss (dB)	Thickness ($\lambda_{\text{passband}}$)	Thickness ($\lambda_{\text{absorbmax}}$)
[80]	Not exist		70	0.3	0.077	0.172
[73]	Not exist		75	1.9	0.025	0.175
[75]	Not exist		102	0.88	0.033	0.1
[77]	99.1		Not exist	0.5	0.4	0.112
[74]	Strip	25	11.5	0.36	0.165	0.1
	Slot	30	34	0.74	0.2	0.112
[23]	Not exist		85	2.4	0.27	0.3
[24]*	Single-pol.	46	18.8	0.2	0.4	0.17
	Dual-pol.	43.2	15.3	0.5	0.44	0.19
This work	Fig. 7.7	54	76	1	0.31	0.16
	Fig. 7.10	67	71	0.26	0.31	0.13

* These bandwidths are for 80% absorption rate.

bands in the transmission curve is limited to -20 dB due to the limitation of the measurement setup [23]. The component's part number L-07C15NJV6T and 04021J0R1ZBSTR\500 are used for the inductor (L_s) with SRF of 21 GHz and microwave capacitor (C_s), respectively.

7.5.3 Dual-Band AFST Response

Dual-band response can be accomplished by adding one more substrate with a different dielectric constant to act as another passband path. Fig. 7.12 shows the unit cell and simulated reflection and transmission coefficients ($b = 10$ mm, $h = 12.048$ mm, $L = 20$ mm, $L_2 = 12$ mm, $d = 1.524$ mm, $\epsilon_r = 3$, $w = 1$ mm, $w_1 = 0.4$ mm, $d_1 = 3$ mm, $C_{al} = 0.1$ pF, $R_{al} = 600 \Omega$, $R_{sl} = 300 \Omega$, $C_{ah} = 0.6$ pF, $R_{ah} = 200$

Figure 7.11: (a) Unit cell used in fabrication. (b), (c) Assembly details with the fabricated structure. (d) Comparison between simulated results with different boundary conditions and measured data under the normal incidence.

Figure 7.12: (a) 3-D View of unit cell of the proposed dual-band AFST with low insertion loss. (b) Simulated reflection and transmission coefficients under normal incidence for TE polarization (E in the y -direction).

Ω , $R_{sh} = 150 \, \Omega$, $d_{p1} = 1.524 \, \text{mm}$, $\epsilon_{rp1} = 3$, $w_{\text{plate}} = 10 \, \text{mm}$, $r_1 = 1.2 \, \text{mm}$, $L_{s1} = 23 \, \text{nH}$, $C_{s1} = 0.05 \, \text{pF}$, $d_{p2} = 1 \, \text{mm}$, $\epsilon_{rp2} = 1$, $r_2 = 1.4 \, \text{mm}$, $L_{s2} = 20 \, \text{nH}$, $C_{s2} = 9 \, \text{fF}$). As shown, a dual-band bandpass response is achieved concurrently with absorption at low and high sides and in between the two bands. The losses at both bands are bypassed using two L-C networks, each of which resonates at one passband. This may be useful if integrated with a dual-band antenna to achieve dual-band antenna with low out-of-band RCS.

7.6 Practical Considerations

Because the structures are manually assembled and use lumped components, some factors can affect the measured reflection/transmission coefficient. Because the absorber part is common and difficult to be assembled for the proposed structures, sensitivity analysis is conducted for this part. The first factor is the change in the air region of the unit cell ($h - d$). Fig. 7.13(a) shows the effect of this factor on the absorption performance of the structure presented in Fig. 7.3(a). Another important factor is the tilting of the R - C strips in x - y or y - z planes. Fig. 7.13(b) shows this tilting effect in both planes. Some other simulations are made for various component values. It is concluded that these factors have little influence on the performance of both fabricated structures.

Working as an absorber or AFSR/T under single-polarization represents the main limitation of the proposed designs, while not for the tailoring methodology introduced in Section 7.2. This limitation may be solved by employing double-sided-strip line instead of the microstrip line in the unit cell, as presented in Chapter 6, which may be a subject for a future work. Using many lumped components is also considered as another drawback of the proposed designs. The number of components may be decreased by replacing lumped inductor and capacitor with the distributed counterpart as the meander line inductor and the interdigitated capacitor, but the parasitic effect at high frequencies should be considered for the distributed elements.

7.7 Summary

By employing multiple lossy resonators in the unit cell of the bandstop 3-D frequency-selective structure, an ultra-wideband absorber has been constructed. Detailed explanation and equivalent circuit models for an absorber with single-/double-resonator have been presented. Absorptive frequency-selective reflection/transmission structures have been proposed utilizing the ultra-wideband absorber. Two prototypes have been fabricated using printed circuit board technology to highlight the simplicity of the fabrication for such kind of structures. Dual-band frequency-selective transmission has been accomplished

(a)

(b)

Figure 7.13: Sensitivity studies of the proposed UWB absorber under normal incidence, (a) The air region of unit cell, (b) Tilting effect of R-C strips.

by integrating two transmission paths with an ultra-wideband absorber. The losses have been bypassed by utilizing two different series resonant circuits, each of which works at one band. Two comparisons have been performed between the proposed designs and others in the literature. Because the fabricated structures have been manually assembled, some practical

considerations have been discussed. Some future work may be considered regarding dual-polarization performance and integration with antennas to achieve out-of-band RCS reduction.

