Spatial periodic array with frequency-selective bandpass response for GPS and DCS1800 mobile communications

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**Spatial Periodic Array with
Frequency-Selective Bandpass Response
for
GPS and DCS1800 Mobile Communications**

Teo Peng Thian

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfilment of the requirement for the degree of
Master of Engineering

2005

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STATEMENT OF ORIGINALITY

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

23 FEB. 2005

Date

Teo Peng Thian

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SUMMARY

This thesis explores the design and implementation of periodic arrays with frequency selective response for a novel antenna that is developed for low frequency microwave communications usage.

It begins with a review of the physics and the analytical techniques essential for the design of infinite periodic arrays. Several test cases are implemented to verify the accuracy of the design approach. A unique test case is constructed to illustrate, with the aid of visualisation tool, the field and wave phenomena associated with the grating lobe, bandstop and bandpass conditions that are otherwise too abstract to appreciate.

Five uniquely-designed periodic arrays are then presented with various degrees of roll-off and bandstop performance while maintaining a consistent bandpass response for the low frequency Global Positioning Satellite (GPS), Mobile Satellite (MSAT) and Global System for Mobile Communications-Digital Communications System (GSM-DCS1800 band) operation. The transmission line-cum-equivalent circuit method is then extended to the smith chart analysis of these arrays, which comprises of bandpass and bandstop, novel convoluted, single, double, quad-layer and quad-layer with staggered-tuning design variants.

The construction of the novel balun-fed circularly polarised Maltese cross-shaped antenna is also presented in this thesis. Results of integrating the antenna with the periodic array shows that out-of-band scattering reduction is achievable with minimal impact on the antenna in-band performance. Finally, the seemingly contradictory two-fold impact of the frequency-selective periodic array performance on the link budget equation is highlighted and explained.

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Chapter 1

Introduction

1.1 Background and Motivation

With the advent of satellite, mobile and personal communication systems, there is an increasing demand to provide a single integrated antenna and front-end receiver system [1]. Applications of such an integrated system can be extended to mobile communications networks with tracking and location-determination technology. Some of the specific communication products include: wireless enhanced 911 (E-911) service [2], [3], adaptive routing for road traffic [4] and intelligent transport system [5], [6]. For most of the traffic telematics, quasi-simultaneous reception of GSM and satellite navigation (GPS) signal is required [7]. To avoid the conventional approach of using two separate receivers, a radio-frequency back-end receiver circuit was reported for a combined GPS-GSM900 terminal [7]. System architecture and software algorithms were developed to couple GPS with GSM services [7], [8]. In Asia and Europe, GSM is available in both the 900 MHz (GSM900) and 1800 MHz (DCS1800) band [5].

While GPS requires circular polarisation (CP) operation, mobile communication (GSM) operates with linear polarization. A CP antenna can be used to receive a linearly polarized signal, with a 3-dB loss for receiving a perfect horizontal or vertical polarized signal [9]. Hence, the antenna presented in this thesis is designed for optimal GPS operation and then extended for GSM coverage rather than vice versa. Moreover, in most cases, a much more stringent voltage standing-wave ratio (VSWR) is required for the GPS operation, while for mobile communication, a VSWR of up to 3 (i.e. higher mismatch) or the equivalent of a return loss of 6 dB is tolerable [10]. The coaxial balun-fed antenna presented in this thesis can provide a circular polarization over a wide bandwidth compared to conventional CP microstrip antennas where certain tradeoffs are required between the impedance and CP axial ratio bandwidth [11]. It will be shown that the antenna gain, CP axial ratio bandwidth, and radiation pattern consistency of this novel Maltese Cross-shaped antenna is sufficiently attractive for GPS (1575.42MHz), MSAT

satellite-to-mobile band (1545MHz-1559MHz) [1] and mobile-to-satellite band (1646.5MHz -1660.5MHz) and GSM/DCS1800 operations.

Similarly, there is an increasing interest to determine and reduce the scattering of antennas. This is advantageous in the mitigation of interference from external sources, [12] as well as scattering and coupling towards external antennas and electronic systems [13]-[15] and the design of low observable sensors [16]. The scattering from an antenna consists of mainly the antenna mode and structural mode scattering [17]. The antenna mode scattering is associated with the radiation properties of the antenna while the structural mode scattering is mainly due to the re-radiation of the induced current on the antenna physical structure. In addition, antenna scattering can also be examined both in the operating band as well as in the out-of-band frequency range.

For in-band scattering, the incident wave energy will travel through the resonating antenna element into the feed network. Proper matching will reduce the scattering due to the feed and junction mismatch. In the ideal case where perfect matching is obtained for all the components, all incident energy that enters through the resonating element will be absorbed by the termination. This minimizes the antenna mode scattering. For out-of-band scattering (other than the higher harmonics), the antenna elements are not at resonance and hence the scattering is due mainly to the element rather than the feed network. Although it is possible to obtain a certain degree of scattering reduction through proper optimization of the distance between the antenna element and the ground plane [18], [19], this is usually effective over a narrow frequency band and is hence not attempted. On the contrary, frequency selective surfaces (FSS) have been shown to reduce the out-of-band scattering of antenna over a wide frequency band for specified directions via proper shaping [20]. Examples include the hybrid radome [21], conical [22] and tilted flat panel [23] FSS geometry. Besides application in scattering reduction, FSS can be used to reject unwanted interference and coupling that is affecting the antenna at its higher harmonics.

