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Abstract

This dissertation provides a synthesis technique for the design of thickness-customizable high-order ($N \ge 2$) bandpass frequency selective surface (FSS) and its application in realizing versatile multi-layered FSS and absorbers. Admittance inverters layers are used to synthesize the transfer response of the filter given desired characteristics such as filter type, center frequency, and bandwidth. These inverter layers are essentially electromagnetic coupling interlayers that can be adjusted to customize the thickness of multilayered FSS without degrading the desired filter performance. A generalized equivalent circuit model is used to provide physical insights of the proposed design. This synthesis technique is adopted to deliver a versatile implementation capability of high-order FSS filters using various dielectric spacers with arbitrary thicknesses. Such technique enables the realization of spatial filters with variable size, while maintaining the desired filter response. To highlight the significance of the proposed synthesis technique, its concept is applied to two practical problems including the design of compact switchable FSS and adaptive/tunable microwave absorbers as it may allow simpler integration of active components that require specific physical dimensions.

In the first practical problem, the feasibility of deploying plasma switchable compact spatial filter in harsh electromagnetic radiation environments is investigated. The proposed FSS integrates contained plasma (plasma-shells) as active tuning elements. These ceramic, gas-encapsulating shells are ideal for high-power microwave and electromagnetic pulse protection because they are rugged, hermetic, operable at extreme temperatures, and insensitive to ionizing radiation. A 2D periodic second-order switchable spatial filter is implemented. It is composed of electrically small Jerusalem cross structures embedded with discrete plasma shells strategically located to effectively switch the transfer function of the filter. This technique is used to realize compact low profile second order band pass spatial filter operating at S-band. It also has the ability to switch its transfer function within 20 to 100 ns while enabling in-band shielding protection for aerospace or terrestrial electromagnetic systems subjected to high power microwave energy (HPME) and high electromagnetic pulse (HEMP) in harsh space environment. Experimental results are shown to be in good agreement with simulation results.

The second practical problem is addressed by deploying a large-scale adaptable compressed Jaumann absorber for harsh and dynamic electromagnetic environments. The multilayered conductor-backed absorbers are realized by integrating ceramic gasencapsulating shells and a closely coupled resonant layer that also serves as a biasing electrode to sustain the plasma. These active frequency-selective absorbers are analyzed using a transmission line approach to provide the working principle and its frequency tuning capability. By varying the voltage of the sustainer, the plasma can be modeled as a lossy, variable, frequency-power-dependent inductor, providing a dynamic tuning response of the absorption spectral band. To study the power handling capability of the tunable absorber, dielectric and air breakdowns within the device are numerically emulated using electromagnetic simulation by quantifying the maximum field enhancement factor (MFEF). Furthermore, a comprehensive thermal analysis using a simulation method that couples electromagnetics and heat transfer is performed for the absorber under high power continuous microwave excitations. The maximum power level handling capability of the microwave absorber has been numerically predicted and validated experimentally.

SYNTHESIS TECHNIQUE OF THICKNESS-CUSTOMIZABLE MULTILAYERED FREQUENCY SELECTIVE SURFACE FOR PLASMA-BASED ELECTROMAGNETIC STRUCTURES

by

Komlan Payne

B.S.E.E., Syracuse University, 2014 M.S.E.E., Syracuse University, 2019

Dissertation

Submitted in partial fulfillment of the requirements for the degree of Doctor of Philosophy in Electrical and Computer Engineering.

Syracuse University May 2020 Copyright © Komlan Payne 2020 All Rights Reserved To my wonderful parents, Assiba and Antoine To my lovely wife, Jillianne

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Chapter 1

Introduction

1.1 Overview of Frequency-Selective Surfaces

Frequency Selective Surfaces (FSSs) are periodic structures arranged in a one or two dimensional lattice, designed to filter electromagnetic (EM) spectrum ranging from microwave to terahertz technology [2, 4]. They are constructed of arbitrary conductive geometry using single or stacked layer backed by dielectric material and designed to reflect, transmit or absorb EM wave based on frequency, angle of arrival and polarization. Consequently, depending on their physical construction, material and geometry, they can be categorized into low-pass, high-pass, band-pass and band-stop spatial filters.

These artificial surfaces are widely used in defense and wireless applications such as in designing radomes [5] and dichroic surfaces for reflectors and sub-reflectors [6]. In fact, early work concentrated on the design of FSS as hybrid radomes mounted at the front of an aircraft or terrestrial military vehicle. In such case, FSSs are used as a front-end filter to control EM waves before they couple with the antennas or sensors in the system. By doing so, the FSS layers protect the enclosed antennas from the outside environment using a bandpass FSS that is transparent within the operation frequencies of the antennas and opaque at other frequencies [7–9]. The concept has also begun to transfer to commercial sectors for cyber security purposes and the design of novel electromagnetics devices such as microwave lenses [10, 11], polarizers [12, 13] and remote sensors [14]. For example, FSS-based wallpapers have been recently researched to contain private Wi-Fi signals within the desired residential/industrial buildings while permitting radio and emergency waves to pass through [15, 16]. Unlike a microwave guided filter where the electromagnetic waves propagate through transmission lines or waveguides, FSSs are used to filter free-space



Figure 1.1. FSS building blocks composed of patch and wire grid. A capacitance is formed between the adjacent edges of two coplanar patches while inductance is associated with a metallic strip for a TEM wave as illustrated in the picture. Figure reproduced from [1].

(unguided) propagating EM wave. In its simplest form, the constituent periodic elements of FSS can be divided in two metallic building blocks (wire grid or patch type) as shown in Figure 1.1. A capacitance is formed between the adjacent edges of two coplanar patches for a transverse electromagnetic (TEM) wave with the electric field perpendicular to the gap while inductance is associated with a metallic strip for a TEM wave in which the electric field is parallel to the wire. The value of the capacitance and inductance is related to the geometrical parameter of the metallic patch and wire grid using approximated analytical solution given in [17] as:

$$C = \frac{2a}{\pi} \varepsilon_0 \varepsilon_{reff} ln\left(csc(\frac{\pi s}{2a})\right); \ L = \frac{2a}{\pi} \mu_0 \mu_{reff} ln\left(csc(\frac{\pi w}{2a})\right), \tag{1.1}$$



Figure 1.2. Common FSS elements unit cell (a) dipole; (b) cross dipole; (c) Jerusalem cross; (d) four-legged loaded element; (e) tripole; (f) square loop. Figure reproduced from [2]



Figure 1.3. Common resonator (bandpass) elements unit cell (a) slot dipole; (b) cross dipole slot; (c) Complementary Jerusalem cross; (d) Complementary four-legged loaded element; (e) tripole slot; (f) square loop slot. Figure reproduced from [2]

where ε_0 , μ_0 are the free space permittivity and permeability, ε_{reff} , μ_{reff} are the effective permittivity and permeability of the substrate, a is the periodicity of the FSS, s is the separation between two adjacent patches, and w is the width of the wire grid. A variety of FSS elements were introduced for bandpass and bandstop applications using combination of the two building blocks. A complete list of bandstop elements is shown in Figure 1.2. Their complementary topologies constitute a bandpass counterpart as shown in Figure 1.3.

1.2 Background

Traditional FSSs, at resonance frequency, have their constituent elements with effective length that are multiple of half the wavelength $\lambda/2$. The dependence of the FSS frequency response with respect to these factors could be a major drawback for some applications.

For example, the size of each element becomes problematics as it enables early grating lobe (harmonics) of the intended frequency which themselves depend upon the incidence angle. In general the desired properties for FSS are as follows:

- Sub-wavelength unit lattice dimensions (periodicity *« λ*) provide a more stable filtering response to the wave incident from oblique angles and avoid early onset of the grating lobes.
- Multi-pole (multi-layer) FSSs are preferred in applications requiring broadband operation with sharper band skirts and higher out-of-band rejections.
- Controllable spacing between the layers: thinner spacing between the layers help stabilize the filtering responses of multi-layered FSSs for waves impinging from oblique angles. Low profile design also adds benefits in improving overall aerodynamics, conformability, and weight loss. However, flexible design thickness is more desired to applications that demand requirements on the physical features of the constituent subcomponents.
- Polarization insensitive FSSs minimize polarization mismatch loss and help reduce the total insertion loss of the filter. This feature is also best suited for randomly moving and/or rotating systems.
- Simple and low cost realization solutions such as scalable, planar or conformal topologies without grounding vias help improve the versatility of the spatial filters while relaxing fabrication requirements. FSSs based on non-exotic materials can further reduce overall fabrication complexity and cost.

However, simultaneously optimizing all above mentioned properties has been a challenging engineering task. Furthermore, one of the more challenging design requirements is to simultaneously obtain multipole FSS response in a low-profile design.

1.3 Motivation

FSSs are typically designed using classical filter synthesis techniques [2]. In general, steepness of a filter response curve is controlled based on the filter order. Thus, the higher order filter responses are preferred for practical application demanding flat in-band top and sharper band skirt for higher out-of-band suppression. In analogy of the design of high-order guided microwave filter using inverters, multipole FSS design can also be implemented using spatial inverters. However, the use of quarter-wavelength separations to implement inverters between adjacent resonant FSS layers [2] not only results in a relatively thick structure but also often deteriorates the filter performance, in particular for waves impinging from oblique angles. An Nth-order filter using this approach has an overall thickness in the order of $(N-1)\lambda_g/4$. To address this issue, a closely coupled three layers FSS arranged as resonator-aperture-resonator configuration [18, 19] has demonstrated stable second-order filter response for broad range of incident angle in a compact form. A miniaturized multipole spatial filter is also realized by alternating non-resonant patches and wire grid layers [1, 20, 21]. These compact designs also provide stable filtering performance for waves impinging from oblique angles and can achieve the overall thickness of about $(N-1)\lambda_0/30$. Although such design method may be ideal in realizing low-profile and/or conformal FSSs applications, it may not lend much freedom in adjusting the separation between each metallic layer. A general design method that enables thickness control of multipole (multilayered) FSSs is yet to be explored. Thus, this thesis provides a new multilayered FSS design technique that can be applied for different thickness FSSs, while maintaining the desired filter responses. We envision the presented synthesis technique would bring a practical alternative design solution for future FSS applications requiring precise thickness specifications.

1.4 Concept of the Thickness-Customizable Multilayer Bandpass FSS



Figure 1.4. Topology of three-pole FSS bandpass filter using coupling interlayers.

Recently, we have introduced the feasibility of realizing miniaturized three-pole bandpass FSS based on inverter interlayers. The design was realized by sandwiching coupling interlayers between three bandpass FSS layers comprised periodic miniaturized complementary Jerusalem cross structures as shown in Figure 1.4. The total thickness of the design can be adjusted between $\lambda_0/4$ and $\lambda_0/10$. Although the initial studies served well in indicating the possibility of transforming the concept to design thickness variable multilayered FSS, a comprehensive and systematic design guide to simultaneously control filter characteristics and overall filter thickness has not been presented. The filter responses to the incident waves were obtained through brute-force optimization-based techniques using High-Frequency Structure Simulator (HFSS) from the ANSYS Corporation. Also, the filters deliver maximally flat bandpass response for various FSS thicknesses, but in doing

so, the filter characteristics including operation frequency and fractional bandwidth are unintentionally altered (see Figure 1.5). This was due to the lack of a synthesis procedure that can be employed to compensate the detuning of desired filter characteristics.



Figure 1.5. Three-pole bandpass FSS with coupling interlayer that deliver highly selective bandpass response with (a) total thickness $d = 7.7mm \approx \lambda_0/4$; (b) total thickness $d = 3.175mm \approx \lambda_0/10$.

In Chapter 2, a generalized technique for synthesizing bandpass FSS filters of arbitrary order ($N \ge 2$) that allows simultaneous control of the thickness and transfer response for multilayered FSS filters is propesed. Our approach provides the capability to control the thickness to arbitrary dimensions that might be needed for a particular application. The use of equivalent circuit model (ECM) provides the physical insight behind the simultaneous control of the filter response and thickness. Thickness customizability allows realization of multilayered FSS filters with desired dimensions in order to flush mount FSS filters onto arbitrary surrounding structures, or to allow the use of readily available dielectric substrates. Commercial dielectric slabs (e.g., from Rogers Corp.) are typically available in limited thickness options (e.g., h = 0.127, 0.254, 0.381, 0.508, 0.76, 1.524, 1.27, 1.9, 2.5, and 3.175 mm). In order to validate the design's versatile implementation capability, a high order bandpass FSS is synthesized using the proposed technique and implemented in four different predefined thicknesses while maintaining the same desired filter characteristics.

1.5 Potential Applications

1.5.1 Plasma-Enabled Compact Switchable FSS for the Protection of Electromagnetic Systems against High Power Microwave

In modern communication systems, high power microwave (HPM) energy constitutes a serious threat for the survival of the electronic and navigation system including satellite, guided missile, manned or unmanned aircraft, ground vehicle, submarines, etc. Intensive research toward the feasibility of electromagnetic shielding surface to protect electronic systems in harsh EM/RF environment in order to provide secure space sensors has been the focus in electronic warfare. FSSs can be used as spatial microwave limiters to control EM waves before they couple with the antennas or sensors in many systems [8]. But a passive bandpass FSS can only protect the system from out-of-band high intensity field because they behave as a transparent medium within the operation spectrum. As a result, jamming attacks, nuclear blasts or high power microwave weapon within the operation band constitute a potential threat for the survival of these systems. Reconfigurable FSSs are therefore required to protect both in-band and out-of-band against this vulnerability. These types of FSSs have been previously investigated using RF MEMS switch, semiconductor varactors, PIN diode, liquid crystal polymers, etc. [9, 22]. However, the use of existing reconfigurable technologies requires trade-offs in terms of design tuning range/speed, cost, reliability, large scale integration, power handling/consumption, and linearity.

In Chapter 3, a switchable electronic protection structure based on plasma-shell technology is implemented. The applicability of voltage controlled plasma as a tuning component embedded in FSS renders this technology achievable. The behavior of the electric field intensity within the operating band of a switchable band pass spatial filter based on plasma is illustrated in Figure 1.6 for both OFF and ON states. The electric field is transparent to



Figure 1.6. (a) Switchable FSS with layers separated by quarter-wavelength. (b) Compact Switchable FSS design using coupling inter-layers.

the spatial filter when the plasma is OFF state. When the plasma is turned ON, transition from lossless dielectric shell to a lossy conductive plasma is rapidly achieved and produces strong shielding effectiveness for the transmission of the electric field. Based on this concept, we demonstrate a switchable second order FSS based on plasma technology. Two topologies (as shown in Figure 1.6) are adopted to provide a fast switchable passband response with both in-band and out-of-band protection. In the first topology (Figure 1.6(a)), a second order bandpass FSS is realized by cascading two single pole FSS layers about quarter wavelength apart from each other. By introducing inverter inter-layer between two resonators, the design thickness can be considerably reduced. For the second topology, this technique is used to design an alternative switchable FSS in a much compact manner (Figure 1.6(b)). The specification and performance of the proposed designs are detailed in Chapter 3.

1.5.2 Plasma-Enabled Compact Adaptive/Tunable Absorber for the Protection of Electromagnetic Systems against High-Power Microwave

The need for tunable, low-profile, light-weight, and high-power handling absorbers has increased significantly as aerospace applications demand reliable radio-frequency/microwave radar cross section (RCS) reduction devices that can rapidly adapt to harsh and dynamic electromagnetic environment while enabling a more agile flight maneuverability [23, 24]. These camouflage systems, used particularly for stealth applications, have the capability to alter the RCS of a target [25, 26]. Among many, conductor-backed absorbers are used extensively in military and aerospace applications to reduce or deflect the backscattered EM signal from their defense equipment (manned or unmanned aircrafts, ground vehicles, naval vessels, guided missiles, etc.) that can be detected by hostile radar [26]. Such absorbers comprised of various materials, shapes, sizes, and design patterns are intended to dissipate the incident wave at the spectrum of interest while reflecting the out-of-band frequency waves. Absorbers can be broadly categorized into material-based and structuralbased. The most well-known material-based absorber is the pyramidal absorber, which presents a gradual transition in impedance from air to absorber. However, it can only be used in limited applications, such as in anechoic chambers because the material used is fragile and bulky.

On the other hand, structural-based absorbers use resonant behavior to absorb the EM wave. The Salisbury screen [27], which uses a lossy homogenous sheet backed by a completely reflective plane at quarter-wavelength distance at the operating center fre-

quency, is one of the most basic resonant type absorbers. This single layer design has limited bandwidth due to its resonant behavior. To improve the absorption bandwidth, the Jaumann absorber [28] uses multiple resistive layers cascaded quarter-wavelengths apart. However, this technique is impractical in most applications due to its large design thickness. Also, the sheet resistance values of the layers increase exponentially from the ground outward to the free space, which leads to manufacturing limitations. Another type of multilayers absorber based on the quarter wavelength resonator concept is named after Dallenbach [29]. Dallenbach absorber depends on the permittivity and permeability of lossy substrates cascaded quarter-wavelengths apart. With the evolution of FSS designs, new radar absorbers including circuit analog absorber [2, 30] and metamaterial absorbers [31–34] have been introduced. Circuit analog absorbers use cascaded lossy bandstop FSS sheets about quarter-wavelength between layers, similar to the Jaumann absorber, but with better performance. In contrast to all aforementioned absorbers that are limited to quarter-wavelength spacers between adjacent layers, metamaterial absorbers based on high impedance surfaces offer designs with flexible thickness from near contact to quarter-wavelength. These periodic structures provide artificial boundaries to the electromagnetic waves and can be used to implement either very thin narrow-band absorbers or thick wideband absorbers. Nevertheless, the performance of passive absorbers with fixed reflectivity-bandwidth characteristics cannot withstand modern frequency agile radar (FAR) systems.

In Chapter 4, we investigate the feasibility of devising electronically tunable compressed Jaumann absorbers based on plasma electrical properties to address critical limitations encountered in harsh and dynamic electromagnetic environments. The technical approach relies on applying thickness controlling inter-coupling layers and integrating tiny plasmashells to enhance the performance and tunability of the absorber (as shown in Figure 1.7). This Inter-coupling layer concept has recently proven to be useful in realizing low-profile



Figure 1.7. Topology of the plasma-based tunable compact adaptive/tunable absorber using coupling interlayer

multi-layered FSS bandpass filter and has also demonstrated dual-purpose capability by serving as a biasing network for the active components. Encapsulated-plasmas are unique among all competing technologies because they provide a reliable control element to the devices operating under electromagnetically harsh conditions. Due to the vulnerability of microwave absorber when exposed to high power microwave/electromagnetic pulse (HPM/EMP), the power handling capability of the tunable plasma-based multilayer absorber is studied using multiphysics analysis, then validated by experimental data. Fabricated prototype absorbers are measured using standard low power and relatively high power measurement setup in free space. Then, we suggest design options that are more resilient to thermal stress to help maintain a more stable operation when applied for systems that are subjected to HPM.