CHAPTER 8

CONCLUSIONS AND RECOMMENDATIONS

8.1 Conclusions

This thesis focuses on overcoming existing limitations and introduce several applications of the 3-D frequency-selective structure. A comprehensive literature review has been conducted on 2-D frequency-selective surfaces, 3-D frequency-selective structures, absorbers, and absorptive frequency-selective transmission and reflection structures in Chapter 2. It has been found that it is difficult to obtain bandstop performance with multiband and high-order response using 2-D frequency-selective surfaces, and most of the existing designs exhibit bandpass response. In addition, although 3-D frequency-selective structure has a sharp response with a stable performance under oblique incidence, it suffers from two main problems. It is a thick structure and mostly operates for a single-polarized wave. Moreover, most of the absorptive frequency-selective transmission structures that were presented in the literature have one-sided absorption band beside the transmission band.

In Chapter 3, two structures have been presented to demonstrate the general methodology proposed for realizing multiband high-order bandstop frequency-selective structures. The first structure has been designed to exhibit a tri-band second-order response and the second has been designed to achieve a dual-band third-order response. Even-odd mode analysis has been discussed for the coupling between two identical resonators that consequently control the separation between the two in-band transmission zeros and equivalent circuit model has been constructed. Measured results of these two designs show a very good agreement with simulated ones and stable filtering performance under various angles of incidence. It is expected that the same methodology may also apply to bandpass frequency-selective structure.

Chapter 4 has introduced a solution for the first drawback of 3-D frequency-selective structure, and an 80% reduction in the 3-D bandstop frequency-selective structure thickness has been achieved by employing the loop resonator instead of microstrip line resonator in the unit cell. A comparison between the microstrip line resonator and our proposed resonator shows comparable performance in terms of the bandwidth and selectivity sharpness. Three structures have been designed and fabricated to verify the advantages of the proposed resonator. The first two structures operate only under TE polarization, while the third structure operates under both TE and TM polarizations. Stable frequency response has been achieved for all structures for more than 60° oblique incidence.

A thin bandpass 3-D frequency-selective structure has been presented in Chapter 5 based on a folded substrate. It has been demonstrated that a 79% thickness reduction is possible by employing a folded structure compared to the basic structure. By increasing the folding layers, more thickness reduction can be achieved. A fabricated prototype has been measured in a parallel-plate measurement setup and measured results show good agreement with the simulated ones. Dual-polarized and dual-band versions have been introduced and they demonstrate stable performance under oblique incidence. A half-cylindrically shaped radome with filtering properties has been integrated with a broadband horn antenna. The effect of the beamwidth of the horn antenna has been investigated. A fabricated prototype has been tested to validate the simulations.

By employing double-sided parallel-strip lines as the resonator in the unit cell of a 3-D frequency-selective structure, dual-polarized bandstop performance has been achieved in Chapter 6. Three quasi-TEM modes are supported in the unit cell. Based on the theoretical explanation provided, equivalent circuit models have been constructed to explain the operating principle of the proposed structures. Two applications have been presented; one is a bandstop frequency-selective structure and the other is an ultra-wideband absorber. By utilizing the general methodology introduced in Chapter 3, a dual-band second-order and

dual-polarized bandstop 3-D frequency-selective structure has been designed, fabricated and measured. The detailed explanation for the absorber's absorption mechanism has been provided. An absorber prototype has been designed, fabricated and measured to validate the simulated results. The measured data agree well with the simulated results. The proposed design has been fabricated using the printed circuit board technology, which highlights the simplicity of the fabrication for such structures. A comparison with other absorber designs in the literature has been performed and it shows that our proposed absorber exhibits the widest absorption bandwidth. Some practical considerations have been made to show the insensitivity of the proposed absorber's response to the fabrication tolerance.

In Chapter 7, a tailoring methodology has been proposed to achieve arbitrary frequency response within a wide absorption band. By employing multiple lossy resonators in a unit cell of the bandstop 3-D frequency-selective structure, an ultra-wideband absorber has been constructed. Absorptive frequency-selective reflection/transmission structures have been proposed utilizing the ultra-wideband absorber. Two prototypes have been fabricated using printed circuit board technology to highlight the simplicity of the fabrication for such kind of structures. Dual-band frequency-selective transmission has been accomplished by integrating two transmission paths with an ultra-wideband absorber. The losses have been bypassed by utilizing two different series resonant circuits, each of which works at one band. Comparisons have been performed between the proposed designs and others in the literature. Because the fabricated structures have been manually assembled, some practical considerations have been discussed. Some future work may be considered regarding dual-polarization performance and integration with antennas to achieve out-of-band RCS reduction.