It should be highlighted that the electrical requirements for the FSS can differ for antennas that are similar, but deployed in different environments. In a different environment, the frequencies for the unwanted interference or hostile signal might change and this will result in different sets of out-of-band attenuation requirements for the FSS. In addition to the difference in electrical requirements, different environment present different mechanical constrains for the FSS installation. For applications with either space constrains or one that requires a curved filter surface, the complexity (e.g. fabrication and alignment) and thickness of a multi-layer FSS makes it less attractive compared to a single layer structure.

Given the possibility of deploying the GPS-cum-DCS1800 antenna [24] with different electrical and mechanical constrains, variants of a low-pass FSS will be designed and presented in this thesis. These FSS (single and multi-layer design) will produce various degree of roll-off while allowing the GPS and DCS1800 signal to pass through unattenuated. The electrical design, performance and trade-off with mechanical complexity will be compared. The effect of the FSS on the antenna in-band radiation performance and out-of-band scattering reduction will also be evaluated.

Various methods have been proposed for the analysis of FSS under the infinite array approximation, namely the periodic methods of moments [25], the spectral variant of the method of moments [26], the modal analysis method [27], the equivalent circuit models [28], the finite difference time-domain (FDTD) [29], the finite element method (FEM) [30] and Keylov-based reduced-order modeling [31]. The focus for the computational aspect is to achieve a more efficient code that could provide accurate analysis for finite size, complex (multi-layer and diverse material properties) FSS structures with wide incidence angle and diverse polarisation. The transmission line-cum-equivalent circuit [32]-[45] method, Febry Perot Interferometer (FPI) analysis [46]-[50], FEM CAD tool CAD [51]-[53] and the smith chart approach [54]-[57] will be used for the analysis and design of the FSS presented in this thesis.

1.2 Objectives

There are a few objectives for this project. The first objective is to develop a circularly polarised antenna that has a high gain, wide axial ratio ($AR < 3\text{dB}$) and impedance bandwidth capable of meeting both the GPS and DCS1800 operation. The proposed antenna should yield better performance for all the above aspects when compared to a microstrip antenna. Its fabrication tolerance should also be relatively less sensitive when compared to that of a microstrip antenna whose dimensions and alignment are critical for the excitation of the appropriate modes. The second objective of this project is to design and analyse variants of periodic array that have a frequency-selective bandpass response at the GPS and DCS1800 frequency range (i.e. 1.5GHz to 1.9GHz) while rejecting signals at the higher frequencies. The pass-band (in-band) is characterised by an insertion loss of approximately 0.5dB while the stop-band (out-of-band) is characterised by an attenuation of 10dB in the transmission coefficient. Periodic arrays in the form of single-layer and multi-layer FSS, as well as bandpass and bandstop FSS will be designed. The degree of roll-off and the variation of the stop-band bandwidth will be investigated for the various designs. It is yet another objective of the project to investigate the effect of the FSS on the antenna radiation and scattering performance.

1.3 Major Contributions of the Thesis

There are four major contributions from this research. Firstly, a novel coaxial balun-fed circularly polarized antenna is developed. This Maltese cross-shaped antenna consists of two orthogonal pairs of bow-tie elements with the antenna and phasing network designed independently to provide a more robust CP performance that is easy to implement and less sensitive to fabrication tolerance. This work was presented in a conference and published in two journal papers.

In the course of researching into analysis and design synthesis method for FSS, the FEM CAD tool Ansoft HFSS [53] has been used intelligently with appropriate test cases to illustrate and explain the field and wave phenomenon associated with the onset of grating lobes, propagation of free space grating lobes, bandstop and bandpass conditions. Also,

the need to de-embed the grating lobe field from the neighboring cells is highlighted for the first time. This work was also presented in one conference and accepted for publication in a journal.

The third contribution is the design, analysis and comparison of four FSS prototypes that meet the requirements for the low frequency GPS and DCS1800 frequency of operation. Amongst the prototypes, a novel convoluted single layer band-pass FSS has been developed with a relatively wide pass-band bandwidth of 22.47%. Selective signal reflection is achieved via selecting the length of the convoluted segments to be approximately a quarter-wavelength at the frequency of rejection. This work is currently under peer-review for possible publication in a journal. A quad-layer FSS with staggered tuning has also been developed with a relatively wide stop-band bandwidth. The adaptation of the surface reflection and array transmission phase for the analysis of the convoluted array and the detailed analysis of the admittance transformation on the smith chart for the multi-layer design are of considerable uniqueness. The preparation for a journal paper submission is in progress.

The final major contribution is a result of the attempt to integrate the FSS with the antenna. The effect of the FSS on the antenna radiation and the backscatter reduction is quantified. Unique phenomena as a result of the finite edges of the FSS are discovered. Also, the seemingly contradictory two-fold impact of the FSS performance on the radar equation is highlighted. The preparation for a journal paper submission is also in progress.

Overall, a total of five papers have been generated and several are in preparation for journal submission. These are listed in the List of Author's Publications.

1.4 Organization of the Thesis

There are six chapters in this thesis. In Chapter 1, the background, objectives and major contributions of this research are introduced. Chapter 2 is a review of the theory and

analysis of FSS based on a combination of the transmission line cum equivalent circuit method as well as the FPI approach. The usage of FEM CAD tool for the analysis of the surface current and wave phenomenon associated with the bandpass FSS is also included. Chapter 3 describes the design and performance of the various bandpass and bandstop FSS developed. Further analysis of each design using the smith chart is also presented in this chapter. In Chapter 4, details of developing a novel antenna intended for subsequent integration with the FSS are presented. The effect of the integration of the FSS with the antenna is reported in Chapter 5. The impact of the FSS on the RF radar equation is also presented. Finally, the conclusions and recommendations for future work are given in Chapter 6.