Chapter 2

Synthesized Technique for the Design of Thickness Customizable High-Order Bandpass FSSs

2.1 Chapter Introduction

The revolutionary concept of engineering structures to emulate the atomic lattice with controllable electrical properties was first proposed by J. B. Pendry et al [35, 36] in the late nineties. After successful experimental demonstration of negative index of refraction phenomena using metamaterials (materials engineered to have properties that have not yet been found in nature) by R. A. Shelby et al [37], various novel microwave and optical devices have been realized. In recent years, the extraordinary properties of metasurfaces (two-dimensional equivalent of metamaterial) have stimulated the interest of research community. Metasurface broadly refers to two dimensional artificial structures that enable extreme control of electromagnetic fields [4]. These synthetic surfaces are realized by arranging subwavelength metallic and/or dielectric structures in an array fashion. However, the modern microwave and RF systems demand ever more stringent design requirements on the physical features of the constituent subcomponents. For example, flexible implementation solutions that can deliver desired element size and weight present important design challenges. In this Chapter, advanced design techniques is devised to address theses issues.

2.2 Filter Theory Concept and Configuration of the Proposed Multipole FSS

Coupled filter theory has been widely applied and used to design waveguide filters, dielectric resonator filters, ceramic combline filters, and microstrip filters [38–42], and has served as an extremely powerful tool in both synthesizing and optimizing guided filters [43, 44]. As for traditional guided filters, inverters enable various alternative implementation options to achieve the same desired transfer responses. Recently, coupled-resonator filter theory based on impedance inverter [45] was used to synthesize multipole FSS composed of dielectric slab and wire grid array. However, in this design, the dielectric spacers are used as resonators with half-wavelength electric thickness. Not only does the design implementation prevent overall thickness customizability but also will result in relatively thick structure at lower operating frequency, which often deteriorates the filter performance. Such performance degradation becomes more pronounced when the waves impinge the FSS from oblique angles. In order to compress the total thickness of the multipole FSS filter, coupling interlayers can be integrated with subwavelength or half-wavelength [18, 46] resonator-based periodic structures. Coupling interlayers are essentially inverter layers. When applied to FSS filters, this concept allows one to replace quarter-wavelength spacers (a type of inverter) to a more compact or even thickness adjustable artificial inverter layers. A maximally flat bandpass response can be realized by placing single-pole FSS layers quarter-wavelength apart [2]. However, if one tries to reduce the separation between the FSS layers, the filter performance degrades due to increased coupling strength between the resonant layers. FSS based on aperture coupling interlayers is a powerful method that allows control of the coupling strength, thus enabling the design of selective filtering response for multilayered FSS filters with desirable separation between

successive layers.



Figure 2.1. Unit cell representation of the topology of the Nth-order spatial bandpass filter based on aperture coupling (inverter) interlayers.

The proposed multilayered bandpass spatial filter is shown in Figure 2.1, where the design is symmetrical about the transverse xy plane located at the middle of the structure. An Nth-order multilayered FSS is comprised *N* metallic bandpass resonators layers, (N - 1) aperture coupling interlayers and 2*N* dielectric slabs. For the purpose of demonstration, we selected a relatively simple square loop bandpass resonator and fixed the dielectric constant (ε_r) to the same values in all the dielectric slabs, with relative permeability $\mu_r = 1$. In this example, the magnetic coupling is dominant since the center of the aperture is located away from the square loop slot (where the electric fields are predominant) [47]. Thus, the nonresonant aperture layers placed between pairs of resonators behave as inductively coupled interlayers. For this implementation, the major magnetic coupling contribution comes between adjacent resonant layers linked by direct coupling. Cross-coupling and mixed coupling may be useful in designing more complex multipole FSS configurations to improve its performance as it was demonstrated for guided filter to introduce additional transmission zeros (TZs) and reduce insertion loss [48–50]. However,
adopting this technique can unveil a much more complicated synthesis level for the design of multipole FSS as the model will become more cumbersome. This was the case in our previous approach design where both electric and magnetic coupling coexist.



Figure 2.2. ECM of the Nth-order spatial bandpass filter consisting of shunt parallel resonators and shunt inductors separated by short transmission lines.

2.3 Equivalent Circuit Model and General Design Guide

The equivalent circuit model (ECM) of the Nth-order multipole FSS in Figure 2.1 is illustrated in Figure 2.2. Individual bandpass FSS layers are modeled as parallel $L_{pi}^{"}$, $C_{pi}^{"}$ ($1 \le i \le N$) resonant tanks and the aperture layers as shunt inductors $L_{i,i+1}^{r}$ ($1 \le i \le N - 1$). Each substrate is modeled by a short transmission line with characteristic impedance $Z_d = Z_0 \times \sqrt{\mu_r / \varepsilon_r}$, and length $h_{[i,i+1]L/R}$. The lengths $h_{[i,i+1]L}$ and $h_{[i,i+1]R}$ represent the thickness of the dielectric slabs sandwiched between the *i*th and the (i + 1)th bandpass resonator layers, placed on the left and right sides of the aperture layer, respectively. The length of the outer dielectric slabs is denoted by h_{A1} and h_{NB} . The semi-infinite spaces on each side of the multilayered structure are represented by semi-infinite transmission line with characteristic impedance Z_0 . The goal of our design guide procedure is to obtain electrical parameter combinations shown in Figure 2.2 given specific filter response characteristic. Most importantly, it will be demonstrated that predetermined thickness and transfer function of multilayered FSS filters.



Figure 2.3. (a) Generalized ECM of bandpass filter based on admittance inverters. (b) Admittance inverters are modeled as lumped component (inductive networks). (c) Coupling networks at the source and load are further modified to avoid the vegetative inductance at the terminations. (d) Simplified model after all the negative inductances are absorbed by adjacent resonators.

Our synthesis technique begins with the generalized ECM of a bandpass filter of order N using admittance inverters ($J_{0,1}$, $J_{1,2}$, \cdots , $J_{N-1,N}$, $J_{N,N+1}$). Given an Nth-order bandpass filter with center frequency f_0 and fractional bandwidth $\Delta = BW/f_0$, its ECM can be modeled as shown in Figure 2.3(a). The parallel L_{pi} , C_{pi} tanks denote the self inductance and self capacitance of the uncoupled resonators, whereas $J_{i,i+1}$ ($1 \le i \le N-1$) represent

the mutual coupling between two consecutive bandpass FSS layers due to the effect of the inverter layer and dielectric slabs. The coupling at the source and load is denoted by $J_{0,1}$ and $J_{N,N+1}$, respectively. The electrical parameters of the equivalent circuit can be obtained using the following equations in [51]:

$$L_{pi} = \frac{\Delta Z_0}{\omega_0 g_i}; \ \ C_{pi} = \frac{1}{\omega_0^2 L_{pi}}. \ \ (1 \le i \le N)$$
(2.1)

$$J_{i,i+1} = \Delta \omega_0 \sqrt{\frac{C_{pi} C_{p(i+1)}}{g_i g_{i+1}}}. \quad (1 \le i \le N-1)$$
(2.2)

where $\omega_0 = 2\pi f_0$ denotes the center angular frequency, and g_i ($0 \le i \le N + 1$) are the normalized parameters of the low-pass prototype filter type (e.g., Butterworth or Chebyshev response). These normalized values are used to find the required inductances and capacitances of the bandpass filters using frequency and element transformation [51]. For the predefined size of the outer substrates, the coupling at the source and the load are set to

$$J_{0,1} = Y_0 \sqrt{\frac{1}{1 + (\mu_0 \mu_r h_{A1} \omega_0 Y_0)^2}}$$
(2.3)

$$J_{N,N+1} = Y_0 \sqrt{\frac{1}{1 + (\mu_0 \mu_r h_{NB} \omega_0 Y_0)^2}}$$
(2.4)

The lumped equivalent circuit of each $J_{i,i+1}$ inverter consists of the series inductor $L_{i,i+1}$ between two shunt inductors of negative value $(-L_{i,i+1})$ such that $J_{i,i+1} = 1/(\omega_0 L_{i,i+1})$, as it is shown in Figure 2.3(b). The coupling network at the source and load is further modified to avoid the negative inductance value at the terminations, which cannot be absorbed by adjacent shunt elements [see Figure 2.3(c)]. The new inductance values

obtained at both terminals are

$$L_{A,1} = \frac{1}{\omega_0 Y_0} \sqrt{\left(\frac{Y_0}{J_{0,1}}\right)^2 - 1} = \mu_0 \mu_r h_{A1}$$
(2.5)

$$L_{N,B} = \frac{1}{\omega_0 Y_0} \sqrt{\left(\frac{Y_0}{J_{N,N+1}}\right)^2 - 1} = \mu_0 \mu_r h_{NB}$$
(2.6)

$$L_{A,1}^{'} = \frac{1 + (\omega_0 L_{A,1} Y_0)^2}{L_{A,1} (\omega_0 Y_0)^2}; L_{N,B}^{'} = \frac{1 + (\omega_0 L_{N,B} Y_0)^2}{L_{N,B} (\omega_0 Y_0)^2}$$
(2.7)

It should be noted that, when the outer substrates are not needed we have $J_{0,1} = J_{N,N+1} = Y_0$. Then from Equations (2.5) to (2.7), the series inductances at both terminals are shorted $(L_{A,1} = L_{N,B} = 0)$ while the shunts inductances are opened $(L'_{A,1} = L'_{N,B} = \infty)$. The simplified version of the ECM after absorption of the negative inductance values by adjacent resonators is shown in Figure 2.3(d). The inductance values obtained in Figure 2.3(d) are obtained as follows:

$$\frac{1}{L'_{p1}} = \frac{1}{L_{p1}} - \frac{1}{L'_{A,1}} - \frac{1}{L_{1,2}}$$
(2.8)

$$\frac{1}{L'_{pN}} = \frac{1}{L_{pN}} - \frac{1}{L_{N-1,N}} - \frac{1}{L'_{N,B}}$$
(2.9)

$$\frac{1}{L'_{pi}} = \frac{1}{L_{pi}} - \frac{1}{L_{i-1,i}} - \frac{1}{L_{i,i+1}}. \quad (2 \le i \le N - 1)$$
(2.10)

By predefining the size of the intersubstrates, the ECM in Figure 2.3(d) can be transformed to the one in Figure 2.4. This new model is obtained by converting the π -inductor network between consecutive resonators to T-network such that

$$L_{[i,i+1]L/R} = \mu_0 \mu_r h_{[i,i+1]L/R}. \quad (1 \le i \le N-1)$$
(2.11)



Figure 2.4. π -inductor network between adjacent resonators is used from Figure 2.3(d) and converted to T-inductor network.

The other required constituting elements in the modified ECM can be obtained as follows:

$$\frac{1}{L_{p1}''} = \frac{1}{L_{p1}'} - \frac{L_{1,2} - L_{[1,2]L} - L_{[1,2]R}}{L_{1,2} \times L_{[1,2]L}}$$
(2.12)

$$\frac{1}{L_{pN}^{''}} = \frac{1}{L_{pN}^{'}} - \frac{L_{N-1,N} - L_{[N-1,N]L} - L_{[N-1,N]R}}{L_{N-1,N} \times L_{[N-1,N]R}}$$
(2.13)

$$\frac{1}{L_{pi}''} = \frac{1}{L_{pi}'} - \frac{L_{i-1,i} - L_{[i-1,i]L} - L_{[i-1,i]R}}{L_{i-1,i} \times L_{[i-1,i]R}} - \frac{L_{i,i+1} - L_{[i,i+1]L} - L_{[i,i+1]R}}{L_{i,i+1} \times L_{[i,i+1]L}}.$$
(2.14)

$$(2 \le i \le N - 1)$$

$$\frac{1}{L_{i,i+1}^r} = \frac{L_{i,i+1} - L_{[i,i+1]L} - L_{[i,i+1]R}}{L_{[i,i+1]L} \times L_{[i,i+1]R}}. \quad (1 \le i \le N - 1)$$
(2.15)

In order to reflect the topology of the bandpass spatial filter proposed in Figure 2.1, the series inductors and their required shunt capacitances are used to model each transmission line [52] with characteristic impedance $Z_d = Z_0 \times \sqrt{\mu_r/\varepsilon_r}$, as shown in Figure 2.2. This transformation is accurate as long as the electric length of each transmission line is small such that their phase $\theta = \beta h < 30^\circ$ within the operation band of the filter. The effective shunt capacitance values shown in Figure 2.2 are obtained from telegrapher's equation of the short transmission lines

$$C_{p1}'' = C_{p1} - \frac{\varepsilon_0 \varepsilon_r h_{A1}}{2} - \frac{\varepsilon_0 \varepsilon_r h_{[1,2]L}}{2}$$
(2.16)

$$C_{pN}^{"} = C_{pN} - \frac{\varepsilon_0 \varepsilon_r h_{NB}}{2} - \frac{\varepsilon_0 \varepsilon_r h_{[N-1,N]R}}{2}$$
(2.17)

$$C_{i}^{''} = C_{i} - \frac{\varepsilon_{0}\varepsilon_{r}h_{[i-1,i]R}}{2} - \frac{\varepsilon_{0}\varepsilon_{r}h_{[i,i+1]L}}{2}. \quad (2 \le i \le N-1)$$
(2.18)

The calculated $(L_{pi}^{"}, C_{pi}^{"}, L_{[i,i+i]}^{r})$ combinations provide the desired filter response for predefined dielectric thickness between the metallic layers for the design of multilayered FSS. It should be emphasized that the design flexibility for various dielectric thicknesses is compensated by adjusting the resonators and the aperture features [i.e., for different predefined dielectric electrical lengths, different values of resonators features $(L_{pi}^{"}, C_{pi}^{"}, L_{[i,i+i]}^{r})$ will be obtained from the synthesis]. This is due to the fact that the coupling coefficient between two consecutive resonators not only depends on the aperture size but also the resonator features as well as the dielectric slabs. This scenario explains the unintentional change in the filter response obtained in our previous approach due to the lack of a synthesis procedure that should be used to compensate the detuning of desired filter characteristics.

The final implementation step maps these electrical parameters to the physical dimensions of the FSS. For the example structures, the dimensions of the bandpass FSS layers shown in Figure 2.1 can be obtained from approximate resonance equations of the array of square loop slots using their constituent elements $(L_{pi}^{"}, C_{pi}^{"})$ [53]. The associate reactance $X_{L_{pi}^{"}}$ and susceptance $B_{C_{pi}^{"}}$, of the inductance and capacitance, respectively, are obtained using

$$\frac{X}{Z_0} = \omega_{pi} L''_{pi} = \cos(\theta_{inc}) F(a_i, a_i - b_i, \lambda_i, \theta_{inc})$$
(2.19)

$$\frac{B_1}{Y_0} = 4\sec(\theta_{inc})F(a_i, b_i, \lambda_i, \theta_{inc})$$
(2.20)

$$\frac{B_2}{Y_0} = 4F(b_i - s_i, \, s_i, \, \lambda_i, \, \theta_{inc})$$
(2.21)

$$\frac{B}{Y_0} = \omega_{pi} C_{pi}'' = \left(1.75 \frac{B_1}{Y_0} + 0.6 \frac{B_2}{Y_0}\right) \varepsilon_{reff}$$
(2.22)

where λ_i , ω_{pi} , and *F* are given by [53]

$$\lambda_{i} = \frac{2\pi}{\omega_{pi}\sqrt{\varepsilon_{0}\mu_{0}\varepsilon_{reff}}}; \quad \omega_{pi} = \frac{1}{2\pi\sqrt{L_{pi}^{"}C_{pi}^{"}}}$$
(2.23)

$$F(a_i, u, \lambda_i, \theta_{inc}) = \frac{a_i}{\lambda_1} \left[\ln\left(\csc\frac{\pi u}{2a_i}\right) + G(a_i, u, \lambda_i, \theta_{inc}) \right]$$
(2.24)

$$G(a_i, u, \lambda_i, \theta_{inc}) = \frac{1}{2} \times \frac{(1 - \beta^2)^2 \left[\left(1 - \frac{\beta^2}{4} \right) (A_+ + A_-) + 4\beta^2 A_+ A_- \right]}{\left(1 - \frac{\beta^2}{4} \right) + \beta^2 \left(1 + \frac{\beta^2}{2} - \frac{\beta^4}{8} (A_+ + A_-) + 2\beta^6 A_+ A_- \right)}$$
(2.25)

with

$$A_{\pm} = \frac{1}{\sqrt{\left[1 \pm \frac{2a_i \sin \theta_{inc}}{\lambda_i} - \frac{a_i \cos \theta_{inc}}{\lambda_i}\right]}} - 1$$
(2.26)

$$\beta = \sin\left(\frac{\pi u}{2a_i}\right). \tag{2.27}$$

In the above equations, ε_0 and μ_0 are the free space permittivity and permeability, ε_{reff} is the effective relative permittivity of the dielectric surrounding the metallic bandpass resonator layer, and θ_{inc} is the angle of incidence of the EM wave. The effective relative permittivity $\varepsilon_{reff} \approx \varepsilon_r$, since the metallic FSS is embedded on both sides by the same substrate. However, an accurate expression of ε_{reff} that depends on the substrates' thicknesses and the period of the FSS can be found in [54].

Since the aperture layer is a wire grid, the physical size $r_{i,i+1}$ of the apertures can be approximated using [17] where *a* is the unit cell size, and μ_{reff} is the effective permeability of the medium.

$$L_{[i,i+1]}^{r} = \frac{a}{2\pi} \mu_{0} \mu_{reff} \ln\left[\csc\left(\frac{\pi(a-r_{i,i+1})}{2a}\right)\right]$$
(2.28)

This reverse engineering technique provides a better understanding of the physics behind the proposed structure. For this synthesis technique, the periodicity (*a*) of the design can be fixed in order to find the remaining physical dimensions of the design. Thus, it is noteworthy to mention that the set of solutions (b_i , s_i , $r_{i,i+1}$) are not unique. This flexibility of the synthesis process allows the designer to obtain feasible physical dimensions. As a proof of concept, this design guide will be used in the next section for the design of Butterworth multilayer filters of order N = 3 with various predefined (customized) thicknesses.



Figure 2.5. Topology of the proposed third-order bandpass FSS.

2.4 Proof of the Proposed Synthesized Technique

2.4.1 Synthesis of a Third-Order Bandpass FSS with Customized Thickness

In order to validate the procedure described in the previous section, a third-order Butterworth response-type bandpass FSS filter operating at center frequency $f_0 = 10 \ GHz$ and fractional bandwidth of $\Delta = 10\%$, is chosen to be implemented with various design total thicknesses comprised between $\lambda_0/12$ and $\lambda_0/6$ (where λ_0 is the free space wavelength at f_0). Following the design approach, four different configuration thicknesses of the FSS are implemented. RT/duroid 6010 from Rogers Corp. with available standard thicknesses is used as dielectric slabs. This substrate has a relative permittivity value of $\varepsilon_r = 10.2$ and loss tangent of tan $\delta = 0.0023$. The third-order bandpass spatial filter is shown in Figure 2.5. The design has a symmetrical architecture due to the Butterworth response type.

Table 2.1. Low-pass (Butterworth) prototype normalized element and electrical parameters of the
third-order bandpass FSS for the ECM shown in Figure 2.3(a).