In summary, the problem of designing bandstop frequency-selective surface with the multiband and high-order response has been solved utilizing 3-D frequency-selective structure. Moreover, the thickness problem associated with the 3-D bandstop frequency-selective structures has also been addressed by

employing the loop resonator instead of the traditional microstrip line resonator in the unit cell. Furthermore, the dual-polarized performance has been achieved by employing double-sided parallel-strip lines as a resonator in the unit cell of a 3-D frequency-selective structure. By investigating the modes supported by this unit cell, it has been found that this unit cell can support three quasi-TEM modes. A thin and ultra-wideband dual-polarized absorber has been designed by properly absorbing these modes. Finally, a tailoring methodology has been introduced in order to achieve an arbitrary frequency response within an ultra-wide absorption band.

8.2 Recommendations for Future Work

This thesis has presented several designs with novel concepts and improved performance. However, it still offers many opportunities for us to have a further investigation. Some of the suggested studies are listed as follows.

1. In Chapter 3, a generalized methodology has been proposed to achieve multiband high-order bandstop 3-D frequency-selective structure. By cascading these structures with an ultra-wideband absorber, multiband absorptive frequency selective reflection structure can be achieved. This new structure can be employed as a ground plane for a multiband antenna to achieve a multiband antenna system with low out-of-band radar cross-section. Fig. 8.1(a) shows the 3-D view of the unit cell for dual-band bandstop 3-D AFSR structure. The simulated reflection coefficients for this dual-band AFSR structure under the normal and oblique incidences for TE- and TM-polarized waves are shown in Fig. 8.1(b).
2. In Chapter 7, absorptive frequency-selective transmission structure has been presented. This structure can be integrated with an antenna as a superstrate to achieve low radar cross-section antenna system. Moreover, a tunable version of this structure may be useful to achieve a tunable antenna system with low radar cross-section. A simulation model is shown in Fig. 8.2. The unit cell consists of two main parts. The first part is an ultra-wideband absorber constructed from two lossy resonators, as

Figure 8.1: (a) 3-D unit cell view of the proposed dual-band AFSR structure. (b) Simulated reflection coefficient under normal and oblique incidence for TE and TM polarizations ($L_{\text{stop}} = 20$ mm, $L_{\text{stop}h} = 14.7$ mm, $b = 12$ mm, $d = 1.57$ mm, $\epsilon_r = 2.2$, $w = 1.5$ mm, $g_l = g_h = 0.7$ mm, $D_v = 1$ mm, $D_{\text{airhole}} = 1.5$ mm).

Figure 8.2: (a) 3-D structure view, (b) detailed unit cell of the proposed tunable AFST structure.

shown in the figure. The second part is the passband substrate to provide the transmission window within the absorption band. Two varactor diodes with a part number of MAVR-011020-1411 are utilized to achieve tunable capacitor $C_{passtune}$ to tune the transmission window and C_{stune} to tune the series resonator to bypass the losses. The detailed biasing network is also shown in the figure for both varactors.

3. The structures proposed in Chapter 7 suffer from two main limitations:
 - (i) operation under single-polarization, which may be solved by utilizing

double-sided-strip line instead of microstrip line in the unit cell, as presented in Chapter 6. and (ii) using many lumped elements, which may be decreased by replacing lumped inductor and capacitor with the distributed counterpart as meander line inductor and interdigitated capacitor. These two limitations may be addressed in the future.

AUTHOR'S PUBLICATIONS

Journal papers

1. **A. A. Omar** and Z. Shen, "Multiband high-order bandstop 3-D frequency-selective structures," *IEEE Trans. Antennas Propag.*, vol. 64, no. 6, pp. 2217–2226, Jun. 2016.
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4. **A. A. Omar**, Z. Shen, and H. Huang, "Absorptive frequency-selective reflection and transmission structures," *IEEE Trans. Antennas Propag.*, vol. 65, no. 11, pp. 6173–6178, Nov. 2017.
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1. **A. A. Omar** and Z. Shen, “Multi-band second-order bandstop frequency selective structure with controllable band ratios,” *IEEE AP-S Antennas and Propag. Symp.*, Vancouver, British Columbia, Canada, July 19-24, 2015.
2. **A. A. Omar** and Z. Shen, “A loop resonator for 3-D frequency selective structure,” *International Symp. On Antennas and Propag. (ISAP)*, Hobart, Tasmania, Australia, November 09-12, 2015.
3. **A. A. Omar** and Z. Shen, “Dual-polarized absorber based on 3-D frequency-selective structure,” *IEEE AP-S Antennas and Propag. Symp.*, Fajardo, Puerto Rico, USA, June 26- July 01, 2016, pp. 675–676.
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5. **A. A. Omar** and Z. Shen, “Ultra-wideband harmonic-suppressed bandpass/stop frequency-selective surface based on L - C lumped elements,” *12th European Conference on Antennas and Propagation (EUCAP)*, London, UK, April 09-13, 2018.

6. **A. A. Omar** and Z. Shen, “Tunable absorptive frequency-selective transmission structure,” *IEEE AP-S Antennas and Propag. Symp.*, Boston, MA, USA, July 08-13, 2018.

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