Chapter 2

Theory and Analysis of Frequency Selective Array

2.1 Introduction

This chapter describes the development of the transmission line [32]-[44] cum equivalent circuit model [44], [45] for the design of bandstop and bandpass FSS. The understanding of the phase transformation, admittance and reflection coefficient variation upon the progressive additions of materials and FSS sheets to the multi-layer stack-up will be useful in explaining the movement of the admittance locus on the smith charts presented for the analysis of multi-layer FSS in subsequent chapters. Also included in this chapter is a review of the concept of using the Fabry-Perot Interferometer (FPI) [46]-[50] to represent multi-layer FSS. This method only considers the fundamental mode present in the FSS structure. However, it provides physical insight as well as design guidelines for determining the spacing between multi-layer FSS design. Finally, the FEM CAD [51]-[53] will be introduced as an alternative means of analysis and design of FSS. The wave phenomenon and the physics associated with the resonance and grating lobes conditions that are otherwise difficult to quantify can be easily visualised and explained using the field display option provided by this CAD tool.

2.2 Transmission Line cum Equivalent Circuit Model [44], [45]

2.2.1 Reflection Properties for Single Layer Bandstop FSS

For a free-standing unloaded band-stop FSS, the impedance of the equivalent circuit shown in Figure 2.1 can be expressed by the following:

$$Z_{\text{FSS}} = j\omega L + \frac{1}{j\omega C} \quad (2.1)$$

and simplified to:

$$Z_{\text{FSS}} = j\sqrt{\frac{L}{C}}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \quad (2.2)$$

where Z_{FSS} is the impedance of the FSS layer, L is the equivalent inductance, C is the equivalent capacitance, ω is the angular frequency and $\omega_0 \left(= 1/\sqrt{LC}\right)$ is the resonant angular frequency.

In the analysis of FSS using the transmission line concept, the admittance expression is much preferred and Eqn. (2.1) can be expressed as:

$$Y_{\text{FSS}} = -\frac{j}{\left(\omega L - \frac{1}{\omega C}\right)} \quad (2.3)$$

while Eqn. (2.2) can be re-written as:

$$Y_{\text{FSS}} = -j\frac{1}{\sqrt{\frac{L}{C}}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} = -j\beta_{\text{FSS}} \quad (2.4)$$

where β_{FSS} is the equivalent susceptance.

Figure 2.1. Equivalent circuit for bandstop FSS.

Figure 2.2. Determination of effective admittance and reflection coefficient

Considering the FSS layer with air in the background (Figure 2.2), the effective admittance when viewed in a transmission-line model can be expressed as:

$$Y_{\text{eff}} = Y_0 + Y_{\text{FSS}} \quad (2.5)$$

Substituting Eqn. (2.3) into Eqn. (2.5) results in:

$$Y_{\text{eff}} = Y_0 - \frac{j}{\left(\omega L - \frac{1}{\omega C}\right)} \quad (2.6)$$

and substituting Eqn. (2.4) into Eqn. (2.5) results in:

$$Y_{\text{eff}} = Y_0 - j\beta_{\text{FSS}} \quad (2.7)$$

where Y_0 is the free space admittance. The reflection coefficient seen by a wave propagating towards the structure can be expressed as:

$$\Gamma = \frac{Y_0 - Y_{\text{eff}}}{Y_0 + Y_{\text{eff}}} \quad (2.8)$$

Substituting Eqn. (2.6) into Eqn. (2.8) results in:

$$\Gamma = -\frac{1}{1 + j2Y_0\left(\omega L - \frac{1}{\omega C}\right)} \quad (2.9)$$

The reflected power can be written as:

$$|\Gamma|^2 = \frac{1}{1 + 4Y_0^2\left(\omega L - \frac{1}{\omega C}\right)^2} \quad (2.10)$$

The half-power (3dB) can then be expressed as:

$$|\Gamma_{\text{3dB}}|^2 = \frac{1}{2} \quad (2.11)$$

Substituting Eqn. (2.11) into Eqn. (2.10) results in,

$$1 + 4Y_0^2 \left(\omega_{\pm} L - \frac{1}{\omega_{\pm} C} \right)^2 = 2 \quad (2.12)$$

where $\omega_{\pm}$ are the corner frequencies at the edges of the bandstop bandwidth. Further simplification results in:

$$2Y_0 \left(\omega_{\pm} L - \frac{1}{\omega_{\pm} C} \right) = \pm 1 \quad (2.13)$$

$$\omega_{\pm}^2 LC - 1 = \pm \frac{C}{2Y_0} \omega_{\pm} \quad (2.14)$$

$$\left(\frac{\omega_{\pm}}{\omega_0} \right)^2 - 1 = \pm \frac{Y}{2Y_0} \left(\frac{\omega_{\pm}}{\omega_0} \right) \quad (2.15)$$

$$\left(\frac{\omega_{\pm}}{\omega_0} \right)^2 - 1 = \pm \frac{1}{Q} \left(\frac{\omega_{\pm}}{\omega_0} \right) \quad (2.16)$$

where $Q = \frac{2Y_0}{Y} = \frac{\omega_0}{BW_{3dB}}$ is the quality factor and BW_{3dB} is the 3-dB bandwidth. When

plotting Eqn. (2.16), the term on the left-hand side corresponds to a curve crossing the vertical axis at -1 , while the right-hand side term corresponds to two straight lines in opposite sloping directions as shown in Figure 2.3.