Parameter	$g_1 = g_3$	82	$J_{0,1}(\Omega^{-1})$	$J_{1,2}(\Omega^{-1})$	
Value	1	2	0.0026	0.0027	
Parameter	$J_{2,3}(\Omega^{-1})$	$J_{3,4}(\Omega^{-1})$	$L_{p1}(nH)$	$L_{p2}(nH)$	
Value	0.0027	0.0026	0.60001	0.30001	
Parameter	$L_{p3}(nH)$	$C_{p1}(pF)$	$C_{p2}(pF)$	$C_{p3}(pF)$	
Value	0.60001	0.42216	0.84432	0.42216	

The synthesis procedure begins by finding the electrical parameters of a third-order bandpass filter from admittance inverters shown in Figure 2.3(a) using Equation (2.1)-(2.4). The results are accessible in Table 2.1. By following the procedure described in Section 2.2, Equations (2.5)-(2.18) are then used to find the electrical parameters $(L_{pi}^{"}, C_{pi}^{"}, L_{[i,i+i]}^{r})$

Design #	Total	h_1	$h_2 = h_3$	а	b_1	b_2	$s_1 = s_2$	r _{1,2}	$f_0(GHz)$	$\Delta(\%)$
_	Height									
1 (Sim.)	$\lambda_0/12$	0	0.635	4.5	4.30	3.95	0.19	1.75	10	10
2 (Sim.)	$\lambda_0/10$	0.254	0.635	4.4	3.84	4.10	0.24	1.60	10	10
3 (Sim.)	$\lambda_0/8$	0.635	0.635	4.1	3.45	3.90	0.20	1.66	10	10
4 (Sim.)	$\lambda_0/6$	0.760	0.760	4.0	3.40	3.80	0.24	1.78	10	10

Table 2.2. Summary of the four design configurations (physical parameters are in mm).

in Figure 2.2 for four different configuration design total thicknesses comprised between $\lambda_0/12$ and $\lambda_0/6$ using only available standard thicknesses. Thus, different values of the resonators and aperture layers features $(L_{pi}^{"}, C_{pi}^{"}, L_{[i,i+i]j}^{r})$ for $1 \le j \le 4$, where *j* is denoted for each customized FSSs design, are obtained. The next step of the procedure is to map these electrical parameter values to the physical dimensions of the resonators and aperture layer using Equations (2.19)-(2.28). However, these initial values are used as a starting point because approximate equations for the constituent elements of the square loop slot and the wire grid are used. Taking away the intense computational time, a simple fine-tuning process using HFSS is carried out to achieve the desired response.



Figure 2.6. FW simulation results for transmission and reflection coefficients of the four configuration designs FSSs along with the ECM results.



Figure 2.7. Frequency response of Design #4. Simulated ECM (Figure 2.5) and FW result using initial physical parameters compared with the one after the fine-tuning process.

The detailed final physical dimensions resulting from the above steps are summarized in Table 2.2. As expected, when the structure is compressed down to a thinner thickness, stronger coupling occurs between resonators, requiring smaller aperture dimensions to compensate the coupling strength for selective filtering response. The frequency response of all four design configurations based on full-wave (FW) EM-simulation along with the ECM results is illustrated in Figure 2.6. The results obtained show that the design can be scaled to different thicknesses while virtually maintaining the same filter characteristics (center frequency and bandwidth) and clearly abide to our expectations. As can be noticed that all the dielectric thicknesses used for the four designs are commercially available. In order to perceive the accuracy of the synthesis process, the ECM result and the FW result using initial physical parameters are compared with the one after the fine-tuning process for Design #4. As can be seen from Figure 2.7, the margin of error is almost imperceptible.



Figure 2.8. FSS filter measurement setup.

2.4.2 Fabrication and Measurement Results

Both Design #2 and Design #4 are fabricated and then tested to experimentally validate our design technique. RT/duroid 6010 laminated with 1-oz copper is used for all dielectric layers. However, beforehand, the dielectric substrates are bonded together using the 4-mil Rogers RO4450F ($\varepsilon_r = 3.52$ and $\tan \delta = 0.004$) bonding layer, as shown in inset of Figure 2.8. The inserted bonding layers have a noticeable effect on the frequency response of the design since it adds extra length to the design thickness. So it is convenient to take into account their presence during the synthesis process. Only the capacitive gap between the resonators and the aperture size are modified to accommodate this perturbation. The new dimensions for the prototype #2 are: $s_1 = s_2 = 0.2$; $r_{1,2} = 1.8$ (in millimeters). For the prototype #4, we obtained: $s_1 = s_2 = 0.15$; $r_{1,2} = 2$ (in millimeters).

The performance of the FSSs (Design #2 and Design #4) with the bonding layer is



Figure 2.9. Simulated filter response of Design #2 at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.



Figure 2.10. Simulated filter response of Design #4 at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.

investigated across different angles of incidence for both transverse electric (TE) and transverse magnetic (TM) polarizations. Figures 2.9 and 2.10 predict a robust filter response when the spatial filters are illuminated from various polarizations and oblique angles $(0^{\circ} \le \theta \le 60^{\circ})$. A total array size of 120 mm × 120 mm corresponding to $4\lambda_0 \times 4\lambda_0$ (where λ_0 is the free space wavelength at 10 GHz) is fabricated for both designs. Two frames using 3-D printer (uPrint SE Plus) are fabricated to align the multilayers and clamp each FSS prototype. A pair of PE9887-11 horn antennas is used to obtain the frequency response of the design in free-space measurement setup. A large metallic wall is used to block direct coupling between the transmitting and receiving antennas. A small aperture of the size of the FSS is carved out at the center to mount the FSS, as shown in Figure 2.8.



Figure 2.11. Transmission coefficient of the FSSs showing simulated ECM and FW results taking into account the bonding layers along with the experimental measurement (Exp. Meas.) results for TEM plane wave.

The calibration of the system is carried out by first measuring the transmission coefficient without the device under test, and then the reflection coefficient is also obtained when the aperture is covered by a thick conductive sheet. Gating window is applied to reduce ripples in the measurement caused by multiple reflections of the waves. The actual frequency response of the spatial filters is obtained by normalizing their measured results obtained in the presence of the FSSs with the calibrated values. As can be seen from Figure 2.11, a good agreement between simulation (ECM and FW) and measurement results from both configuration designs is obtained. Although slight discrepancies between measured and simulated filter response can be observed, these minor inconsistencies are mainly due to fabrication tolerance and imperfect assembly of the multilayers. However, the important design parameters (center frequency and bandwidth) are well maintained. The measured performance of the designs across different angles of incidence ($0^\circ \le \theta \le 60^\circ$) for both TE and TM polarizations is also experimentally investigated. Results obtained (Figures 2.12 and 2.13) show robust and stable filter response for both prototypes. Although at 60° , a slight deviation of the center frequency is obtained. The average insertion loss obtained

within the scan angle is about 1.5 dB as shown in the inset of Figures 2.12 and 2.13. The increase of the insertion loss compared to the simulated result can be due to the finite size of the prototypes.



Figure 2.12. Measured filter response of Prototype #2 at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.



Figure 2.13. Measured filter response of Prototype #4 at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.

The responses of the proposed design show a good filter performance for practical applications demanding flat in-band top and sharper band skirt for higher out-of-band suppression. Although no TZ is generated neither at the lower or upper sideband, at least 20-dB out-of-band suppression is obtained at frequencies less than 9 GHz and above 11 GHz for all designs which show better passband selectivity compared to the results obtained in [21]. For our previous approach, TZs are located at the upper side of the passband due to the presence of the mixed electric and magnetic coupling but the synthesis

would be very complicated. A more robust response (compared to [45]) is achieved for a wide range of angle of incidences up to 40° for both TE and TM.

2.5 Design Profile and Fabrication Tolerance

Through our analysis, we have provided a degree of conformity for thickness customizable high-order bandpass FSS. However, some limitations are required for the design of a very thin profile. During the synthesis process, we have neglected the coupling between adjacent resonator layer and aperture layer and only consider the direct coupling between the resonators. However, for much closer proximity between the metallic layers, (in the order of $\leq \lambda_0/100$), the mutual coupling between the aperture layer and the resonator layer caused by the evanescent higher order Floquet modes, can be high and should be taken into account. This is not the case for all four designs, as the separation between adjacent resonators and aperture layer was thick enough ($\geq \lambda_0/47$). This assumption is validated since no perceptible difference is obtained between the FW simulation response and the ECM response. For the design of very thin bandpass filter with much closer proximity between adjacent metallic layers ($\leq \lambda_0/100$), two major problems can arise. First, the ECM needs to be adjusted. Second, the larger capacitance or inductance will be required which can limit the fabrication process.

While the complexity of the fabrication is pertained by the dielectric profile, the sensitivity of the filter response to its design parameter should also be investigated. Uncertainties in the property of the material used can affect the response of the FSS in terms of center frequency, bandwidth, and insertion loss. A parametric study is provided to show the variation of the response against unavoidable tolerance in the dielectric constant, thickness, and the dielectric loss tangent of the substrate used. The sensitivity analysis is also performed for a slight change in the aperture size which can be due to the technology used for the fabrication process. Note that this feature is chosen because the aperture size mainly controls the external quality factor which depends on the coupling between the resonators.

Characteristics of the frequency response include the center frequency, the bandwidth, and the insertion loss. The physical dimensions for the Design #2 (see Table 2.2) are used as the nominal dimension for the analysis. Based on the coupled-resonator filter topology, the insertion loss of the FSS is usually dependent on the coupling level of the resonators and the material losses including the dielectric and metallic losses. For this parametric study, the coupling level is satisfied and we obtained the transmission coefficient of the FSS for various dielectric losses ($\tan \delta_{\varepsilon}$). The impact of the variation of the dielectric loss from 0.002 to 0.02 is shown in Figure 2.14(a). It can be seen that by increasing the dielectric loss within that range, the insertion loss in the passband is mainly affected while the bandwidth is slightly reduced leaving the location of the center frequency virtually unchanged. With no dielectric loss the insertion ($\tan \delta_{\varepsilon} = 0$), loss is about 0.4 dB, which is due to the metallic loss. When the dielectric loss increases to 3.2 dB which is unacceptable for practical application. Therefore, a low-loss dielectric is required for this design.

For most available commercial substrates, the tolerance of the dielectric constant is at most ± 0.25 and about $\pm 0.05 \text{ mm}$ for the thickness. The analysis on the uncertainty of the dielectric constant [see Figure 2.14(b)] shows that the center frequency of the spatial filter is only affected as it slightly changes from 9.9 to 10.2 GHz. Then, a variation of all the substrates thickness within the range of their tolerance is investigated. The results shown in Figure 2.14(c) predict a slight deviation of the center frequency from 9.9 to 10.1 GHz. Finally, the variation of the aperture size from 1.5 to 1.7 mm shows that the required external quality factor for the Butterworth response-type bandpass filter is disturbed. As a result, when the aperture size increases, the bandwidth of the FSS also increases, as



Figure 2.14. Impact of the parametric analysis on the frequency response of the third-order bandpass FSS. The nominal dimension (in millimeters) of Design #2 is used except for the parametric variable. (a) Effect of the dielectric loss from the substrate. (b) Uncertainty of the dielectric constant ($\varepsilon_r \pm 0.25$). (c) Uncertainty of the substrates thickness ($h \pm 0.05$). (d) Imprecision of the aperture size due to fabrication issues.

shown in Figure 2.14(d). The slight discrepancy in the bandwidth that were obtained in Figure 2.11 could be due to this fabrication error. All these behaviors obtained from the parametric analysis are normal as it also happens for microwave filters. However, the sensitivity level of our proposed design is within a norm that can be tolerated by most practical applications.

2.6 Chapter Conclusion

Advanced design techniques for high-performance FSSs are implemented in this chapter. A design synthesis based on microwave filter theories and inverters is develop to realize compact/size controllable high order bandpass FSS for radiative applications. With a better design control of various artificial surfaces, researchers and engineers will be equipped with alternate design tools to meet and exceed the technological demands of future electromagnetic devices for wireless applications. The proposed technique is by no means limited to FSSs as it can be applied to any applications (spatial filters, absorbers, and antennas) that require high-performance size-control on the physical features of the constituent subcomponents.

Chapter 3

Plasma-Enabled Reconfigurable Low-Profile Frequency Selective Surface for Harsh EM/RF Environment

3.1 Chapter Introduction

Reconfigurable FSSs have been investigated using RF MEMS switch and semiconductor varactors. Unfortunately, they suffer from complex biasing control; they also have a limited tunable range capability and are not well suited to handle high power intensity field. The technology of using plasma to attenuate high power microwave energy is investigated to address these issues. The proposed active FSS is implemented to stand up to the harshness of aerospace, military, or satellite applications and therefore improve the state of the art in power dependent frequency-selective protection to provide a secured space sensor operating in harsh EM/RF environments. Various applications may require different specification, but the desired performances for abovementioned EM radiation protection systems are as follows:

- low profile plasma based reconfigurable band pass spatial filter centered at 3 GHz with the ability to integrate into existing structures
- switchable passband response with fast turn-on times
- greater than 30 dB out of band shielding effectiveness at OFF state and negligible in-band insertion loss at OFF state
- greater than 30 dB in-band transmission suppression at ON state with low power consumption

- fractional bandwidth (BW): > 5%, selected as trade off between selectivity and insertion loss
- polarization independent (this feature is best suited for randomly moving and/or rotating systems as it minimizes polarization mismatch loss
- scan angle capability up to 45°
- material thickness expected to be less than 50 mm
- capable of withstanding ionizing radiation, shock, vibration and temperature range of -90 to +90°C.

In this chapter, we demonstrate a switchable second order FSS based on discrete plasmashells technology. Two designs are investigated for this purpose to provide a fast switchable passband response with both in-band and out-of-band protection. In the first topology, a second order bandpass FSS is realized by cascading two single pole FSS layers about quarter wavelength apart from each other. A much compact and flexible design is adopted for the second topology by introducing inverter inter-layer between two resonators such that the design thickness can be considerably reduced.

3.2 Plasma-shells Concept and its Electrical Properties

Plasma's electrical properties and its advantages in stealth applications have been extensively studied [55, 56]. Theoretically, plasma medium can be manipulated at the appropriate frequency to transmit/delay, reflect, or absorb electromagnetic wave. Notably, it has the unique property to withstand high power microwave energy [57, 58]. However, the realization of a practical large scale plasma device has presented challenges in integrating the plasma with the EM device [59–61], and assuring control over its properties (plasma density, gas composition, and pressure). Often plasma devices previously employed have been bulky, fragile and not hermetic. Mostly, there is no control over the gas content and pressure. Consequently, they have a short operating life and will not stand up to the rigors of aerospace, military, or satellite applications.



Figure 3.1. (a) Plasma-shells cutaway showing internal plasma. (b) IST manufacturing capability produces shells of various size, shape, and texture [3]

Recently, controlled gas encapsulating ceramic chambers patented and commercially available by Imaging Systems Technology Inc. circumvents the aforementioned problems. These plasma-shells generate plasma when properly biased. The ceramic, gasencapsulating shells are ideal for high-power microwave and electromagnetic pulse protection because they are rugged, hermetic, operable at extreme temperatures, and insensitive to ionizing radiation. When energy is applied across the exterior surface of the plasmashell, the encapsulated gas ionizes as shown in Figure 3.1(a). Ionized gas can emit, reflect, or absorb EM energy. Standard sizes of plasma-shells range from 0.5 to 10 mm, and are compatible with surface-mount technology assembly techniques. Shells are extremely lightweight and can be manufactured in different shapes: spheres, cylinders, cubes, oblate spheroids, rectangular prisms, and other complex shapes as shown in Figure 3.1(b). Primary materials of shells include: Y_2O_3 , Z_rO_2 , S_iO_2 , Al_2O_3 , carbon steel, and various glasses. Noble gases (helium, neon, argon, krypton, xenon) are often used as a mixture inside the hermetic shell. The ability to customize the shells (material, size, shape, texture, density) allows them to be easily integrated into many enabling applications. We will use these plasma-shells as real-time fast electronic switches in the design of practical large scale switchable electromagnetic field blocking FSS for the protection of EM systems subjected to high power microwave.

The internal gas volume can be modeled as a complex frequency dependent material, with the permittivity ε_p expressed as follows [62]:

$$\varepsilon_p = \varepsilon_0 (\varepsilon_r' - j \frac{\sigma_p}{\omega \varepsilon_0}), \tag{3.1}$$

where $\varepsilon_0 = 8.854 \times 10^{-12} F/m$ is the permittivity of free-space, ε'_r the real part of relative permittivity, $\sigma_p[S/m]$ is the conductivity and $\omega[rad/s]$ is the operating angular frequency.

$$\varepsilon'_r = 1 - \frac{\omega_p^2}{\omega^2 + \nu^2} \text{ and } \sigma_p = \frac{\omega_p^2 \cdot \nu \cdot \varepsilon_0}{\omega^2 + \nu^2} \left[\frac{S}{m}\right],$$
(3.2)

where $\omega_p \ [rad/s]$, and $\nu \ [rad/s]$ are plasma frequency and electron collision frequency, respectively. It is noted that the plasma frequency ω_p depends on the electron density $n_e \ [m^{-3}]$ as:

$$\omega_p = \sqrt{\frac{n_e \cdot e^2}{\varepsilon_0 \cdot m_e}} = 56.4\sqrt{n_e},\tag{3.3}$$

where e is the electron charge ($e = 1.6 \times 10^{-19}$ C), and m_e is the electron mass ($m_e = 9.1 \times 10^{-31}$ Kg). The electron density of the ionized plasma is electrically controlled by

applied voltage. The electron collision frequency for a noble gas is obtained from [63]:

$$\nu[s^{-1}] = \frac{8}{3 \times \pi^2} N\left(\frac{m_e}{2K_B T_e}\right)^{5/2} \int_0^\infty \vartheta^5 Q^{(m)}(\vartheta) exp\left(-\frac{m_e \vartheta^2}{2K_B T_e}\right) d\vartheta, \tag{3.4}$$

where $T_e[K]$, $N[cm^{-3}]$, $K_B[J \cdot K^{-1}]$, $\vartheta[m \cdot s^{-1}]$, and $Q^{(m)}(\vartheta)$ are temperature, gas number density, Boltzmann's constant, velocity, and momentum transfer cross section of the electrons, respectively. The gas number density $N[m^{-3}]$ is obtained from the ideal-gas equation of state as:

$$N = \frac{P}{K_B T_e}.$$
(3.5)

where P [Pa] is the gas pressure. It can be seen that while the electron collision frequency can be controlled by the gas contents, the plasma frequency is controlled by the biasing voltage. Thus, for fixed gas contents, the real part of relative permittivity of the plasma medium can be altered by adjusting the biasing voltage.