Figure 2.3. Graphical representation of Eqn. (2.16).

The intersection of the two straight lines $\pm \frac{1}{Q} \left(\frac{\omega}{\omega_0} \right)$ with the quadratic curve $\left(\frac{\omega}{\omega_0} \right)^2 - 1$

will establish the corner frequencies (refer Figure 2.3) and these two corner frequencies can be expressed as:

$$\left(\frac{\omega_-}{\omega_0}\right) = \frac{1}{2Q} \left[\sqrt{1 + 4Q^2} - 1 \right] \quad (2.17)$$

$$\left(\frac{\omega_+}{\omega_0}\right) = \frac{1}{2Q} \left[1 + \sqrt{1 + 4Q^2} \right] \quad (2.18)$$

Taking the product of Eqns. (2.17) and (2.18) results in:

$$\left(\frac{\omega_-}{\omega_0}\right) \left(\frac{\omega_+}{\omega_0}\right) = 1 \quad (2.19)$$

$$\omega_0 = \sqrt{\omega_- \omega_+} \quad (2.20)$$

Eqn. (2.20) implies that the resonant frequency is the geometric mean of the two corner frequencies.

Substituting Eqn. (2.5) into Eqn. (2.8) results in:

$$\Gamma = \frac{j\beta_{\text{FSS}}}{2Y_0 - j\beta_{\text{FSS}}} \quad (2.21)$$

$$\Gamma = \frac{1}{\frac{2Y_0}{j\beta_{\text{FSS}}} - 1} \quad (2.22)$$

$$\Gamma = -\frac{1}{1 + j2Y_0 \sqrt{\frac{L}{C}} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)} \quad (2.23)$$

$$\Gamma = -\frac{1}{1 + jQ \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)} \quad (2.24)$$

$$\Gamma = -\frac{1}{1 + j\zeta} \quad (2.25)$$

where $\zeta = Q \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)$ is the transparency of the FSS and

$Y = \sqrt{\frac{C}{L}}$ is the admittance of the stand-alone bandpass FSS.

At resonance where $\omega = \omega_0$, the transparency parameter will be zero, indicating no power can be transmitted through the FSS. Also, the reflection coefficient will return a value of -1 (i.e. with a magnitude of 1 and a phase of 180°) and when plotted on a Smith Chart, this corresponds to a short circuit condition. The phase for the reflection coefficient of the single layer FSS can further be simplified from Eqn. (2.25) as:

$$\angle \Gamma = \pi - \tan^{-1} Q \left(\frac{f}{f_0} - \frac{f_0}{f} \right) \quad (2.26)$$

Figure 2.4 illustrates the relationship between the phase of the reflection coefficient and frequency. The plot shows the bandpass FSS resonating at 590GHz with the phase equal to 180 degree.

Figure 2.4. Typical trend for the phase of the reflection coefficient of a bandstop FSS. In this case, a FSS resonating at 590GHz with Q of 1.5526 is calculated.

2.2.2 Reflection Properties for Single Layer Bandpass FSS

The equivalent circuit for the bandpass FSS is shown in Figure 2.5. The parallel configuration of the inductor and capacitor differs from that of the bandstop FSS. The admittance for such an arrangement is expressed as:

$$Y_{\text{FSS}} = j\omega C + \frac{1}{j\omega L} \quad (2.27)$$

$$Y_{\text{FSS}} = j\sqrt{\frac{C}{L}}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) \quad (2.28)$$

$$Y_{\text{FSS}} = jY\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) = j\beta_{\text{FSS}} \quad (2.29)$$

The effective admittance (Figure 2.6) consisting of a parallel combination of the FSS admittance Y_{FSS} and the free space admittance Y_0 is given as:

$$Y_{\text{eff}} = Y_{\text{FSS}} + Y_0 \quad (2.30)$$

Following Eqn. (2.5) and Eqn. (2.6), the reflection coefficient for single layer bandpass FSS can be expressed as:

$$\Gamma = \frac{-j\beta_{\text{FSS}}}{2Y_0 + j\beta_{\text{FSS}}} \quad (2.31)$$

$$\Gamma = \frac{-j\left(\frac{1}{2Y_0}\right)\beta_{\text{FSS}}}{1 + j\frac{1}{2Y_0}(\beta_{\text{FSS}})} \quad (2.32)$$

$$\Gamma = \frac{-j\left(\frac{Y}{2Y_0}\right)\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}{1 + j\frac{Y}{2Y_0}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} \quad (2.33)$$

$$\Gamma = \frac{-j(Q)\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}{1 + jQ\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} \quad (2.34)$$

$$\Gamma = \frac{-j\xi}{1 + j\xi} \quad (2.35)$$

where ξ is opaqueness. The phase for the reflection coefficient of a single layer bandpass FSS can then be obtained as:

$$\angle \Gamma = \begin{cases} \frac{\pi}{2} - \tan^{-1} Q \left(\frac{f}{f_0} - \frac{f_0}{f} \right) & \text{for } \left(\frac{f}{f_0} - \frac{f_0}{f} \right) > 0 \\ -\frac{\pi}{2} - \tan^{-1} Q \left(\frac{f}{f_0} - \frac{f_0}{f} \right) & \text{for } \left(\frac{f}{f_0} - \frac{f_0}{f} \right) \leq 0 \end{cases} \quad (2.36)$$

Form Eqn. (2.19) and Eqn. (2.20), it can be observed that at resonance ($\omega = \omega_0$), the opaqueness parameter will be zero, indicating zero obstruction to power transmission through the FSS.