Generally, a plasma medium can be treated as a conductor with conductivity σ_p or as a dielectric with permittivity ε_p based on its electron collision frequency, its plasma frequency and the operating frequency of the system [62]. The first scenario is met for frequencies where $\omega \ll \omega_p, \nu$. The conductivity of the plasma is then reduced to its dc form such that $\sigma_p \rightarrow \sigma_{dc} = \frac{\omega_p^2 \cdot \varepsilon_0}{\nu}$. Such characteristics of the plasma is used in this Chapter to design switchable frequency selective surfaces for electronic protection in the following sections. For the second scenario regime, where the bulk plasma is used as a dielectric with variable permittivity, we have $\nu < \omega < \omega_p$ which is practical for plasma driven at RF discharge.

3.3 Plasma-based ON-OFF Switchable Second Order FSS (Quarter Wavelength Concept)

3.3.1 Detail Specification



Figure 3.2. Unit cell representation of the first order switchable plasma-shell integrated FSS design. Units are in mm

In [64], plasma-shells have been used to demonstrate a tunable second order large area band pass FSS operating in X-band. In this example, however, only 7 dB in-band suppression is achieved when plasmas were sustained with 1400 V peak-to-peak square voltage. An alternative design that gives more flexibility and agility implementation has been proposed to improve switchable in-band and out of band suppression in the S-band. For example in [64], plasma- shells about quarter wavelength height are sandwiched between two FSS layers that serve as plasma excitation. Thus, this topology requires relatively long plasma shell when the design is to be scaled to low frequencies. Plasma shells with such thickness are in general more challenging to realize. Also in case where few shells breakdown or not functional, it will be impossible to replace them.



Figure 3.3. Simulated first order band pass filter response in OFF state at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.

First we proposed, a single pole FSS with top layer composed of miniaturized Jerusalem cross structure enclosed by inductive metallic loop arranged in two dimensional periodic lattice. As shown in the unit cell (Figure 3.2), each Jerusalem cross structure is connected to the metallic wire by a micro plasma shell. The optimized shell dimension used in this design is 6 x 2.2 x 2.2 mm. Each shell is filled with 0.1 % Argon - 99.9 % Neon mixture at 300 K electron temperature. The electron collision frequency is set to $v = 4.85 \times 10^{10} rad/s$. The bottom layer is a high impedance sub-wavelength metallic wire grid. This layer is connected to the JC layer through conductive vias and mainly serve as biasing network for the active components. FR4 dielectric with $\varepsilon_r = 4.4$, loss tangent of tan $\delta = 0.017$ and a



Figure 3.4. Simulated first order filter response in ON state ($n_e = 3.6 \times 10^{12} cm^{-3}$) at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.

thickness of t = 1.524 mm is used as substrate. The model utilizes a three-layers stack-up FR4 in order to accommodate the biasing traces within the electrically insulating materials. Due to the symmetrical layout of the design, this filter is insensitive to the polarization of the electric field. Physical dimensions of the unit cell are optimized to provide a first order band pass response centered at $f_0 = 3 \text{ GHz}$ when the plasma is in OFF state. The simulated filter response with an incident plane wave illuminated from various incidence angles ($0^{\circ} \le \theta \le 45^{\circ}$), for both TE and TM polarizations in Figure 3.3 demonstrates a stable filter response. The results show a first order band pass response centered at 3 GHz with 1.2 dB in-band insertion loss across the scan angle. The 3 dB bandwidth is about 9 %. For TE case, transmission peak is formed around 3.6 GHz as the incidence angle is increased. This is due to the crooked mode that is emblematic of the Jerusalem cross structure. For plasma ON state, the plasma electron density value is approximated to $n_e = 3.6 \times 10^{12} \text{ cm}^{-3}$ equivalent to $\omega_p = 1.07 \times 10^{11} \text{ rad/s}$. The simulated transmission and reflection response for various angle of incidence for both TE and TM polarization (Figure 3.4) exhibit robust switchable average effective shielding of about 20 dB for both in

band and out of band operation.



Figure 3.5. Unit cell representation of the second order switchable plasma-shell integrated FSS design by cascading two of the single pole FSS layers apart from each other about quarter wavelength (12.7 mm). Total thickness is about $\lambda_0/5$. Units are in mm

The multi-poles FSS is realized by cascading two of the single pole FSS layers about quarter wavelength apart from each other to realize a traditional second order bandpass filter. The topology of the second order tunable FSS is shown in Figure 3.5. The total thickness of the structure is about 21 mm (equivalent to $\lambda_0/5$) and the periodicity in the order of $0.15\lambda_0$. Although the overall thickness of the device is much thicker than the first proposed design, it still meets the design specifications in term of thickness. At OFF state ($\omega_p = 0$), the filter response is characterized for wave incident from normal and oblique angles (Figure 3.6). The simulated results show highly selective second-order bandpass response centered at 3 GHz with 1.9 dB in-band insertion loss. The results also show



Figure 3.6. Simulated second order band pass filter response in OFF state at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.



Figure 3.7. Simulated second order filter response in ON state ($n_e = 3.6 \times 10^{12} \ cm^{-3}$ equivalent to $\omega_p = 1.07 \times 10^{11} \ rad/s$) at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.

from various oblique angles for both TE and TM waves compared to previous prototypes. The 3 dB bandwidth is about 6 %. The simulated transmission and reflection response for various angle of incidence for plasma ON state ($n_e = 3.6 \times 10^{12} \text{ cm}^{-3}$) are shown in Figure 3.7. The in-band and out-of-band transmission coefficient is suppressed about 40 dB average. The performances of this second order FSS meet the design requirements.



3.3.2 Fabrication and Experimental Results

Figure 3.8. Illustration of one fabricated panel at OFF and ON state.

Two large panels of size 12 in. \times 12 in. each (array of 19 \times 19 unit cells) are fabricated on FR4 dielectric substrates. Each panel hosts 1,400 discrete plasma shell filled with 0.1 % Argon – 99.9 Neon mixture at 300 K electron temperature. The value of the gas pressure (P = 240 Torr) is estimated by solving a simple theoretical expression for neon gas derived in [65] at 300 K electron temperature:

$$\nu[s^{-1}] = 8.63 \times 10^{-18} \times \left(\frac{P}{K_B T_e}\right) \times T_e^{0.833} \text{ for } 10^2 K \le T_e \le 5 \times 10^3 K, \tag{3.6}$$

Plasma-shell are manually bonded to the metallic layers with 0.5 mm thickness low resistivity epoxy. One fabricated panel is shown in Figure 3.8 at OFF state. The panel is excited with boost DC-AC converter used to generate the power and sustain the plasma. The transformer steps up the DC voltage by a factor of 12 and produces a square wave output with a low driving frequency ranging from 25 kHz to 1 MHz. The excited panel is also shown in Figure 3.8. Two panels are cascaded about quarter wavelength (12.7



Figure 3.9. (a) Photograph of the test setup (free space) for the FSS measurement. Two panels are cascaded about quarter wavelength; (b) Measured second order band pass filter response in OFF state for both TE polarization and TM polarization.

mm) and the test setup is illustrated in Figure 3.9(a). A pair of S-band horn antennas is used to obtain the frequency response of the design in free-space measurement setup. A large metallic wall is used to block direct coupling between the transmitting and receiving antennas. An aperture of the size of the FSS is carved out at the center of a large metallic frame to mount the test sample as shown in Figure 3.9(a). The calibration of the system is carried out by first measuring the transmission coefficient without the device under test, and then the reflection coefficient is also obtained when the aperture is covered by a thick conductive sheet. Gating window is applied to reduce ripples in the measurement caused by multiple reflections of the waves. The actual frequency response of the spatial filters is obtained by normalizing their measured results obtained in the presence of the FSSs with the calibrated values. Transmission scatter (S)-parameters were measured with a Hewlett-Packard/Agilent 8720B vector network analyzer (VNA) with through calibration and smoothing over 0.3 GHz to remove ringing. The response of the FSS for both TE and TM wave at OFF state is illustrated in Figure 3.9(b). At OFF state, the structure achieved a second order bandpass filter centered at 2.7 GHz and achieved a sharper roll off and higher out of band rejection. The shift of the center frequency from 3 GHz to 2.7 GHz is due to fabrication errors tolerance. The 3 dB bandwidth is about 7%. Also the insertion

loss has increased about 4.5 dB. This increase is due to the overall dielectric and Ohmic loss from the stacked boards and also the lossy conductive epoxy used as electrodes.



Figure 3.10. Measured second order filter response at ON state for both TE polarization and TM polarization. (a) with 900 V peak voltage; (b) with 1200 V peak voltage

Both layers are then biased in parallel connection with low-frequency (1 MHz) square wave of maximum 900 V peak voltage. The plasma activation response time is less than 100 ns and switchable protection level of about 35 dB is obtained for both TE and TM modes as shown in Figure 3.10(a). Another test was conducted with a new power supply that allows maximum voltage up to 1200 V peak voltage. The second order design performance is again characterized using the free space measurement setup. The data obtained are illustrated in Figure 3.10(b). At ON state, the device responds strongly to the EM field and switchable passband attenuation level exceeding 45 dB is obtained for both perpendicular and parallel polarization of the EM wave. The improved shielding effectiveness compared to the previous measurements is due to the increase of the plasma electrons density as a result of the higher input voltage.

3.4 Plasma-based ON-OFF Compact Switchable Second Order FSS (Inverter Inter-Layer Concept)

3.4.1 Detail Specification

An alternative FSS design is realized by inserting metallic wire grid in between two identical band pass FSS layers composed of periodic miniaturized complementary Jerusalem cross structure. Discrete tunable plasma-shells are incorporated between the metallic layers to add configurability feature to the spatial filter. The design model shown in Figure 3.11 is optimized to provide a maximum flat band pass response around 3 GHz when the plasma is in OFF state. Available plasma shells with dimension of 4.4 x 4.4 x 2.6 mm are used. The shells are filled with 0.1 % Argon - 99.9 % Neon mixture at 300 K electron temperature. The electron collision frequency is set to $\nu = 4.85 \times 10^{10}$ rad/s. The electron density n_e of the ionized internal volume is electrically controlled by the sustainer voltage. The plasma shells are strategically located where the electric field is maximum in order to effectively switch the transfer function of the filter for plasma ON state. The model utilizes the three-layer stack-up FR4 in order to accommodate waveguide measurements. Specifically, this configuration is used in a way to avoid the shorting of the FSS structures to the waveguide and also maintain the symmetry of the structure to achieve polarization independence of the structure. Plasmas are sustained with low-frequency (1 MHz) square waves voltage connected in parallel configuration. Detail dimensions of the unit cells of each layer of the model are also denoted in Figure 3.11. The periodicity and overall thickness of the structure are both in the order of $\lambda_0/10$, where λ_0 is free space wavelength at 3 GHz.

The simulated filter response for plasma OFF state with an incident plane wave illumi-



Figure 3.11. Topology of the proposed design unit cell including detailed unit cells of each layer (Top view of the CJC resonator and Top view of the inverter inter-layers in contact with the plasma shells); All unit are in mm. Total thickness is about $\lambda_0/10$

nated from various incidence angles ($0^{\circ} \le \theta \le 45^{\circ}$) for both TE and TM polarizations are shown in Figure 3.12. The results show a second order band pass response centered at 3 GHz with 1.3 dB in-band insertion loss across the scan angle. Both material and conductor losses were included in the simulation. The 3 dB bandwidth is about 15 %. Simulated results show a stable filter response with high out-of-band rejection when the structure is illuminated from $\theta = 0$ to 45 degrees.

For plasma ON state, the electron density value $n_e = 3.6 \times 10^{12} cm^{-3}$ is approximated using plasma shell properties obtained from preliminary work. The simulated transmission and reflection response for various angles of incidence for both TE and TM polarization are shown in Figure 3.13. The structure provides robust average effective shielding of over 25 dB for both in band and out of band operation.



Figure 3.12. Simulated second order band pass filter response in OFF state at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.



Figure 3.13. Simulated second order filter response in ON state ($n_e = 3.6 \times 10^{12} \text{ cm}^{-3}$) at various oblique angle of incident wave. (a) TE polarization; (b) TM polarization.

3.4.2 Fabrication and Experimental Results

An array of 6 x 6 elements is fabricated using FR4 dielectric substrates. These dielectric slabs have relative permittivity value of $\varepsilon_r = 4.4$ and loss tangent of tan $\delta = 0.017$. The finite dimension of the FSS is 63 x 63 mm (equivalent to $0.6\lambda_0 \times 0.6\lambda_0$). Plasma-shells



Figure 3.14. (a) Assembled two-pole FSS and Plasma-shell primed with conductive electrode on the top side of the shells. (b) Photo showing the tunable FSS at ON state



Figure 3.15. sketch and photograph of the flanges spacers and waveguide measurement set up.

are carefully bonded to the metallic layers with 0.5 mm thickness low resistivity epoxy. The biasing network is sandwiched between both outer FSS layers using a fine alignment method. The assembled board along with plasma shells primed with electrode are shown in Figure 3.14(a) and the excited device is shown in Figure 3.14(b).

A waveguide experiment is adopted in this case by enabling a controlled measurement environment and overall cost reduction. A sketch and photograph of the waveguide measurement set up is shown in Figure 3.15. The cut off frequency of the dominant mode (TE_{10} or TE_{01}) is 2.4 GHz which is below our operating frequency. For a fair comparison, a waveguide simulation model of the finite array element is evaluated prior


Figure 3.16. Waveguide simulated result of the finite size second order band pass filter response at (a) OFF state $n_e = 0$; (b) ON state $n_e = 3.6 \times 10^{12} \text{ cm}^{-3}$ equivalent to $\omega_p = 1.07 \times 10^{11} \text{ rad/s}$

to the measurement test. Wave ports are used to transmit and receive the electromagnetic wave. The four boundaries are assigned as perfect conductor to relax the computational time. The simulated transmission and reflection coefficient of the wave propagation at dominant mode is shown in Figure 3.16 for both OFF and ON state ($n_e = 3.6 \times 10^{12} \text{ cm}^{-3}$). The results show a second order band pass response centered at 3 GHz with 1.4 dB in-band insertion loss at OFF state. Also, an average effective shielding of about 25 dB for both in band and out of band operation is observed at ON state. The fabricated prototype perfectly covers the entire aperture of an S-band waveguide to prevent the energy to leak near the walls of the waveguide. Via holes are also added around the perimeter of the FSS to electrically connect the waveguide to the metallic spacer thereby allowing proper current flow along the waveguide walls. Prior to the measurement, R&S ZVB20 network analyzer is accurately calibrated in the range of operation using a full two-port calibration kit. In order to eliminate ripples, the normalized response is transformed to the time domain and an 8-10 ns gating window is applied to the transmission and reflection coefficient. Then, the gated data is transformed back to the frequency domain. The waveguide measurement results of the structure for plasma OFF state are shown in Figure 3.17(a). The filter exhibits a band pass response around 3.15 GHz with 2 dB insertion loss. The 3 dB bandwidth is



Figure 3.17. Waveguide measurement result of the prototype at ON state. (a) OFF state $V_P = 0$; (b) ON state $V_P = 600,700$ V peak voltage

about 12 %. The slight shift of the center frequency may be due to fabrication tolerance, imperfect alignment of the three boards, placement error of the plasma shell, and flatness of the final board. The small increase of the insertion loss can also be due to the lossy epoxy which was not included in the simulation model. Nevertheless, measurement and simulation results are in good agreement.

In order to switch the transfer function of the filter, the shells are biased with the lowfrequency (1 MHz) square wave with 600 V peak and 700 V peak voltage. For plasma ON state, the electric field intensity is coupled through the conductive electrodes and excites the confined gas to form a plasma. As a result the shells light up (Figure 3.14(b)). The plasma activation response time is about 20 to 100 ns. When measured inside the waveguide, switchable average attenuation of about 25 dB and 30 dB is achieved in the passband (Figure 3.17(b)) for both biasing voltage. The measurement results prove to be in very good agreement with simulation results. Characteristics of this alternative design reflect compact low profile, insensitivity to polarization, tunable frequency-selective protection with stable filter response. However it can be seen that the switchable attenuation level is compromised by the overall reduction thickness of the device.

3.4.3 Discussion

When the device is excited with 600 V peak voltage square wave, simulation and measurement follow the same trend. The total power used to excite the second order FSS is estimated about 150 watts. The power per unit volume for noble gas plasma dominated by two-body recombination can be approximated using equation in [55]:

$$\frac{P}{V} = kn_e^2 E_i \approx k \left(\frac{\omega_p}{5.64 \times 10^4}\right)^4 E_i,\tag{3.7}$$

where $k [cm^3/s]$ is the two-body dissociative rate constant, $E_i [joules]$ is the energy required to ionize a noble gas, and $V [cm_3]$ is the plasma volume. The value of $k = 1.8 \times 10^{-7} cm^3/s$ and $E_i = 36.2eV = 5.8 \times 10^{-18} joules$ can be found in [55]. The volume of each plasma volume is about $V = 0.05 cm^3$. Thus by using the electron density value ($n_e = 3.6 \times 10^{12} cm^{-3}$) estimated in the simulation, the power per unit shell-plasma is calculated to be 0.6765 watt/unit. Since 72 shells are used for the first experiment assembled board, the total power absorbed by the device is found to be 48.7 watts. There is about 33 % efficiency used of the RF power. The power loss can be due to the mismatch of the biasing network. A proper network system can help increase the plasma density and improve the degree of protection that the device can provide.

Besides using a matching biasing network to increase the switchable protection level of the device, other variables can be optimized to improve the device performance. In this study, the design was carried out by taking into account available shell's size and gas properties. Different sized plasma shells, gas mixture, and gas pressure give flexibility design in order to further increase attenuation capability of the device. As an example, plasma from low pressure noble gas mixture produce high conductivity at lower frequencies. It can be seen from Equation (3.2), for a fixed electron density value, a lower electron collision frequency (or gas pressure) increases the conductivity of the plasma thus can improve the shielding capability of the device. Additionally, instead of the FR4 dielectric substrate, alumina material would be preferred because they can operate at extremely high temperature environment, and have a better capability to handle high power microwave and electromagnetic pulse. Alumina dielectric also has lower dielectric loss and can potentially reduce insertion loss of the device at OFF state.

3.5 Chapter Conclusion

In this Chapter, we demonstrated improved protection for RF/microwave systems using switchable EM field blocking FSS based on plasma shells. The proposed design delivers a very low profile solution where the total thickness of two-pole bandpass filter is around 10 mm corresponding to $\lambda_0/10$. Due to its small periodicity and overall thickness of the structure (both in the order of $\lambda_0/10$), the filter response is less sensitive to the wave impinging from oblique angles. Attenuation level in both in-band and out-of-band has reached at least 30 dB. The controllability of the structure thickness is achieved by inserting coupling layers between resonator layers. The proposed design can be further expanded to a higher order to increase the attenuation level using the same technique design.