Figure 2.5. Equivalent circuit for bandpass FSS.

Figure 2.6. Determination of effective admittance and reflection coefficient.

2.2.3 Reflection Properties for Bandstop FSS with Dielectric Half-Space Loading

The physical structure for a single FSS with dielectric loading in the half-space is shown in Figure 2.7. In the equivalent circuit model, only the capacitance is affected by the dielectric loading while the inductance remains unchanged.

Figure 2.7. Single layer bandstop FSS with dielectric loading in the half-space.

The equivalent admittance Y_{FSS} of the FSS layer under the dielectric loading is given as:

$$Y_{FSS} = \frac{-j}{\sqrt{L} \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)} = -j\beta'_{FSS} \quad (2.37)$$

where C' and β'_{FSS} are the equivalent capacitance and susceptance of the FSS with the effect of dielectric loading respectively. The reflection coefficient seen from the dielectric towards the FSS and free space can be simplified from Eqn. (2.8):

$$\Gamma = \frac{Y_1 - (Y_0 + Y_{FSS})}{Y_1 + (Y_0 + Y_{FSS})} \quad (2.38)$$

Substituting Eqn. (2.37) into Eqn. (2.38) and rearranging it results in:

$$\Gamma = \frac{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2}{(Y_1 - Y_0)^2 - (2Y_1)j\beta'_{FSS} - (\beta'_{FSS})^2} \quad (2.39)$$

Dividing the numerator and denominator by the numerator yields:

$$\Gamma = \frac{1}{\frac{(Y_1 - Y_0)^2 - (\beta'_{FSS})^2}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2} - j \frac{(2Y_1)\beta'_{FSS}}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2}} \quad (2.40)$$

$$\Gamma = \frac{-1}{1 - \frac{(Y_1 - Y_0)^2 - (\beta'_{FSS})^2}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2} - 1 + j \frac{(2Y_1)\beta'_{FSS}}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2}} \quad (2.41)$$

$$\Gamma = \frac{-1}{1 - \frac{2Y_1^2 - 2Y_1Y_0 - j(2Y_1)\beta'_{FSS}}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2}} \quad (2.42)$$

$$\Gamma = \frac{-1}{1 - \left[\frac{(-j) \left[j(2Y_1^2 - 2Y_1Y_0) + 2Y_1\beta'_{FSS} \right]}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2} \right]} \quad (2.43)$$

$$\Gamma = \frac{-1}{1 + j \frac{2Y_1}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2} \left(\beta'_{FSS} + j(Y_1 - Y_0) \right)} \quad (2.44)$$

$$\Gamma = -\frac{1}{1 + j\zeta'} \quad (2.45)$$

where $\zeta' = \frac{2Y_1}{(Y_1 - Y_0)^2 + (\beta'_{FSS})^2} \left(\beta'_{FSS} + j(Y_1 - Y_0) \right)$ is the transparency for the entire dielectric-loaded FSS structure,

$Y_0 = \frac{1}{377}$ is the free space admittance,

Y_1 is the wave admittance in the dielectric and

$\sqrt{\epsilon_1}$ is the relative permittivity in medium 1 (as shown in Figure 2.7).

The transparency expression can further be simplified and expressed as:

$$\zeta' = \frac{2Y_1}{\beta_{\text{FSS}} - j(Y_1 - Y_0)} \quad (2.46)$$

It can be observed from Eqn. (2.45) that if $\zeta = 0$, the reflection coefficient will be -1 (i.e. with a magnitude of 1 and a phase of 180°), indicating a short circuit. If ζ equals to infinity, the reflection coefficient will be zero, indicating a perfect transmission.

From Eqn. (2.45), it shows that in the case for a FSS loaded with dielectric in the half-space, the reflection coefficient takes the same form as the reflection coefficient for a free-standing FSS as described in Chapter 2.2.1, except that the transparency is a complex value in this case. For the purpose of purely determining the phase of the reflection coefficient, a simpler equation can be used. Following Eqn. (2.37) and Eqn. (2.38), the reflection coefficient for the FSS with dielectric in the half-space can be expressed as:

$$\Gamma = \frac{Y_1 - Y_0 + j \frac{1}{Y' \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)}}{Y_1 + Y_0 - j \frac{1}{Y' \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)}} \quad (2.47)$$

$$\Gamma = \frac{\sqrt{\epsilon_1} Y_0 - Y_0 + j \frac{2Y_0}{Q \left(\frac{f}{f_0} - \frac{f_0}{f} \right)}}{\sqrt{\epsilon_1} Y_0 + Y_0 - j \frac{2Y_0}{Q \left(\frac{f}{f_0} - \frac{f_0}{f} \right)}} \quad (2.48)$$

where $Q = \frac{2Y_0}{Y'}$ is the 3-dB bandwidth of the FSS with dielectric loading and

$Y' = \sqrt{\frac{L}{C}}$ is the admittance of the FSS.

This equation is verified by plotting the phase of the reflection coefficient of a FSS resonating at 590GHz with $Q=5.55263$ and the substrate having dielectric constant of 3.78 as shown in Figure 2.8. It agrees well with Figure 3 in an earlier publication [46].