Chapter 4

Plasma-Enabled Adaptable Low-Profile Absorber for Harsh and Dynamic EM/RF Environments

4.1 Chapter Introduction

For target detection and tracking, the frequency agile radar (FAR) system performs multifunctional measurement by sending signals through different carrier frequencies [66, 67]. In order to account for the operating frequency shift, wideband tunable/adaptive radar absorbing materials are crucial in avoiding battlefield surveillance. Nowadays, many existing tunable radar absorbers achieve variable impedance at their input by integrating tuning elements such as PIN diodes, semiconductor and ferroelectric varactors, MEMS switches, liquid-crystal polymers, graphene and liquid metal [68–77]. However, intrinsic characteristics of these tunable components have repercussions on one or more device performance factors, such as: tuning range or speed, reliability, linearity, polarization sensitivity, cost, weight, and fabrication complexity. In terms of reliability, the aforementioned tunable absorbers are vulnerable when operated in extreme environmental conditions including high temperatures and EM radiation from HPME tracking radar systems. This is because such tuning components have limited power handling capability. Moreover, the choice of lossy material is important when the thermal stress due to time average Ohmic losses is of primary concern. In a case where the temperature exceeds the combustion level of the lossy material, burning will occur and the failure in the system will increase the probability of detection.

Our goal is to investigate the feasibility of devising large scale tunable / adaptive

absorbers that can maintain stable operation in harsh and dynamic electromagnetic battlefield surveillance subject to high power microwave energy. The development of compact, lightweight, and high-performance electromagnetic absorbers is vital for stealth applications against high power radar tracking systems. We propose a low-profile tunable absorber based on discrete plasma-shells that provide a wide dynamic range of absorption spectral band. The use of discrete plasma-shells allows the deployment of large-scale prototypes in which the properties of the confined gas can be easily manipulated. For demonstration purpose, the proposed tunable absorbers are deigned to meet the following requirements:

- excellent in-band absorption greater than 90% (i.e. at least 10 dB absorptivity)
- real-time tunability of absorption spectral band that can dependably perform under harsh EM environments, stable absorption response over wide incident angle (> ± 45 degrees)
- low-profile design (« λ/4 separation between adjacent FSS layers), and polarization independent (this feature is best suited for randomly moving and/or rotating systems as it minimizes polarization mismatch loss)
- capable of withstanding ionizing radiation, shock, vibration and temperature range up to 250°C
- ability to integrate into existing structures to meet the needs of a broad range of Air Force applications.

This chapter is organized into three sections elaborated as follows:

The first section covers the feasibility of devising a low-profile, practical, large scale, electronically tunable/adaptive absorber based on plasma technology. We demonstrate

the concept by integrating discrete plasma-shells with a well-designed metallic backed absorber. The control of the absorption spectral band and energy level is allowed by systematically changing the plasma density. The proposed FSS-based absorbers are loaded with different types of resistive sheet (lossy inductive FSS, lossy capacitive FSS, and lossy resonant FSS layer).

In the second section, we investigate potential implementations of the proposed absorber subjected to HPME and HEMP by selecting the core materials that are suitable for higher-temperature service. A magneto-dielectric substrate is incorporated in the design as a substitute for the resistive sheet to account for the loss. This alternate design allows the EM energy to be dissipated in lossy magneto-dielectric substrates. Instead of relying on exotic lossy materials to achieve good performance, our design simply makes use of commercially available lossy magneto-dielectric substrate (from Laird's Eccosorb MF500F series) suitable for high power microwave compared to the resistive sheet used in the prior design. In the first hand, a single pole FSS-based absorber using the magneto-dielectric substrate is investigated. In the other hand, we expand the single pole FSS-based absorber to a higher order spatial FSS based absorber to increase the absorption bandwidth. The proposed multilayer absorber is based on a technique developed for the design of a thickness customizable high-order bandpass frequency selective surface. Such technique allows simple integration of the tuning elements, while simultaneously providing the design option to realize the absorber with specific or desired thicknesses. These multi-layered absorbers with potential integration of tuning components enable tunable/adaptable operation modes by controlling the plasma density at each FSS layer.

And finally, in the last section, the performance of two-pole HPM absorber is investigated under high power excitation. The maximum power level handling capability of the device will be numerically predicted and validated experimentally.



Figure 4.1. Unit cell representation of the dynamic plasma-based absorber along with the detailed geometry (units are in mm). (a) Design #1 indicates the absorber with the inductive lossy layer. (b) Design #2 is the one with the capacitive lossy layer. (c) Design #3 is the one with lossy resonator layer.

4.2 Plasma-Based Tunable Absorber Loaded with Resistive Sheet

4.2.1 Design Specification and Model Analysis

Unlike our previous designs where these plasma-shells' electrical properties were manipulated and used to design spatial filter switches, in this project, they serve as tuning components of reconfigurable absorbers. The topology of the proposed design is based on a resistive metamaterial sheet absorber. A lossy inductive, capacitive or resonator layer, which is embedded with discrete plasma-shells, and a practical biasing layer (as shown in Figure 4.1) is investigated. The lossy layer is made of resistive film with specific surface resistivity $R_S[\Omega/Sq]$. The pattern of the metallic biasing layer combines a square patch array with a wire grid. In the proposed designs, the lossy layer and biasing network are printed on RO4003C, a hydrocarbon ceramic laminate from Rogers Corporation's RO4000 series materials, which are known for their reliability when subject to severe thermal shocks. The thickness of the substrate is $h_1 = 0.813$ mm and relative permittivity is $\varepsilon_{r1} = 3.38$ with ± 0.05 tolerance. The plasma shells (shell size: 4 mm × 4 mm × 3 mm), placed between the ground plane and the biasing layer, are used not only as the tuning component, but also for structural support. The shells are hollow ceramic dielectric (Al_2O_3) material with wall thickness of 0.2 mm and dielectric constant of 9.6. The shells are bonded to the metallic layers with 0.5 mm thickness of silver epoxy (not shown in the figure) used as electrodes to facilitate the coupling of energy through the dielectric wall of the shell.



Figure 4.2. Equivalent circuit model of the tunable absorbers for normal angle of incidence.

In this proposed active absorbers, the metallic resonators are used as practical biasing network. As a result, the effective relative dielectric constant loading the FSS capacitors can be controlled to add tunability to the system. To elucidate the working principle of the absorption and its frequency tuning capability, a transmission line model approach is adopted. The ECM of the design is shown in Figure 4.2. The biasing layer is modeled by a parallel L_b , C_b resonant tank and the bulk plasma as a shunt component with admittance Y_p . The biasing network and the ground plane are separated by a transmission line with length $h_0 = 3$ mm (height of the shell) and characteristic admittance Y_0 (characteristic admittance of free space). The lossy layer is also modeled as a shunt admittance Y_L , which is separated from the biasing network by a short transmission line of length h_1 and characteristic admittance $Y_1 = Y_0\sqrt{\varepsilon_{r1}}$. For the first topology, the lossy layer is inductive and is seen as a series R_I , L_I , and for the second topology it is capacitive, thus seen as

a series R_C , C_C . on the other hand, for the third design, the lossy layer is a resonant component seen as a series R_r , L_r , C_r . From the ECM, the input admittance of the circuit can be obtained as follows:

$$Y_{in} = Y_L + Y_1 \frac{(Y_{bias} + Y_p + Y_d) + jY_1 \tan(\beta_1 h_1)}{Y_1 + j(Y_{bias} + Y_p + Y_d) \tan(\beta_1 h_1)},$$
(4.1)

$$Y_d = -jY_0 \cot(\beta_0 h_0); \ Y_{bias} = j(\omega C_b - \frac{1}{\omega L_b}), \tag{4.2}$$

$$Y_{L} = \begin{cases} (R_{I} + j\omega L_{I})^{-1} & \text{for the lossy inductive sheet} \\ (R_{C} - \frac{j}{\omega C_{C}})^{-1} & \text{for the lossy capaciticesheet} \\ (R_{r} + j\omega L_{r} - \frac{j}{\omega C_{r}})^{-1} & \text{for the lossy resonator sheet} \end{cases}$$
(4.3)

The bulk plasma is modeled based on parallel plane geometry and because equal electrode area is used at each side of the shell, a symmetric discharge is adopted. As we mentioned in the previous chapter, a plasma medium can be treated as a conductor with conductivity σ_p or as a dielectric with permittivity ε_p based on its electron collision frequency, its plasma frequency and the operating frequency of the system [62]. The first scenario for frequencies where $\omega \ll \omega_p, \nu$, the conductivity of the plasma is then reduced to its dc form such that $\sigma_p \rightarrow \sigma_{dc} = \frac{\omega_p^2 \cdot \varepsilon_0}{\nu}$. Such characteristics of the plasma have been used to design switchable frequency selective surfaces for electronic protection in the Chapter 2. For this current chapter, the second scenario is considered where the bulk plasma is used as a dielectric with variable permittivity such that we have $\nu < \omega < \omega_p$ which is practical for plasma driven at RF discharge.

The admittance Y_p of the bulk plasma is given in [62] as:

$$Y_p = \frac{j\omega\varepsilon_p A}{h_0},\tag{4.4}$$

where A is the effective cross-sectional area of the plasma. By assuming uniform electron density, this expression is approximated to:

$$Y_p = j\omega C_0 + (j\omega L_p + R_p)^{-1}.$$
(4.5)

where $C_0 = \varepsilon_0 A / h_0$ is the capacitance of the vacuum chamber, $L_p = \omega_p^{-2} C_0^{-1}$ is the inductance of the plasma, and $R_P = \nu L_p$ is the plasma resistance. As shown in Figure 4.2, the shunt component with impedance Y_p is modeled as a series (L_p, R_p) in parallel with C_0 . Theoretically, the resonant frequency ω_0 of the absorber is obtained when $Im[Y_{in}(\omega_0)] = 0$, with a matching condition such that $Re[Y_{in}(\omega_0)] = Y_0$. Thus, the operating frequency of the absorber can be simply tuned by adjusting its input impedance. This task can be achieved with the variation of the bulk plasma admittance Y_p . At OFF state ($\omega_p = 0$), the series $(L_p R_p)$ circuit is opened and the bulk plasma model is reduced to the shunt capacitance C_0 . At ON state ($\omega_p = 0$), the admittance Y_p of the plasma discharge is a parallel combination of the shunt capacitance C_0 with the series $(L_p R_P)$ circuit. As the plasma frequency ω_p increases, the values of the inductance and resistance decrease. Therefore the effective inductance of the system decreases leading to the tuning of the resonant absorption frequency as it shifts to higher frequency. In fact, from Equation (4.5), it can be perceived that for fixed electron collision frequency (gas content and pressure), the impedance of the plasma can be controlled by either its plasma frequency or its cross-sectional area as will be conveyed in the next section.

4.2.2 Implementation and Simulation Results

The tunable absorbers are designed to operate in the X-band (8 GHz – 12 GHz) as this specific band is commonly used for radar and military electronic warfare applications, as well as satellite, terrestrial and space-based communication. A full wave EM analysis

Phys. para.	<i>p</i> (<i>mm</i>)	<i>a</i> (<i>mm</i>)	<i>b</i> (<i>mm</i>)	ω (mm)	g (mm)	s (mm)	$R_{S}\left(\Omega/Sq.\right)$
Design #1	13	11		9.55	0.25		110
Design #2	13	11	•••	7.9	0.25	•••	100
Design #3	15	10.9	12.6	9.35	0.25	1	9.3
Elec. para.	$L_b(nH)$	$C_b(pF)$	$L_{I}(nH)$	$C_c(pF)$	$L_r(nH)$	$C_r(pF)$	$R(\Omega)$
Elec. para. Design #1	$\begin{array}{c} L_b (nH) \\ 2.3 \end{array}$	$\begin{array}{c} C_b \ (pF) \\ 0.2 \end{array}$	$\begin{array}{c} L_I(nH) \\ 1.6 \end{array}$	$\begin{array}{c} C_c \ (pF) \\ \cdots \end{array}$	$L_r(nH)$	$\frac{C_r(pF)}{\cdots}$	R (Ω) 320
Elec. para. Design #1 Design #2		$ \begin{array}{c} C_b \ (pF) \\ 0.2 \\ 0.08 \end{array} $	$ \begin{array}{c} L_I(nH) \\ 1.6 \\ \dots\end{array} $	$\begin{array}{c} C_c \ (pF) \\ \dots \\ 0.1 \end{array}$	$ \begin{array}{c} L_r(nH)\\ \dots\\ \dots\\ \end{array} $	$\frac{C_r(pF)}{\cdots}$	R (Ω) 320 146

 Table 4.1. Physical/electrical parameters of the proposed absorber.

using ANSYS HFSS is utilized to characterize the response of the structures at OFF state as shown in Figure 4.3. Based on the physical parameters and the full wave frequency response of the absorber without the plasma-shells, the electrical parameters of the ECM are extracted in conjunction with a circuit simulator (Keysight's Advanced Design System).



Figure 4.3. Simulated FW simulation results for the reflectivity of the passive absorbers along with the ECM results.

First, the values of the parallel L_b , C_b resonant tank are obtained using simple analytical expressions for wire grid and patch array [17]. Next, only the lossy layers with the RO4003C are simulated using full wave EM simulation and then the responses are matched to those

obtained from their counterpart ECM comprised of a series R_I , L_I or R_C , C_C or R_r , L_r , C_r followed by the short transmission line of length h_1 using a curve fitting technique. The value of the capacitance that accounts for the dielectric constant shells C_0 is numerically obtained from the full wave simulation of the design (in Figure 4.1) with and without the plasma shells. Finally, by linking and slightly adjusting all the electrical parameters obtained, the frequency responses of the complete ECM in Figure 4.2 are simulated for the OFF state case. The physical and electrical parameters of the absorbers are provided in Table 4.1. Both the full wave EM results and the ECM results are illustrated in Figure 4.3 and the comparison reveals the accuracy of the transmission line modeling. The total thickness of each absorber is about 3.813 mm, corresponding to $0.11\lambda_0$, where λ_0 is the free space wavelength of the absorption center frequency ($f_0 = 9 GHz$). The results of the passive absorber predict about 18%, 24% and 30% fractional bandwidth at 10 dB return loss level for the lossy inductive, capacitive and resonant FSS layers design type, respectively. The bandwidth over thickness performance of the proposed absorbers is computed and compared with a Salisbury screen (operating at the same center frequency) by examining their respective figure of merit (FoM), as depicted in Table 4.2. The FoM is found by using the Rozanov [78] physical bound imposed on the thickness to bandwidth ratio for any nonmagnetic metal backed absorber as:

$$FoM = \frac{|\ln(\rho_0)|(\lambda_{max} - \lambda_{min})}{2\pi^2 d},$$
(4.6)

where ρ_0 is the reflectivity, *d* is the total thickness of the absorber, and λ_{min} and λ_{max} are the respective minimum and maximum wavelength in the spectrum range allowed by the specific reflection coefficient. As given in Equation (4.6), a higher FoM indicates the design with superior performance. Table 4.2 shows that beyond 20-dB absorption level (which is typically preferred in most application), wider absorption bandwidth is obtained for

Reflectivity	$\rho_0 = 0.1(10 \ dB)$	$\rho_0 = 0.01(20 \ dB)$	$ \rho_0 = 0.001(30 \ dB) $
Salisbury	0.4064	0.2390	0.1123
Design #1	0.1851	0.1013	0.0473
Design #2	0.2483	0.1470	0.0677
Design #3	0.3130	0.3605	0.3115

 Table 4.2. Comparison in FoM between the proposed absorbers and Salisbury screen.

Design #3. Although Design #1 and Design #2 have similar performance compared to the Salisbury screen, their active counterparts will have superior overall characteristics.



Figure 4.4. Simulated results of the plasma-based absorber (Design #1) subject to different plasma frequencies. (a) Simulated ECM tuning response. (b) Simulated full wave (FW) tuning results.

At ON state, the admittance of the bulk plasma can be engineered to provide the tunability capability to the design. The electron collision frequency of the plasma is set to its optimal value $v = 3.6 \times 10^{10} rad/s$. Then, the response of the absorbers is examined by changing the plasma frequency within the range of 0 (OFF state) to $8 \times 10^{11} rad/s$. Both the simulated ECM and the full wave (FW) EM results for all three designs are illustrated in Figure 4.4, Figure 4.5 and Figure 4.6, respectively . By comparing the frequency response of the absorber using full wave EM simulation to those obtained from the ECM, it is observed that the ECM accurately predicts the behavior of the absorber at ON state. The results show that the absorption center frequency increases from 9 GHz to 10 GHz, providing



Figure 4.5. Simulated results of the plasma-based absorber (Design #2) subject to different plasma frequencies. (a) Simulated ECM tuning response. (b) Simulated full wave (FW) tuning results.



Figure 4.6. Simulated results of the plasma-based absorber (Design #3) subject to different plasma frequencies. (a) Simulated ECM tuning response. (b) Simulated full wave (FW) tuning results.

a dynamic tuning of the absorption spectral band for design #1 and design #2. Instead, for Design #3, the increase of the plasma frequency mostly affects the higher resonance frequency, thus providing a tuning of the absorption spectral bandwidth but at the expense of the absorption rate.

The tuning range of the absorber can be further improved by increasing the crosssectional area of the shell. The size of plasma-shell that can be manufactured ranges from 0.5 mm to 10 mm. The response of the absorber is further explored using larger plasmashell size with dimensions 8 mm × 8 mm × 3 mm such that the thickness of the design remains unchanged (i.e. only the cross-sectional area of the unit cell shell is increased). By changing the plasma frequency, the center frequency of the absorber (design #2) can be



Figure 4.7. Simulated full wave (FW) results of the plasma-based absorber subject to different plasma frequencies and larger shell size. Tuning capability observed for Design #2.



Figure 4.8. Fabricated sample. (a) Front side: lossy sheet. (b) Back side: biasing network. (c) Assembled tunable absorber: side view.

increased with the size of the plasma shells from 9 GHz to 11 GHz (see Figure 4.7), which is about double the tuning range previously obtained. Also the absorption bandwidth for Design #3 has increased considerably under the same tuning procedure.

4.2.3 Fabrication and Measurement Results

To validate the numerical results, one of the prototype circuits (Design #2) is fabricated on 0.813 mm thick RO4003C substrate. An array of lossy capacitive patches made of $R_{CS} = 100\Omega/Sq$ OhmegaPly resistive film (manufactured by Ohmega Technologies, Inc.) and the metallic biasing layer are laminated and patterned on either side of the substrate using wet etching process. The board is made by Brigitflex Inc., a circuit board manufacturer

known for their expertise in the fabrication of customized PCBs embedded with planar resistors. A square panel of about 5" × 5" (equivalent to $4\lambda_0 \times 4\lambda_0$) including an array of 10 × 10 elements is used as the test sample. The front and back of the circuit are shown in Figure 4.8. The plasma-shells (shell size: 4 mm × 4 mm × 3 mm) filled with 0.1 % Argon - 99.9 % Neon mixture at 175 Torr gas pressure are commercially available from Imaging Systems Technology Inc. The value of the gas pressure (P = 175 Torr) is estimated by solving a simple theoretical expression for neon gas derived in [65] at 300 K electron temperature:

$$\nu[s^{-1}] = 8.63 \times 10^{-18} \times \left(\frac{P}{K_B T_e}\right) \times T_e^{0.833} \text{ for } 10^2 K \le T_e \le 5 \times 10^3 K, \tag{4.7}$$

where K_B [J/K], T_e [K], and P [Pa] are Boltzmann's constant, electron temperature and the gas pressure. The plasma-shells are electrically bonded to the metallic layers with 0.5 mm thickness of silver conductive epoxy. A frame with an array of cavities fabricated using a 3D printer (uPrint SE Plus) is used for fine alignment. The silver conductive epoxy is deposited on both the biasing and ground plane using a syringe. Then the shells are manually placed in the cavities. The fabrication process is illustrated in Figure 4.9. The assembled board including the plasma shells and ground plane is shown in Figure 4.8(c).