Figure 2.8. Phase of the reflection coefficient for bandstop FSS (with dielectric in the half-space) resonating at 590GHz (comparing with Figure 3 of [46]).

The reflection coefficient for the TM case is then given as:

$$\Gamma_{TM} = \frac{Y_{ITM} - Y_{0TM} + j \frac{2Y_{0TM}}{Q \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)}}{Y_{ITM} + Y_{0TM} - j \frac{2Y_{0TM}}{Q \left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)}} \quad (2.49)$$

where $Y_{ITM} = \sqrt{\epsilon_1} Y_0 \left(\frac{1}{\cos \theta_{ITM}} \right)$ is the wave admittance in medium 1 and

$Y_{0TM} = Y_0 \left(\frac{1}{\cos \theta_{0TM}} \right)$ is the wave admittance in free space.

2.2.4 Reflection Properties for Bandpass FSS with Dielectric Half-Space Loading

For the bandpass FSS with dielectric loading in the half-space, the admittance is given as:

$$Y_{\text{FSS}} = j\sqrt{\frac{L}{C'}}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) = j\beta'_{\text{FSS}} \quad (2.50)$$

where C' and β'_{FSS} are the equivalent capacitance and susceptance of the FSS with the effect of dielectric loading respectively. For an incoming wave at normal incidence angle, the reflection coefficient seen from the dielectric towards the FSS and free space can be simplified from Eqn. (2.8) as:

$$\Gamma = \frac{Y_1 - Y_0 - j\frac{1}{Q}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}{Y_1 + Y_0 + j\frac{1}{Q}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} \quad (2.51)$$

where $Q = \frac{2Y_0}{Y'}$ is the 3-dB bandwidth of the FSS with dielectric loading,

$Y' = \sqrt{\frac{L}{C'}}$ is the admittance of the FSS,

$Y_1 = \sqrt{\epsilon_1} Y_0$ is the wave admittance in the dielectric and

$\sqrt{\epsilon_1}$ is the relative permittivity in medium 1.

In the case with a scanned TE incidence, the reflection coefficient is modified to:

$$\Gamma_{\text{TE}} = \frac{Y_{\text{ITE}} - Y_{\text{0TE}} - j\frac{1}{Q}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}{Y_{\text{ITE}} + Y_{\text{0TE}} + j\frac{1}{Q}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)} \quad (2.52)$$

where $Y_{\text{ITE}} = \sqrt{\epsilon_1} Y_0 (\cos \theta_{\text{ITE}})$ is the wave admittance in medium 1,

$Y_{\text{0TE}} = Y_0 (\cos \theta_{\text{0TE}})$ is the wave admittance in free space and

$$\theta_{\text{ITE}} = \sin^{-1}\left(\frac{\sin \theta_{\text{0TE}}}{\sqrt{\epsilon_r}}\right).$$

2.2.5 Reflection Properties for Double Layer Bandstop FSS with Dielectric Spacer

When another spacer with a thickness of d is added onto the other side of the bandstop FSS as shown in Figure 2.9, the wave admittance Y' can be written as [41], [42]:

$$Y' = Y_1 \frac{Y_{\text{eff}} \cos \beta d + jY_1 \sin \beta d}{Y_1 \cos \beta d + jY_{\text{eff}} \sin \beta d} \quad (2.53)$$

With the further addition of another FSS layer, the overall effective admittance (refer to Figure 2.10) can be expressed as:

$$Y'_{\text{eff}} = Y_{\text{FSS}} + Y' \quad (2.54)$$

The reflection coefficient will then be expressed as:

$$\Gamma = \frac{Y_3 - Y'_{\text{eff}}}{Y_3 + Y'_{\text{eff}}} \quad (2.55)$$

Figure 2.9. Equivalent circuit for bandpass FSS.

Figure 2.10. Determination of effective admittance and reflection coefficient.

2.3 Variation of Admittance with Scanned Incidence

In relating the wave admittance to the phase constant β_0 , the permeability of free space μ_0 , angular frequency $\omega=2\pi f$ and the angle of incidence θ_0 , a set of more detailed wave admittance for TE and TM cases are given below. Firstly, the wave admittance for TE case in free space is given as:

$$Y_{oTE} = \frac{\beta_0}{\omega\mu_0} \cos\theta_{oTE} = \frac{\frac{\omega}{c}}{\omega\mu_0} \cos\theta_{oTE} = \frac{1}{c\mu_0} \cos\theta_{oTE} \quad (2.56)$$

where c is the velocity of light. Since $c = \frac{\omega}{k} = \frac{1}{\sqrt{\mu_0\epsilon_0}}$, the wave admittance for the TE

case can be also be expressed as:

$$Y_{oTE} = \frac{\sqrt{\mu_0\epsilon_0}}{\sqrt{\mu_0\mu_0}} \cos\theta_{oTE} = \frac{\sqrt{\epsilon_0}}{\sqrt{\mu_0}} \cos\theta_{oTE} \quad (2.57)$$

The relationship between the angle of incidence θ_0 in the free space and the scanned angle θ_1 inside the medium with a permittivity of ϵ_r is related by Snell's Law as:

$$\sqrt{\epsilon_0} \sin\theta_0 = \sqrt{\epsilon_0\epsilon_1} \sin\theta_1 \quad (2.58)$$

$$\sin\theta_0 = \sqrt{\epsilon_1} \sin\theta_1 \quad (2.59)$$

An alternative expression relating the angle of incidence θ_0 in the free space to the scanned angle θ_1 in the medium is given as:

$$\cos\theta_1 = \sqrt{1 - \sin^2\theta_1} = \sqrt{1 - \frac{\sin^2\theta_0}{\epsilon_1}} = \sqrt{\frac{\epsilon_1 - \sin^2\theta_0}{\epsilon_1}} \quad (2.60)$$