The device is excited with a sine wave produced by Keysight N5181A MXG RF analog signal generator operating in the range of 100 KHz – 6 GHz. The wave from the signal generator is amplified with a 50 dB (nominal) CW ENI/E&I 325LA power amplifier (PA), with driving frequency ranging from 250 KHz to 150 MHz and maximum input voltage limited to 1 V. The output voltage of the PA is then connected to the bias traces of the absorber. If desired, for any specific plasma frequency used for each tuning state, the power required to sustain the plasma-shells can be estimated using the approximated



Figure 4.9. Fabrication process. (a) Conductive silver epoxy deposited on ground plane using a syringe. (b) Shells manually placed in holes of 3D printed frame. (c) Conductive silver epoxy also deposited on top side of shells before assembly.

power per unit volume for noble gas plasma dominated by two-body recombination [55]:

$$\frac{P}{V} = kn_e^2 E_i \approx k \left(\frac{\omega_p}{5.64 \times 10^4}\right)^4 E_i,\tag{4.8}$$

where $k [cm^3/s]$ is the two-body dissociative rate constant, $E_i [joules]$ is the energy required to ionize a noble gas, and $V [cm_3]$ is the plasma volume. The value of $k = 1.8 \times 10^{-7} cm^3/s$ and $E_i = 36.2eV = 5.8 \times 10^{-18} joules$ can be found in [55]. Because 100 plasma-shells (shell size: 4 mm × 4 mm × 3 mm) are used for the fabricated prototype, the total plasma volume is about $V = 4.8 cm^3$. Therefore, under perfect conditions, for any specific plasma frequency ω_p , the CW power P [W] required to sustain the plasma volume can be estimated using Equation (4.8). Then the voltage V_P can be theoretically obtained by taking into account the 50 dB gain of the power amplifier. However, due to the complexity of the nature of charged particles in the plasma medium, the power loss due to the mismatch of the biasing network, the choice of the driven RF frequency, and other factors including fabrication errors, the theoretical value of P has not been correlated with the experimental value. Instead, the absorber tuning states given the driven RF frequency and the experimental voltage V_P from the signal generator are empirically approximated.



Figure 4.10. Dimming effect of the shells when plasma discharge is driven with 1 MHz RF signal. By increasing the voltage from the signal generator a glow discharge is obtained as a sign of avalanche discharge. (a) $V_P = 0.5$ V. (b) $V_P = 0.7$ V. (c) $V_P = 0.9$ V.

By setting the driven frequency (f_S) at 1 MHz and increasing the voltage from the signal generator, the glowing effect of the plasma-shells can be seen in Figure 4.10. An electric glow discharge visible to the eye is produced. When $V_P = 0.7 V$, the power output into the load which is monitored by a built-in front panel meter of the PA reads about 50 W (power density of 2.9 KW/m^2). The sustained continuous wave output power generated by the power supply is capable to ionize the large-scale plasma-device and eliminate the need of laser excitation or high pulsed voltage supply which are relatively costly.



Figure 4.11. Sketch and photograph of the test setup (free space) for the absorber measurement.

The experimental setup to validate the device performance is illustrated in Figure 4.11. A pair of PE9887-11 broadband horn antennas are placed about 0.5 m from the device under test (DUT) to ensure the DUT is at far-field from the test antennas. An aperture of the size of the absorber is carved out at the center of a large wood frame to mount the test sample.



Figure 4.12. Full wave simulated reflection coefficient of the passive absorber (Design #2) at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.



Figure 4.13. Measured reflection coefficient of the passive absorber (Design #2) at various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.

The frame is covered with an adhesive surface wave absorber (MR-31-0003-20 from Mast Technologies) to reduce diffraction effects from the edges of the DUT. The calibration of the system is carried out by measuring the reflection coefficient of the system when the aperture is covered by a conductor plane. Then the reflection coefficient of the system is obtained with the device under test. These results are obtained with a calibrated Agilent HP 8719ES (50 MHZ to 13.5 GHz) vector network analyzer. Gating window is applied to reduce ripples in the measurement caused by multiple reflections of the waves and also to reduce cross coupling between the horn antennas. The frequency response of the absorber is obtained by normalizing the measured results obtained in the presence of the device with the one obtained from the conductor plane of the same size. The performance



Figure 4.14. Measured reflection coefficient of the active absorber (Design #2) at normal angle of incidence for various biasing voltages V_P . (a) Driving frequency $f_S = 1$ MHz. (b) Driving frequency $f_S = 2$ MHz.

of the absorber (at OFF state) across different angles of incidence for both TE and TM polarizations is numerically and experimentally obtained. Both simulated and measured results (Figure 4.12 and Figure 4.13) show acceptable and stable reflection response when the system is illuminated from various oblique angles ($0^{\circ} \le \theta \le 45^{\circ}$). At normal incidence, we can notice a slight deviation of the center frequency from 9 GHz to 9.2 GHz of the measured results compared to the ones obtained from full wave EM simulations. The slight discrepancies can be due to many factors including the finite size of the prototype, the thickness and homogeneity of the silver epoxy used, fabrication tolerance (caused by inaccurate material properties) and imperfect assembly of the absorber.

The reflection coefficients of the device at ON state are obtained when the plasma is sustained using the amplified RF source directly coupled to the gas across the thin ceramic dielectric wall. The driven frequency (f_S) from the signal generator is set to 1 MHz and a voltage range between $0 \le V_P \le 0.9V$ is fed to the PA since its maximum allowed input voltage is 1 V. The measured results at normal angle of incidence indicate a tuning of the absorption band in the X-band as shown in Figure 4.14(a). The absorption resonant frequency has increased from 9.2 to 9.9 GHz and wider absorption band is obtained by increasing the biasing voltage. The tuning speed is found to be as short as 10-100 ns.

Experimentally, it has also been previously demonstrated that the percentage of RF power absorbed by the electrons increases with the driven frequency [79]. This is theoretically explained using the equation from homogeneous plasma model derived by Godyak [80] as:

$$\frac{P_i}{P_e} = \left(\frac{\omega_p}{\omega_s}\right)^2 \frac{3u_B}{2d\nu'},\tag{4.9}$$

where P_i is the power absorbed by ions, P_e is the electron heating power, ω_s is the driving frequency, d is the plasma half-width, and $u_B = (eT_e/M)^{1/2}$ is the Bohm velocity (M is the ion mass). Another set of results is further obtained by increasing the driven frequency (f_S) to 2 MHz. As can be seen in Figure 4.14(b), the frequency shift starts for lower biasing voltage at 0.3 V. By increasing the voltage from 0-0.8 V, the plasma density increases with higher driven frequency. The absorption center frequency shifts until it ceases to increase at around 10.5 GHz. When the voltage is near 0.8 V, the electron density remains virtually unchanged, which is independent of the voltage.



Figure 4.15. Power measurement setup using a high power dual directional coupler and a spectrum analyzer.

In addition, power consumption of the plasma-shell tuned absorber is estimated by

$\mathbf{V}(\mathbf{V})$	$D_{\rm c}(M)$	$P_r(W)$		$P_L(W)$		$P_u(W/cm^3)$	
$v_p(v)$	$P_f(\mathbf{v}\mathbf{v})$	1 MHz	2 MHz	1MHz	2 MHz	1 MHz	2 MHz
0.3	44.6	26.3	24.4	18.3	20.2	3.81	4.21
0.4	45.7	26.9	25.0	18.8	20.7	3.92	4.31
0.5	47.8	27.9	25.5	19.9	22.3	4.15	4.65
0.6	48.7	28.6	26.2	20.1	22.5	4.19	4.69
0.7	50.6	29.8	27.8	20.8	22.8	4.33	4.75
0.8	52.8	31.1	29.8	21.7	23.0	4.52	4.79
0.9	54.4	33.4	32.3	22.0	23.1	4.58	4.81

Table 4.3. Summary of the power distribution at different driven frequencies.

measuring the forward power (P_f) and reflected power (P_r) to and from the plasma-shells, respectively. A dual directional coupler with two auxiliary outputs is used to sample both forward and reflected energy. A 40 dB high power dual directional coupler (C40-110-481/1N by Pulsar Microwave Corp.) is used to characterize the power consumption level of our absorber. The test setups (in Figure 4.15) show the output of the reverse coupling port reduced by the 20 dB attenuator and detected by a spectrum analyzer, while the output of the forward coupling port is terminated by the 50 Ω load. Based on the coupling factor and the attenuation level, the sampled data is used to obtain the reflected power from the load (or absorber). Similarly, by interchanging the role of the forward and reverse coupling port, the forward power is obtained. A summary of the device power consumption analysis for the different driving frequencies is provided in Table 4.3, where P_L is the total power absorbed by the plasma volume and P_u is the power per unit volume. As can be observed from this fidelity test, the percentage of RF power absorbed by the plasma increases with the driven frequency. On average, the current biasing setup shows about 42 % efficiency, which can be improved if needed by adding a matching network.

External matching networks between the load (tunable absorber) and the power supply can be designed in order to increase the efficiency of the RF power transfer from the source

Absorber Tun-	Resonant Frequency	Tuning speed	Thermal ef-	Additional notes
ing Mechanism	Tuning range	in the order of	fect capability	
Varactor diodes [69]	4.35 - 5.85 GHz	nanoseconds	< 125°C	Complex biasing net- works; design sensitive to polarization.
MEMS [72]	1.12 - 1.32 THz	milliseconds	N/A	Design sensitive to polar- ization.
Liquid crystal [74]	2.51 - 2.62 THz	milliseconds	< 300°C	Not easy to integrate in many applications; de- sign sensitive to polariza- tion.
Graphene [75]	0.91 - 1.08 THz	milliseconds	< 1500°C	Complex biasing config- uration and not easy to integrate in measure- ment devices. Also, the realization of graphene using chemical vapor de- position (CVD) is costly.
Liquid metal [77]	0.24 - 0.41 THz	milliseconds	< 1300°C	Requires sophisticated control mechanism for synchronous movement of the liquid metal.
Plasma [this work]	9.2 - 10.5 GHz	milliseconds	< 1000°C	Can handle high power at microwave frequen- cies. Simple biasing net- work and easy integra- tion.

Table 4.4. Comparison of our plasma-based absorber with other existing tunable absorbers.

to the plasma discharge and also reduce the need of higher power for activation voltage threshold. However, a variable matching network may be required since the admittance of the bulk plasma is dependent on the discharge voltage. In fact the admittance of the bulk plasma in Equation (4.5) is a function of the plasma frequency or the electron density of the ionized plasma which is electrically controlled by the applied voltage. All limitations considered, the detailed design and analysis of the biasing network are out of the scope of the material presented here and left for future research items.

Although large-scale plasma-based devices may require a relatively larger amount of power than other electronic tuning mechanisms, this dilemma can be overcome using

a high efficiency power amplifier and more matching circuitry. The advantages of our design using plasma technology are highlighted in Table 4.4. Compared with some existing tunable microwave absorbers, superior characteristics of plasma-based devices such as simple biasing schemes, ability to integrate into existing structures to reduce system weight, cost, and most importantly survivability in harsh EM/RF environments (including extreme temperatures and ionizing radiation) outweigh the low power consumption requirement.

4.3 Plasma-Based Tunable Absorber Loaded with Magneto Dielectric Substrate

4.3.1 Single Pole Plasma-Based Tunable Aborber

4.3.1.1 Design Specification

Let us recall that the basis of this project is the investigation of high power handling capability of microwave absorbers that can rapidly adapt to harsh and dynamic EM environments. Although the tuning component (plasma-shells) used in the proposed tunable absorber has the unique property to withstand high power microwave energy [57, 58], the other core material in the design is not suitable for high-power operation. The proposed absorbers, based on lossy (resistive) layers, may be useful for applications that require moderate power handling and operation in dynamic EM environments. However, although they are relatively low cost and simpler to realize, the lossy resistive layers used to absorb the EM wave have very limited thermal stress capability. Usually, these thin resistive sheets, made of nickel phosphorous, nickel chromium, nickel chromium aluminum silicon, or chromium silicon monoxide (foils available on Rogers substrates from Ticer and Ohmega Technologies) operate in relatively lower temperature ranges but

provide superior absorption bandwidth because of their frequency-independent electrical properties over a broader frequency range. Above their maximum operation temperature, these resistive films are susceptible to burn.



Figure 4.16. (a) Topology of the single pole high power absorber; (b) Modified single pole high power absorber due to practical issues; (c) The physical parameters used to obtain absorption in the C-band are provided: a = 12.7; $g_1 = 0.5$; $s_1 = 1.3$; $b_1 = 9.2$; $l_1 = 6.5$ (units in mm).

To address this vulnerability, lossy substrates that can withstand higher power EM waves can be used to improve the peak power handling capability of the absorber. In order to account for the total loss provided by the resistive layer in the previous designs, both dielectric and magnetic losses are needed. Unfortunately, such substrate options are limited in today's market in terms of losses and relative power handling capability. Therefore, a systematic design of FSS element shapes with available materials will be required to effectively match the input impedance of the design to the characteristic impedance of free space. In the modified design, a frequency-dispersive lossy magneto-dielectric is used in lieu of the RO4003C used in the previous absorber as shown in Figure 4.16(a). The lossy substrate (Eccosorb MF500F-112 from Laird Technologies) is rigid and can be used for high transient power levels, allowing high temperatures up to 260°C (500°F) as mentioned in the data sheet. In order to show that the absorption band of the proposed absorber can be scalable to a desired microwave spectrum, an electromagnetic bandgap (EBG) resonator is employed for the IEEE C-band (4-8 GHz) operation. The working principle of the single pole high power HPM absorber is similar to the previous absorber

with the resistive sheet, as the plasma-shells are used to tune the surface impedance of the resonator, which results in a dynamic tuning of the absorption spectral band. However, this design presents some practical issues for fabrication. The commercially available magneto-dielectrics do not have printed copper layers on any side that could be used to etch the EBG resonator pattern. Moreover, they are available in standard thickness options (*h* = 3.2, 6.4, 9.5, 12.7, 15.9, 19.1 mm, etc.). Printing the EBG resonator on a thin dielectric and pasting the magneto-dielectric directly on top can also lead to a thick design especially for a single pole operation. One way to overcome aforementioned limitations, is to perforate the Eccosorb MF500F substrate (with thickness h = 3.2 mm) in order to accommodate the discrete shells as shown in Figure 4.16(b). The plasma-shell dimensions used in this model are 6 mm x 6 mm x 3 mm. Since the absorber is operating in C-band, the value of the electron collision frequency used to improve the performance of the system is set to be $\nu = 2.0 \times 10^{10} rad/s$. The magneto-dielectric MF500F-112 is accurately modeled in the frequency domain using its frequency-dependent data points in HFSS. The frequency-dependent definition is applied to both electric and magnetic properties (real part of permittivity and permeability, electric loss tangent, and magnetic loss tangent) of the material. HFSS interpolates this data at the desired frequencies during the generation of the solution.



Figure 4.17. Full wave simulated reflection coefficient of the single pole HPM absorbers at various oblique angles of incident wave (a) for TE polarization; (b) TM polarization.



Figure 4.18. Full wave simulated reflection coefficient of the single pole HPM absorber subjected to different plasma frequencies. Tuning capability observed for the design (a) with shell size: 6 mm x 6 mm x 3 mm; (b) with larger shell size: 8 mm x 8 mm x 3 mm.

Figure 4.17 shows the full wave simulation result of the HPM absorber performance. At normal incidence angle of the EM wave, results predict about 9% FBW at 10 dB reflectivity level. The total thickness of the structure is only 4 mm (about $\lambda_0/12$) and the periodicity in the order of $0.25\lambda_0$, where λ_0 is the free space wavelength at 6 GHz. The performance of the absorber is also investigated across different angles of incidence for both TE and TM polarizations as shown in Figure 4.17. Simulation results predict stable absorption response and at least 20 dB absorption across the absorption band for both TE and TM polarizations when the system is illuminated from various oblique angles ($0^\circ \le \theta \le 45^\circ$). The effect of the variation of the plasma frequency ($0 \le \omega_p \le 9 \times 10^{11} rad/s$) on the proposed design performance is also investigated. Simulation results investigating different shell sizes (illustrated in Figure 4.18) show that by changing the value of the plasma frequency within that range, the absorption center frequency increases providing real-time/on-demand electronic tunability of the absorption spectral band. It can also be observed that the larger plasma volume increases the tuning range of the single pole HPM absorber.

4.3.2 Multiple Pole Plasma-Based Tunable Aborber

4.3.2.1 Design Specification

The proposed design that incorporates resistive FSS sheets provides broader bandwidth (for the same thickness) compared with the one that incorporates lossy magneto-dielectric substrate , but its operation service is limited to temperatures less than 125°C. On the other hand, the single pole design that uses lossy substrates with higher thermal conductivity is suitable for higher power microwave energy but the operating bandwidth is limited (9% fractional bandwidth at 10 dB reflectivity level) by the dielectric and magnetic losses of the existing MF500F-112 substrate. In order to broaden the absorption bandwidth, a multilayers/multipoles design with the ability to have independent tuning control over each resonant frequency is needed.



Figure 4.19. Topology of the multipole aborbers based on coupling interlayers; (a) Two-pole; (b) Three pole; (c) Detailed geometry of the unit cell of each layer to obtain absorption in the C-band.