The wave admittance inside the medium can now be written as:

$$Y_{1TE} = \sqrt{\frac{\epsilon_0}{\mu_0}} \sqrt{\epsilon_1} \cos\theta_{1TE} = \sqrt{\frac{\epsilon_0}{\mu_0}} \sqrt{\epsilon_1 - \sin^2\theta_{oTE}} \quad (2.61)$$

For TM case, the free space wave admittance is given as:

$$Y_{0TM} = \frac{\epsilon_0}{\sqrt{\mu_0 \epsilon_0} \cos \theta_{oTM}} = \sqrt{\frac{\epsilon_0}{\mu_0}} \frac{1}{\cos \theta_{oTM}} \quad (2.62)$$

Similarly, using Eqn. (2.60) to relate the scanned angle in the dielectric medium θ_i with the scanned angle in free space θ_o , the wave admittance in medium 1 can be written as:

$$Y_{1TM} = \sqrt{\frac{\epsilon_0}{\mu_0}} \sqrt{\epsilon_1} \frac{1}{\cos \theta_{oTM}} = \sqrt{\frac{\epsilon_0 \epsilon_1}{\mu_0}} \frac{1}{\left(\sqrt{\frac{\epsilon_1 - \sin^2 \theta_{oTM}}{\epsilon_1}} \right)} \quad (2.63)$$

$$Y_{1TM} = \sqrt{\frac{\epsilon_0}{\mu_0}} \frac{\epsilon_1}{\sqrt{\epsilon_1 - \sin^2 \theta_{oTM}}} \quad (2.64)$$

It can be observed that the bandwidth is dependent on the angle of incidence. For TM propagation where the magnetic field is normal (or transverse) to the plane of incidence, the admittance and hence the bandwidth varies by a factor of $1/\cos \theta$, with $\theta = 0$ degree corresponding to the normal incidence. For TE propagation, the admittance and hence bandwidth varies by a factor of $\cos \theta$. Due to this variation, it is necessary to incorporate scan stabilisation features into the FSS design [59].

2.4 Fabry-Perot Interferometer (FPI) Approach of Analysis and Design [46]

This section is a review of the FPI technique developed by the other researchers [46]-[49] for the analysis of multi-layer FSS. A program was developed following their formulation for the band-stop FSS and its results were compared to those published in [46]. The program is later modified for the design of band-pass FSS and its accuracy was verified with the measured results of an actual prototype.

The FPI method is useful for the design of multiples of double-layer FSS. These double-layered FSS are analogous to the two reflecting surfaces of the FPI instrument used in optical systems. In between the two surfaces, it is filled with dielectric having thickness

of s and relative permittivity of ϵ_r (Figure 2.11). The propagation constant between the dielectric slabs for an incident wave at an angle of θ is given by [50]:

$$\beta = \frac{2\pi}{\lambda} \sqrt{\epsilon_r - \sin^2 \theta} \quad (2.65)$$

where λ is the free space wavelength.

Figure 2.11. FSS layers in an FPI arrangement [46].

At the interface of the FSS, the transmission and reflection coefficients are respectively given as [46]:

$$\tau_i^{-(+)}(\omega) = T_i^{-(+)}(\omega) e^{j\theta_i^{-(+)}(\omega)} \quad (2.66)$$

$$\rho_i^{-(+)}(\omega) = R_i^{-(+)}(\omega) e^{j\phi_i^{-(+)}(\omega)} \quad (2.67)$$

where $T_i^{-(+)}$ and $R_i^{-(+)}$ are the respective magnitude components;

$\phi_i^{-(+)}$ and $\theta_i^{-(+)}$ are the phases,

sign $-(+)$ as identifying coefficients of incident waves travelling from $-(+)$ z and

with i representing the i th number of interface.

The wave matrix representation for the double layer FSS structure can be expressed as [46]:

$$\begin{bmatrix} b^- \\ a^- \end{bmatrix} = \frac{1}{\tau_1^- \tau_2^-} \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} \begin{bmatrix} b^+ \\ 0 \end{bmatrix} \quad (2.68)$$

with the transfer matrix written as [46]:

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} 1 & -\rho_1^+ \\ \rho_1^- & \Delta_1 \end{bmatrix} \begin{bmatrix} e^{j\beta s} & 0 \\ 0 & e^{-j\beta s} \end{bmatrix} \begin{bmatrix} 1 & -\rho_2^+ \\ \rho_2^- & \Delta_2 \end{bmatrix} \quad (2.69)$$

The term $\begin{bmatrix} e^{j\beta s} & 0 \\ 0 & e^{-j\beta s} \end{bmatrix}$ is a standard transmission line matrix while the parameter s is the physical spacing between the two FSS. The transfer matrix can further be simplified as:

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} e^{j\beta s} & (-\rho_1^+)e^{-j\beta s} \\ (\rho_1^-)e^{j\beta s} & (\Delta_1)e^{-j\beta s} \end{bmatrix} \begin{bmatrix} 1 & -\rho_2^+ \\ \rho_2^- & \Delta_2 \end{bmatrix} \quad (2.70)$$

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} e^{j\beta s} - \rho_1^+ \rho_2^- e^{-j\beta s} & -\rho_2^+ e^{j\beta s} - \rho_1^+ \Delta_2 e^{-j\beta s} \\ \rho_1^- e^{j\beta s} + \rho_2^- \Delta_1 e^{-j\beta s} & -\rho_1^- \rho_2^+ e^{j\beta s} + \Delta_1 \Delta_2 e^{-j\beta s} \end{bmatrix} \quad (2.71)$$

where $\Delta_{l(2)} = \tau_{l(2)}^- \tau_{l(2)}^+ - \rho_{l(2)}^- \rho_{l(2)}^+$. It should be noted that the term for T_{21} and T_{22} differs slightly (in the +/- sign) with that published in [46].