The general synthesis developed in Chapter 2 allows simultaneous control of the electromagnetic waves response and design thickness of multi-layered bandpass FSS filters. The technique is based on coupled bandpass filter theory using admittance inverter. The design concept allows the engineer to use readily available commercial dielectric materials



Figure 4.20. Simulated reflection coefficient of the passive multipole HPM absorbers; (a) Two-pole with physical parameters: $h_1 = 3.2$; $h_2 = 0.8$; a = 12.5; $g_1 = g_2 = 0.35$; $s_1 = s_2 = 1.4$; $b_1 = 9.3$; $b_2 = 8.24$; $l_1 = l_2 = 6.5$; r = 4.2; (b) Three pole with physical parameters: $h_1 = 3.2$; $h_2 = 0.8$; a = 12.5; $g_1 = g_2 = g_3 = 0.15$; $s_1 = s_2 = s_3 = 1.6$; $b_1 = 9.2$; $b_2 = 9$; $b_3 = 7.8$; $l_1 = l_2 = l_3 = 6$; r = 4 (units in mm).

with standard thicknesses for multi-layer FSS design technology. Such capabilities also bring practical benefits for tunable metasurface by providing flexibility in integrating tuning elements or materials that require precise control of physical dimensions. The proposed multipole absorber relies on the same design concept. A multilayer (secondorder and third-order) bandpass FSS using such technique is placed above a ground plane as depicted in Figure 4.19. For such combinations, the absorption band is mainly determined by the operation frequency of the FSS and loss can be introduced in the design using either resistive FSS elements or lossy substrates. The latter is taken into account, and lossy substrates, consisting of the Eccosorb MF500F series (with thickness $h_1 = 3.2mm$) are used to absorb the EM wave energy. The FSS is formed by a periodic array of cascaded electromagnetic bandgap (EBG) resonators that behave as a parallel LC bandpass filter. An interlayer consisting of a mesh grid is placed between the resonators to regulate the coupling level between consecutive resonators for various separations. The metallic layers are printed on the FR4 dielectrics with thickness $h_2 = 0.8mm$. The simulated full wave EM results illustrated in Figure 4.20 predict absorption of the EM wave in the C-band centered at 6 GHz with 17% and 25% fractional bandwidth at 10 dB reflectivity level for both second and third order absorber, respectively. The total thickness of the two-pole

Design	Size	h_1	а	$g_1 = g_2$	$S_1 = S_2$	<i>b</i> ₁	<i>b</i> ₂	$l_1 = l_2$	r
#1	$\lambda_0/9$	2	12.5	0.17	1.56	9.3	8.3	6.5	3.6
#2	$\lambda_0/7$	3	12.5	0.25	1.50	9.3	8.2	6.5	3.9
#3	$\lambda_0/5$	4	12.5	0.35	1.47	9.3	8.2	6.5	4.2

Table 4.5. Physical dimensions (in mm) and electrical size of the three passive absorbers.

absorber is 8 mm (about $\lambda_0/6$) and the periodicity in the order of $0.25\lambda_0$, where λ_0 is the free space wavelength at 6 GHz. For the three-pole absorber the total thickness is 12 mm (about $\lambda_0/4$) and the same periodicity. Other than the reduction of the RCS provided by these multipole frequency selective absorbers, the skirt selectivity of the absorption band with high out of band reflection can be beneficial for other applications. In fact, the high reflectivity obtained outside the absorption band is relevant in the realization of dual-band reflector antennas [81] for satellite communication applications.



Figure 4.21. Simulated reflection coefficient of the passive absorber with three different customized thicknesses.

To demonstrate the absorber capability in adapting to various design thicknesses, the response of the two-pole absorber is obtained for different values of the MF500F-112 thickness h_1 (2, 3 and 4 mm). The physical dimensions of the absorber for each configuration without the plasma-shells are summarized in Table 4.5. As shown in Figure 4.21, all three

design configurations provide virtually the same absorption response. The results predict a second order response of about 17% fractional bandwidth at 10 dB reflectivity level.



Figure 4.22. Proposed unit cells of the two-pole tunable plasma-based absorber based on coupling interlayers. The physical parameters used to obtain absorption in the C-band are provided: $h_1 = 3.2$; $h_2 = 1.57$; a = 10; $g_1 = g_2 = 0.4$; $s_1 = s_2 = 1.6$; $b_1 = 9.24$; $b_2 = 8.48$; $l_1 = l_2 = 6.5$; b = 6.5; $w_1 = w_2 = 6$; r = 3.4 (units in mm)

In order to add the reconfigurability feature to the multipole absorber, cuboid cavities are created within the lossy magneto-dielectric substrate, thereby allowing the hollow ceramic gas-encapsulating chambers (plasma-shells) to be embedded in the design as shown in Figure 4.22. The shells (size: 6 mm x 6 mm x 3 mm) are filled with 0.1% Argon - 99.9% Neon with the electron collision frequency value set to: $v = 2.0 \times 10^{10} rad/s$. In the active layer, both the resonators and the coupling interlayers are used as biasing surfaces for the shells. In order to minimize field enhancement at the sharp corners of the resonators, a fillet of radius 0.7 mm is performed at the corner of the resonators patch. The



Figure 4.23. Full wave simulated reflection coefficient of the multilayer absorber at OFF state for various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.

physical parameters of the active absorber are provided in Figure 4.22. Simulation results shown in Figure 4.23 predict stable absorption response across the absorption band for both TE and TM polarizations when the system is illuminated from various oblique angles $(0^{\circ} \le \theta \le 45^{\circ})$.



Figure 4.24. Full wave simulated reflection coefficient of the multilayer absorber at ON state subject to different plasma frequencies. Tuning capability observed for (a) Only the top plasma layer is activated and (b) Both plasma layers are activated.

The variation of the biasing voltage is mimicked by applying different plasma frequencies through simulation. On the one hand, we assume that only the top plasma layer is excited ($\omega_{P1} = 0$). The variation of the plasma frequency ($0 \le \omega_{P2} \le 9 \times 10^{11} [rad/s]$) predicts a tuning of the absorption spectrum and rate. The higher absorption resonant frequency has shifted from 6.2 to 6.7 GHz (see Figure 4.24(a)). However, wider absorption band is obtained at the expense of the absorption rate. On the other hand, when both plasma layers are excited, the effect of the variation of both plasma frequency such that $\omega_{P1} = \omega_{P2}$ (see Figure 4.24(b)) predicts a tuning of both resonant frequencies across the C-band. It can be seen that the center frequency shifts to higher frequency (from 6 GHz to 7 GHz).



Figure 4.25. Fabricated layers. (a)) EBG resonator #1; (b) Coupling layer; (c) EBG resonator #2; (d) Perforated lossy magneto-dielectric (MF500F-112); (e) Design stacked up without the MF500F-112.

4.3.2.2 Fabrication Processes and Measurement Results

The proposed multilayer plasma-based tunable absorber is fabricated and tested in a free space environment to validate the numerical results. The metallic resonators and coupling layer are patterned on the FR4 substrates using a wet etching process. Also, the magneto-dielectric substrates MF500F-112 are finely drilled and perforated using a milling machine (LPKF ProtoMat S103) as shown in Figure 4.25. The prototype board is an array of 13 × 13 elements, printed on the FR4 substrates with a total size of 130 mm × 130 mm.



Figure 4.26. (a) Photograph of the test setup (free space); (b) Fabricated prototype (without the plasma-shells) inside a 3D printed frame; (c) Measurement vs simulated result of the absorber without plasma-shells.

First, a pre-test is performed to evaluate the performance of the proposed absorber without the plasma shells. A frame made out of thermoplastic material using a 3D printer (Monoprice Maker Select v2) is used for fine alignment and to hold the layers together. A small aperture, the size of the absorber is carved out at the center of a commercially available large metal-backed foam broadband absorber (Eccosorb AN-75) to mount the DUT as shown in Figure 4.26. The Eccosorb AN-75 frame is used to reduce diffraction effects from the edges of the DUT. The calibration procedure is the same as previously described. The measured return loss of the absorber without plasma-shells compared to its counterpart simulated result is illustrated in Figure 4.26. The results predict a second order response of about 14% fractional bandwidth at 10 dB reflectivity level. The link between simulated and measured results show that the MF500F-112 magneto-dielectric substrates have been accurately modeled in the simulation tool. A slight contrast between both results

can be due to the small size of the prototype and the material tolerances. The plasma-shells used for this experiment are commercially available from Imaging Systems Technology Inc. The shells are filled with 0.1 % Argon - 99.9 % Neon mixture at 100 Torr gas pressure. The gas pressure is estimated from the electron collision frequency ($\nu = 2.0 \times 10^{10} \text{ rad/s}$) using Equation (4.7) a theoretical expression for neon gas derived in [65] at 300 K electron temperature.



Figure 4.27. (a) Shells are hand-placed in the perforated lossy magneto-dielectric backed by the ground conductor; (b) Conductive silver epoxy are stenciled on the top side of the shells using a syringe; (c) Bottom plasma excitation layer is added on top of the lossy magneto-dielectric; (d) The perforated lossy magneto-dielectric is added on top of the bottom plasma excitation layer. (e) Shells with epoxy deposited on both top and bottom side are hand-placed in the substrate chambers. (f) Top plasma excitation layer is added on top of the lossy magneto-dielectric.

The fabrication procedure of the absorber with the plasma shells embedded in the lossy perforated magneto-dielectric substrates MF500F-112 is illustrated in Figure 4.27. The performance of the absorber (at OFF state) across different angles of incidence for both
TE and TM polarizations is experimentally obtained. Measured results Figure 4.28 show acceptable and stable reflection response when the system is illuminated from various oblique angles ($0^{\circ} \le \theta \le 45^{\circ}$). Both the measured results (Figure 4.28) and simulated results (Figure 4.23) demonstrate acceptable agreement although the measured center frequency has shifted to 6.3 GHz.



Figure 4.28. Measured reflection coefficient of the multilayer absorber at OFF state for various oblique angles of incident wave. (a) TE polarization. (b) TM polarization.

In order to turn the device ON, a class A, RF power amplifier (ENI 2100L with 50 dB nominal gain and operating frequency ranging from 10 kHz to 12 MHz) is used to sustain the plasma layers. The RF power amplifier (PA) amplifies a sinusoidal wave produced by Keysight's N5181A MXG RF analog signal generator operating at 2 MHz. The output of the PA is then connected to the bias traces of the absorber via a coaxial cable. By monitoring the power level of the continuous wave through the PA front panel meter and changing the voltage from the signal generator, different voltage is supplied to the plasma volume. The tuning speed is less than 100 ns. The excited device under test is shown in Figure 4.29. When the top plasma layer is only excited with various RF power such that $P_1 = 0$ and $0 \le P_2 \le 67$ W, the measured results at normal angle of incidence indicate a tuning of the higher absorption resonant frequency (see Figure 4.30(a)). However, when both plasma layers are activated ($0 \le P_1 = P_2 \le 67$ W), a tuning of the absorption spectrum range is perceived (see Figure 4.30(b)) showing a shift of the absorption center frequency. The



Figure 4.29. Photograph of the test setup (free space) when the absorber is excited with various RF powers.



Figure 4.30. Measured reflection coefficient of the multilayer absorber at ON state. The plasma volume are excited independently excited with various RF power intensities. Tuning capability observed for (a) Only the top plasma layer is activated and (b) Both plasma layers are activated.

measured results obtained in Figure 4.30 agree well with their simulated results shown in Figure 4.24. These experimental results validate the fact that the proposed absorber, when properly biased, can be adapted under a dynamic EM environment.

4.4 Multiphysics Analysis of Two-Pole High Power Microwave Absorber

Multiphysics analysis is a powerful design process that can be used to understand multiple physical phenomena of electromagnetic devices under real world conditions. Such analysis, composed of electromagnetic (EM) simulation coupled with thermal, structural, and/or fluidic simulations [82, 83] are critical for the design liability associated with performance and cost. One of the real world applications of multiphysics analysis involves the impact of high power electromagnetic interference (EMI) with electronics devices. In such events, researchers have focused their understanding on electronic attacks (EA) and electronic protection (EP). These inquiries have led to tremendous development capabilities that are evolving rapidly with modern technology [84–86]. Advanced countermeasures electronic systems, such as electronic coating shielding, isolators, filter limiter, frequency selective surface, microwave absorber [87–91], have been explored to ensure functional safety of EM devices. Conversely, techniques such as electronic jamming and deception, anti-radiation weapons, use of HPM and EMP weapons [92–98] are often exploited to target those protective layers and damage the electronic and electric systems through EM coupling.

The exposure of electronic systems to high power level EM interference, whether it is intentional or not, can disrupt their performance and cause device failure depending on the core material and the microscopic features of the systems architecture. Usually, the electric field induced in the system caused by the incident wave are associated with current and voltage surge that can initiate electromagnetic breakdown, arcing or overheating of the system [99]. Under such conditions, design specifications such as EM-thermal-mechanical management can no longer be neglected. The ANSYS multiphysics simulator has proved to be a reliable and powerful software package for the study of EM-thermal-mechanical coupling [100, 101]. While the EM analysis is performed by ANSYS HFSS, both thermal analysis and stress analysis can be performed using ANSYS Mechanical. For instance, if the electronic system is lossy, the electric fields calculated from Maxwell's equations are used to compute the power dissipated in the system. The EM losses (surface and volume loss densities) are then converted into heat using heat transfer equation. Then, the temperature distribution can be utilized to analyze the thermal stress (expansion or contraction). The deformation caused by the thermal stresses and the change in temperature are fed back into HFSS and will affect the electro-thermal properties of the materials. This cycle is repeated until the temperature reaches its steady state. The interaction between EM and heat transfer using both ANSYS HFSS and ANSYS Mechanical provides an accurate and complete bidirectional coupled physics analysis.

In this section, the physical limitation of a plasma-based tunable conductor-backed absorber under high power microwave levels is demonstrated using multiphysics numerical analysis. The peak power (for very short duration pulse) and average microwave power (for continuous wave) handling capability of the active absorber is investigated. The numerical analysis is carried out to predict the dielectric and air breakdown levels within the system. Also, since the absorber converts the EM energy into thermal energy, the EM wave and heat transfer are coupled together to evaluate the thermal behavior of the absorber. Both the steady state and transient state analysis are performed when the device is exposed to various incident power densities. A non-uniform temperature distribution is obtained with hotter spot located within the lossy magneto-dielectric. Since testing at high power microwave levels is potentially unsafe, experiments are performed with relatively moderate far-field power densities to validate numerical results. The temperature distributions on the top surface of the absorber are obtained using a thermal imaging infrared camera (FLIR E6).

4.4.1 Physical Impact of High Power Microwave Source on the Proposed Multilayers Absorber

High peak and average microwave power sources present severe challenges for the reliability and safety of resonant structures. Under exposure of high-power RF, the physical interaction between the EM field and the electronic system on a molecular level must be investigated. Electromagnetic breakdown and overheating of the device are feasible causes of design failure. In general, EM breakdown events are the primary concerns associated with very short pulse durations (in terms of μ s) of high peak power. This happens when the induced electric field within the device becomes greater than the breakdown limit of air or dielectric. In our resonant absorber, the slots within the metallic layers enhance the localized electric field intensity from a buildup of negative and positive charges. As a result of the electric discharge, current will flow through the dielectric to create a short circuit as the insulating material becomes a good conductor. On the other hand, overheating of the system is associated with high average power with sustained continuous wave. The energy dissipated in the system due to the time averaged metallic and dielectric losses is converted into heat, which can become problematic depending on the selected design materials.

4.4.2 Estimation of the Absorber Breakdown Threshold

The prediction of the absorber electrical breakdown level is a very complex task because the discharge process is transient and occurs on a microscopic level. While many factors can cause microwave-metal electrical discharge mechanism, we exploit the effect of the field enhancement caused by high peak microwave power. Numerically, the dielectric and air breakdown level within the absorber can be estimated by quantifying the maximum field enhancement factor (MFEF) [102]. MFEF is defined as the ratio between the maximum local electric field intensity inside the system and the electric field intensity of the incident EM wave. Therefore, a microwave system resilient to high power RF is expected to have lower MFEF.



Figure 4.31. Simulated E-field distribution in the metallic layers at a unit cell level of the absorber illuminated with an incident power of 1 W. The electric field with intensity $E_0 = 2746$ V/m is coupled to the absorber. The induced field distribution is observed at (a) the first resonant frequency $f_1 = 5.9$ GHz (b) the second resonant frequency $f_2 = 6.1$ GHz



Figure 4.32. Extracted MFEF values at different frequencies of the absorber without and with filleted resonators corners.

However, at resonance the fields inside small metallic gaps are maximally enhanced.

The electric field distribution in the metallic layers of the absorber unit cell is numerically obtained at both resonance frequencies $f_1 = 5.9$ GHz and $f_2 = 6.1$ GHz. When the cross section of the unit cell (with periodicity a = 10 mm) is illuminated with an incident power of 1 W, a maximum incident power density of $10\,000$ W/m² which is equivalent to an electric field intensity of $E_0 = 2746$ V/m is coupled to the system.

Table 4.6. Approximated maximum allowed power density (in MW/m^2) for very short pulse duration of high power.

Design	5.8 GHz	5.9 GHz	6.0 GHz	6.1 GHz	6.2 GHz
Without filleted resonator corners	12.4	11.9	14.8	12.2	13.7
With filleted resonator corners	18.3	16.3	25	17.6	26.5

The results shown in Figure 4.31 predict strong local field enhancement at both EBG resonators. It can be seen that, the field is strongly enhanced at the edges of the metallic layers. The MFEF of the absorber (without/with resonator corners filleted) is extracted across the C-band (4GHz to 8GHz) using full-wave EM simulation and the result is depicted in Figure 4.32. It is perceived that the proposed design with filleted resonators corners shown better performance in handling higher microwave power. The value of the MFEF is higher in the operation band of the absorber with maximum (27 and 26) observed at both resonant frequencies for the proposed design. At OFF state, the air gap existing in the cavity is the most limiting medium with electrical breakdown strength $E_B = 3 \text{ MV/m}$. For very short pulse duration of high peak power, the maximal power density the absorber can withstand at specific frequencies is depicted in Table 4.6. Overall, the data obtained showed the maximum allowed incident power density is about $S = 16.3 \text{ MW/m}^2$ for the proposed multilayer absorber at OFF state.

4.4.3 Thermal Analysis of the Proposed Multilayer Absorber

As mentioned previously, when the absorber is coupled with a high average power continuous wave (CW), thermal analysis needs to be taken into account to avoid design failure due to material heating and possible burning. Therefore, it is important to understand how the material choice can affect the thermal and electrical performance of the design. The CW incident power density allowed by the design can be obtained based on the thermal limit of their temperature dependent materials. A comprehensive multiphysics solution that couples full-wave EM to thermal analysis is performed using HFSS and Icepak from ANSYS.

		1		1
Material	Thermal conductiv-	Density	Specific heat	Operating temper-
	ity $(W.m^{-1}.K^{-1})$	$(Kg.m^{-3})$	$(\tilde{J}.Kg^{-1}.K^{-1})$	ature limit (°C)
Copper	401	8933	356	< 750
Alumina	27	3970	910	< 1750
Neon	0.0498	0.1079	1029.9	< 1000
FR4	0.3	1250	1300	< 140
Rogers RO4003C	0.71	1700	900	< 280
MF500F-112	1.44	3250	1300	< 260

Table 4.7. Thermal properties of the absorber materials.

In general, the temperature dependency of the materials' electrical properties (such as dielectric constant, loss tangent, conductivity) are governed using a quadratic approximated equation [103]:

$$x(T) = x(T_0) \left[1 + C_1(T - T_0) + C_2(T - T_0)^2 \right].$$
(4.10)

where T_0 is the initial temperature, T is the temperature of the heated material (which can be position dependent as well), and C_1 and C_2 are the linear and quadratic expansion

coefficients, respectively. A bidirectionally coupled analysis is possible if the temperature dependency of all the materials electrical properties are known. In such case, the transient frequency response of the design due to temperature drift and structural deformation can be also obtained. However, in this section our analysis focuses on the heat generated solely due to Joule heating within the absorber for different incident power densities.



Figure 4.33. Simulation model of the finite size multilayer absorber illustrating the coupling between ANSYS HFSS and ICEPAK.

The thermal specification of the design modeled in Icepak is illustrated in Table 4.7. These values are taken under ordinary conditions and it can be seen that the FR4 substrate is the limiting material in terms of the maximum operating temperature (<140 °C). Beyond this threshold, known as the glass transition temperature (T_g), the hard polymer will become soft and lose its electrical and mechanical integrity. The simulation set up in Icepak is performed in a natural convection environment (heat transfer coefficient *HTC* = 10 W · K⁻¹ · m⁻²), with the radiation ON (for heat transfer) and ambient external temperature. In order to predict the temperature profile of the design, the exact finite size of the absorber is taken into account (as shown in Figure 4.33) to consider the edge effects, although a heavy simulation CPU time is required.

Both the steady and transient state simulation results are obtained when the absorber is



Figure 4.34. Simulated steady state temperature map throughout the multilayer absorber exposed to various power densities with incident plane wave at $f_0 = 6.0$ GHz.

Table 4.8. Simulated maximum temperature (°C) within the FR4 substrates for different power densities across the absorption spectral band.

Power density	5.8 GHz	5.9 GHz	6.0 GHz	6.1 GHz	6.2 GHz
$S = 0.1 \mathrm{kW/m^2}$	23.5	24.9	24.6	24.8	24
$S = 1.0 \mathrm{kW/m^2}$	58	59	58.5	59.5	58
$S = 3.7 \mathrm{kW/m^2}$	140.5	142	141	151.5	140
$S = 4.0 \mathrm{kW/m^2}$	147.5	148.5	148	148.7	148
$S = 8.0 \mathrm{kW/m^2}$	231	235	235	234	230

exposed to a continuous plane wave. The steady state temperature profile of the absorber exposed to various power densities at the center frequency $f_0 = 6.0$ GHz is illustrated in Figure 4.34 (which reflect the temperature plot in both substrates and metallic surfaces). It appears that a non-uniform spatial temperature distribution is obtained such that the peak temperature is located within the magneto-dielectric MF500F-112, where the substrate losses are the most significant. A report of the simulated maximum temperature generated within the FR4 substrate for different power densities across the absorption spectral band is summarized in Table 4.8. It can be seen that the temperature increases with the power density and slightly higher values are obtained at the resonant frequencies. A transient



Power density $S = 4 kW/m^2$



Figure 4.35. Simulated transient state temperature map throughout the multilayer absorber with incident plane EM wave at $f_0 = 6.0$ GHz. The absorber is exposed to various power densities. (a) S = 0.1 kW/m²; (b) S = 4.0 kW/m².

simulation is also performed on the multilayer absorber. The 3D temperature distribution throughout the absorber is obtained within one-hour time frame for various power density at the center frequency as shown in Figure 4.35. The initial temperature of the absorber is set to the room temperature ($T_0 = 20$ °C). The transient simulated results show that the temperature within the absorber increases with time and converges to the steady state



Figure 4.36. Time variant simulated peak temperature plot throughout the FR4 (plot using symbol) and the MF500F substrate (solid line plot) illustrated for various power densities with incident plane EM wave at $f_0 = 6.0$ GHz.

after one hour. The plot in Figure 4.36 (maximum temperature versus time) is provided to illustrate the trend of the temperature increase over time within the FR4 and the magnetodielectric MF500F. It's observed that the temperature rises quickly for higher incident power density. Based on the data projected, the maximum average power density the absorber can tolerate at OFF state for an incident CW is about $S_A = 3.7 \text{ kW/m}^2$.

4.4.4 Experimental Results of the Temperature Distribution on the Top Surface of the Absorber

Testing the power handling capability of specific microwave components requires technical safety measures. All the test instruments need to be rated for the measurement environment. Safety practices recommended by IEEE standards on high power testing should be adopted. The test setup can differ given the specific device under test (DUT). For example, when testing the power handling of a microwave filter, the input of the DUT can be driven with a coax connection and the output terminated using a high power load. Radiated susceptibility measurement of antennas are often performed in a semi-anechoic chamber. However, all these test procedures require the use of a high power source.





(D)

Figure 4.37. (a) Sketch and (b) photograph of the free space test setup.

To validate the thermal numerical results of the multilayer tunable plasma-based absorber, a 40 dB traveling wave tube (TWT) amplifier from Hughes (Model#: 1177H13F000) operating in the S/C-band (3 GHz to 8 GHz) is used to generate a high power microwave. Since the power handling limit of the absorber is related to the thermal effect, the temperature of the absorbing material under RF fields is measured using a thermal imaging infrared camera (FLIR E6). This IR camera has the capability to detect object temperatures ranging from -20 °C to 250 °C with 2% reading accuracy. As a safety precaution, the thermal analysis tests are performed with relatively moderate far field power densities. Although the incident power level might not test the design to its limits, it is sufficient enough to cause heat generation that can be measured using the IR camera. An illustrative sketch of the test setup is shown in Figure 4.37(a). Prior to the experiment, a 6.3 GHz sinusoidal wave produced by Keysight's E8257D PSG RF analog signal generator is amplified by the TWT power amplifier and the output is monitored using the E4418B EPM series power meter. Thus, by placing the DUT at the far-field (0.5 m) from the PE9887-11 broadband horn antenna, the incident power density is evaluated using the equation:

$$S = \frac{P_t G_t}{4\pi R^2} \tag{4.11}$$

where P_t is the transmitted power, $G_t = 12.6$ is the gain of the transmit horn antenna, and R is the distance from the horn antenna to the center of the top surface of the DUT. The measured thermal distribution is obtained at three different power densities (S = 0.08, 0.1 and 0.3 kW/m²). A photograph of the test measurement is shown in Figure 4.37(b). Since only the top surface of the absorber is visible to the FLIR E6 IR camera, the temperature distribution on the top surface of the absorber is measured and compared to the numerical results. The numerical analysis of the thermal transient state of the absorber upon an incident RF power density at 6 GHz is performed.

The simulated temperature distribution obtained on the top surface of the absorber for a power density of $S = 0.1 \text{ kW/m}^2$ is shown in Figure 4.38(a). In order to accurately measure the radiated temperature on top of the absorber, the IR camera is properly calibrated,



Figure 4.38. (a) Transient state temperature map observed at the top surface of the multilayer absorber exposed to a power density power $S = 0.1 \text{ kW/m}^2$. (a) Simulated results for incident plane EM wave at $f_0 = 6.0 \text{ GHz}$; (b) Measured results for incident plane EM wave at $f_0 = 6.3 \text{ GHz}$.

the emissivity is set to 0.91 (emissivity of FR4) and the distance between the object and camera is set to 0.5 m. Images obtained from the IR camera are depicted in Figure 4.38(b). It shows that the top surface of the absorber becomes hotter for different time frames. The qualitative aspect of the simulated temperature profile does not exactly correspond to the measured temperature map obtained of the top surface of the multilayer absorber. Several parameters such as the color visualization used in the simulation of the design model, the lighting in the test room, or the position of the IR camera surely influence the results. However, the measured temperature obtained at the center of the top surface of the multilayer absorber accurately validates its numerical counterpart as observed in Figure 4.39 for various incident power densities.



Figure 4.39. Simulated/measured time variant temperature plots obtained at the center of the top surface of the absorber versus time illustrated for various power densities with incident plane EM wave.

4.4.5 Techniques to Reduce the Absorber Susceptibility to Fail under High Power Excitation

The design of microwave absorbers resilient to high power is very intriguing, since the absorbed incident energy is converted to heat via phase cancellation. However, multiphysics analysis of absorbers can result in optimized design solutions. In [104], optimization techniques have been investigated to mitigate field enhancement in resonant structure. The widening of capacitive gaps between metallic patches, filleting of sharp metallic corners, and the use of genetic algorithms (pixelized screen geometry) have reduced the MFEF of such design. In addition, by substituting capacitive metallic layer with high permittivity thin dielectric, the MFEF value can be further reduced [102].

In the case of thermal management, custom design techniques such as the use of hollow pyramidal and honeycomb lossy material, integration of coolant system within the design, and the use of high temperature materials can increase the power handling capability. For instance, RO4003C, a hydrocarbon ceramic laminate from Rogers Corporation's RO4000



Figure 4.40. Simulated reflection coefficient of the alternate multilayer absorber (using Rogers) along with the previous design (using FR4) at OFF state. The physical parameters of the alternative design Type 2 (using Rogers RO4003C) used to obtain absorption in the C-band are provided: $h_1 = 3.2$; $h_2 = 1.5$; a = 10; $g_1 = g_2 = 0.4$; $s_1 = s_2 = 1.65$; $b_1 = 9.4$; $b_2 = 8.66$; $l_1 = l_2 = 6.5$; b = 6.5; $w_1 = w_2 = 6$; r = 3.6 (units in mm).



Figure 4.41. Full-wave simulated reflection coefficient of the multilayer absorber at ON state subject to different plasma frequencies when Both plasma layers are activated.

series (which are known for their reliability when subjected to severe thermal shocks with transition temperature $T_g = 280 \,^{\circ}\text{C}$) has been incorporated in the design as a substitute for the FR4 dielectric. Under these circumstances, the magneto-dielectric MF500F becomes the limiting material in terms of the maximum operating temperature (<260 $^{\circ}\text{C}$). For a fair comparison, the two types of absorber have the same thickness, unit cell periodicity, and

operating frequency band, with virtually the same fractional bandwidth.

The reflectivity response of this alternate design along with the previous design is shown in Figure 4.40. The physical dimensions (in mm) of the alternative design are also depicted in the caption of Figure 4.40. Figure 4.41 shows the effect of the variation of both plasma frequency such that $\omega_{P1} = \omega_{P2}$ when both sources are turned ON. It also predicts a tuning of both resonant frequencies across the C-band such that the center frequency shifts from 6 GHz to 7 GHz. The plot of the time variant simulated peak temperature within the magneto-dielectric MF500F (Figure 4.42) indicates that the average power handling capability under sustained CW excitation increases from 3.7 kW/m² (for the design using FR4) to 10 kW/m² (for the one using Rogers RO4003C).



Figure 4.42. Time variant simulated peak temperature plot throughout the RO4003C (plot using symbol) and the MF500F substrate (solid line plot) illustrated for various power densities with incident plane EM wave at $f_0 = 6.0$ GHz.

4.5 Chapter Conclusion

A low profile, compact, high performance electronically tunable microwave absorber based on plasma properties is revealed in this Chapter. Numerical solutions of the propagation of the electromagnetic wave in plasma devices had been provided using transmission line approach. The practicability of the design, the biasing networks, the fabrication process and experimental verification were satisfactory. The feasibility of devising electronically tunable, high power resilient compressed Jaumann absorbers relies on applying thickness controlled inter-coupling layers and integrating tiny discrete plasma-shells. The power handling capability of the proposed multilayer absorber is studied using multiphysics analysis.

Chapter 5

Conclusion

In this work, we demonstrated innovative design techniques for next-generation frequency selective surfaces and absorbers. High performance multipole bandpass spatial filter is implemented by cascading miniaturized resonant element and inverter layers. The realized filters deliver maximally flat multipole stable inband responses. The separation between adjacent FSS layers can be arbitrary (at very close distance or at moderately spaced distance), thus providing flexible design control of both passband response and overall filter thickness. Ultimately, given overall thickness specification and filter characteristics, the proposed generalized synthesis technique provides the optimal inter-coupling dimensions allowing a magnetic coupling path between resonant layers to produce the desired response. Building on this achievement, the concept is applied to practical engineering problems including the design of low profile, compact reconfigurable FSSs and microwave absorbers. The design flexibility allows simpler integration of active components that require specific size. The integration of discrete plasma-shells into the proposed compact multilayered FSS/absorber adds reconfigurability property to their performance while achieving efficient and stable multifunctional operation in harsh environment. This novel electromagnetic radiation protection system has potential application in modern stealth technology and protection from high-power electromagnetic interference.

In Chapter 2, we introduce a new design guide that allows simultaneous control of the EM wave response and design thickness of multilayered bandpass FSS filters. The proposed technique is based on coupled bandpass filter theory using admittance inverter. A generalized ECM is systematically used to discover the physical mechanism behind the interaction of the multilayers. It has been demonstrated that given center frequency and bandwidth of filter specification, there is a degree of conformity for thickness customizable high-order bandpass FSS with some limitations in term of the design profile. The overall thickness of multipoles FSSs can be deliberately designed without losing any generality of the spatial filter performance. The proposed technique is suitable for FSS applications requiring precise thickness specifications and also allows the engineer to use readily available commercial dielectric materials with standard thickness for multilayered FSS design technology.

In Chapter 3, a practical large scale switchable electromagnetic field blocking FSS based on discrete plasma-shells is proposed. The discrete plasma-shells allow the realization of large scale plasma devices and easy control of the plasma density by using a simple FSS layer as a plasma excitation surface. Method of understanding and controlling the plasma behavior has been implemented by extracting electrical parameters of the plasma. Then we examined this feature as switchable field blocking components of FSS. A comparison between simulation and measurement in free space or waveguide environment have proven to be in good agreement. The proposed state of art switchable FSS definitely provides benefits in improving overall aerodynamics, conformability, and reliability. The fabricated large-area switchable plasma device exhibits a fast tunable shielding against low and high power microwave and has potential application in aerospace and terrestrial electromagnetic system protection. The design can be scaled to desired frequency bands including multi-band operation using proper design concept. These novel spatial filter devices address the limitation of tunable FSS based semiconductor varactor and pin diode where high power is a potential threat in electronic warfare.

In Chapter 4, a low-profile tunable absorbers based on lossy FSS layers embedded with discrete plasma-shells is investigated. The bulk plasma is modeled using parallel plate topology and a transmission line approach is provided to explain the working principle of the proposed plasma-based tunable absorbers. Through full wave electromagnetic simulation, it is demonstrated that by adjusting the plasma frequency of the gas volume, a tuning of the absorption band is achieved. The large-scale plasma-based absorber is fabricated and tested. The biasing voltage used to sustain the plasma is properly adjusted to control the transfer function of the absorber to validate the tunability capability that was expected from the simulation results.

Furthermore, by leaning toward applications requiring protection from high power HPM/EMP emanated by frequency agile microwave weapons or radar systems, we introduced a compact multilayer tunable absorber based on lossy magneto-dielectric substrates embedded with discrete plasma-shells. The discrete plasma-shells located on different layers of the structure allow independent control of each resonant frequency. The techniques employed in the proposed design lend much freedom to customize both functional and physical characteristics of the absorber. This thickness customizable compact multilayer absorber is tuned in real time to provide a multifunctional response on demand for a dynamic use of the EM spectrum.

The power handling capability of our proposed multilayer absorber is studied. Notably, we investigate the physical effect of high peak and average power, which can present severe challenges in a harsh electromagnetic environment. A coupled physical phenomenon between both RF and thermal simulation is investigated by rigorous multiphysics simulated analysis. The results predict very well the thermal issues for high-power applications. High power analysis of microwave components is very crucial in a sense that selected materials can be rightfully chosen in the early stages of design in order to avoid failure of the final product for specific applications and assure a safe operation in harsh environments.

The outcomes of our research will lead to a new class of reconfigurable, high performance, high-order spatial filters and absorbers for stable communication links from high power interference. While this research was only applicable to planar EM structures, future research can explore the versatility of the proposed technique on conformal multilayered reconfigurable FSSs/absorbers to validate the usability and the effect of incorporating inverter layers under non-planar structural formations. Also, given high demand for broadband and high data rate communication, plasma-based reconfigurable structures will be investigated over a wide range of spectrum from millimeter wave, terahertz (THz), up to optical frequencies. This forthcoming research will present certain degree of theoretical and experimental challenges due to the micro-scale design features.

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Publications

Journal Articles

- J1. K. Payne, K. Xu, J. H. Choi, and J. K. Lee, "Multiphysics analysis of plasma-based tunable absorber for high power microwave applications," *IEEE Transactions on Antennas and Propagation*, 2020 (in review).
- J2. K. Payne, K. Xu, J. H. Choi, and J. K. Lee, "Electrically tunable microwave absorber based on discrete plasma-shells," *IEEE Transactions on Antennas and Propagation*, vol. 67, pp. 6523–6531, Oct 2019.
- J3. K. Payne, K. Xu, and J. H. Choi, "Generalized synthesized technique for the design of thickness customizable high-order bandpass frequency-selective surface," *IEEE Transactions on Microwave Theory and Techniques*, vol. 66, pp. 4783–4793, Nov 2018.
- J4. K. Payne, K. Xu, J. H. Choi, and J. K. Lee, "Plasma-enabled adaptive absorber for high-power microwave applications," *IEEE Transactions on Plasma Science*, vol. 46, pp. 934–942, April 2018.

Conference Presentations

C1. K. Payne, J. K. Lee, K. Xu, and J. H. Choi, "Higher order plasma-based tun-able absorber using magneto-dielectric substrates," in 2019 IEEE International Symposium

on Antennas and Propagation and USNC-URSI Radio Science Meeting, pp. 2027–2028, July 2019.

- C2. K. Payne, J. K. Lee, K. Xu, and J. H. Choi, "Compact third-order bandpass frequency selective surface with wide spurious-suppression bandwidth," in 2018 IEEE International Symposium on Antennas and Propagation USNC/URSI National Radio Science Meeting, pp. 2037–2038, July 2018.
- C3. K. Payne, J. K. Lee, K. Xu, and J. H. Choi, "Low-profile plasma-based tunable absorber," in 2018 IEEE International Symposium on Antennas and Propagation USNC/URSI National Radio Science Meeting, pp. 2065–2066, July 2018.
- C4. K. Payne, J. H. Choi, E. Peters, D. Wedding, and C. Wedding, "Plasma-tunable electronic protection for microwave applications," in 2017 IEEE International Symposium on Antennas and Propagation USNC/URSI National Radio Science Meeting, pp. 259–260, July 2017.
- C5. K. Payne, J. H. Choi, and T. Jang, "Ultra-low profile tri-polarized antenna for WLAN/MIMO application," in 2017 IEEE International Symposium on Antennas and Propagation USNC/URSI National Radio Science Meeting, pp. 753–754, July 2017.
- C6. T. Jang, K. Payne, L. J. Guo, and J. H. Choi, "Half-mode hexagonal substrate integrated waveguide structure and its application," in 2017 IEEE MTT-S International Microwave Symposium (IMS), pp. 487–490, June 2017.
- C7. K. Payne and J. H. Choi, "Development of dual-band/dual-polarized metamaterial antenna in an electrically-small form factor for efficient wireless data and power transfer," in 2016 IEEE Conference on Antenna Measurements Applications (CAMA), pp. 1–4, Oct 2016.

- C8. K. Payne, E. F. Peters, J. Brunett, D. K. Wedding, C. A. Wedding, and J. H. Choi, "Secondorder plasma enabled tunable low-profile frequency selective surface based on coupling inter-layer," in 2016 46th European Microwave Conference (EuMC), pp. 309–312, Oct 2016.
- C9. K. Payne and J. H. Choi, "Third-order frequency selective surface based on inverter interlayers," in 2016 IEEE International Symposium on Antennas and Propagation (AP-SURSI), pp. 953–954, June 2016.
- C10. K. Payne, J. H. Choi, M. A. Ali, and C. M. Wu, "Highly-selective miniaturized firstorder low-profile dual-band frequency selective surface," in 2016 IEEE International Symposium on Antennas and Propagation (APSURSI), pp. 955–956, June 2016.