The transmission coefficient for the double layer FSS is given as [46]:

$$\tau_{FPI} = \frac{b^+}{b^-} = \frac{\tau_1^- \tau_2^- e^{-j\beta s}}{1 - \rho_1^+ \rho_2^- e^{-j2\beta s}} \quad (2.72)$$

Assigning variable A and B to the numerator and denominator of the expression in Eqn.

(2.73) resulted in:

$$A = \tau_1^- \tau_2^- e^{-j\beta s} = T_1^+ e^{j\theta_1(+)} T_2^- e^{j\theta_2(+)} e^{-j\beta s} \quad (2.73)$$

$$A^* = \tau_1^- \tau_2^- e^{+j\beta s} = T_1^+ e^{-j\theta_1(+)} T_2^- e^{-j\theta_2(+)} e^{+j\beta s} \quad (2.74)$$

$$B = 1 - \rho_1^+ \rho_2^- e^{-j2\beta s} = 1 - R_1^+ e^{j\phi_1(+)} R_2^- e^{j\phi_2(+)} e^{-j2\beta s} \quad (2.75)$$

$$B^* = 1 - R_1^+ e^{-j\phi_1(+)} R_2^- e^{-j\phi_2(+)} e^{+j2\beta s} \quad (2.76)$$

$$B.B^* = 1 - [2R_1^+ R_2^+ \cos(\varphi_1 + \varphi_2 - 2\beta s)] + (R_1^+ R_2^+)^2 \quad (2.77)$$

The power of the transmission coefficient t_{FPI} can be obtained by multiplying the transmission coefficient τ_{FPI} with its conjugate. Taking into account of Eqn. (2.73) to Eqn. (2.77), it can be expressed as:

$$t_{\text{FPI}} = \tau_{\text{FPI}} \tau_{\text{FPI}}^* = \frac{(T_1^- T_2^-)^2}{1 - 2R_1^+ R_2^+ \cos \Phi + (R_1^+ R_2^+)^2} \quad (2.78)$$

where

$$\Phi = -\varphi_1^+(\omega) - \varphi_2^-(\omega) + 2\beta s \quad (2.79)$$

is the total FSS insertion (S21) phase. This total insertion (S21) phase within the dielectric slab consists of the path length s and the phase of the reflection coefficient from the two surfaces. To consider the power of the transmission coefficient with respect to transmission from the internal faces of the interferometer into the free space, Eqn. (2.78) can be written as [46]:

$$t_{\text{FPI}} = \left(\frac{Y_0}{Y_1} \right)^2 \frac{(T_1^+ T_2^-)^2}{1 - 2R_1^+ R_2^+ \cos \Phi + (R_1^+ R_2^+)^2} \quad (2.80)$$

where Y_1 and Y_0 are the wave admittance in the dielectric and free space respectively.

For maximum power transmission coefficient, $\cos(\Phi)$ equals to one and Eqn. (2.80) becomes:

$$t_{\text{FPI}} = \left(\frac{Y_0}{Y_1} \right)^2 \frac{(T_1^+ T_2^-)^2}{(1 - R_1^+ R_2^+)^2} \quad (2.81)$$

If the FSS on both layers are identical (i.e. $R_1^+ = R_2^- = R$, $T_1^+ = T_2^- = T$), it can be further simplified as:

$$t_{\text{FPI}} = \left(\frac{Y_0}{Y_1} \right)^2 \frac{(T)^4}{(1 - R^2)^2} \quad (2.82)$$

Maximum transmission gives rise to a phase shift condition of $\Phi = 2\pi n$, where n is a positive integer. Equating this phase relationship with Eqn. (2.79) and substituting Eqn. (2.65) into Eqn. (2.79) results in [46]:

$$\frac{2\pi f}{c} s \sqrt{\epsilon_r - \sin^2 \theta} - n\pi = \frac{\Phi_1^+ + \Phi_2^-}{2} \quad (2.83)$$

The terms on the right hand side of Eqn. (2.83) represents the average phase curve for the reflection coefficient of the two FSS. The term on the left represents the relative phase delay due to the path difference between the two FSS. It is a set of straight lines when plotted against frequency. The graphs representing the two terms in Eqn. (2.83) are plotted in Figure 2.12. The intersection between the two plots yields maximum power transmission coefficient. This can be further illustrated with Figure 2.13 and Figure 2.14. The intersection points (A and B) in Figure 2.13 correspond to the peak transmission shown in Figure 2.14. The FSS element in this case is a slotted tripole with a total perimeter of 13.99mm and inter-element spacing of 4.167mm. A substrate with dielectric constant of 1.7 and thickness of 0.32cm is placed between the two identical FSS layers.

Figure 2.12. A combined plot showing the average phase of the reflection coefficient of two FSS and a family of straight lines representing the path length between the two FSS.

Figure 2.13. Determining the intersections between the plot for the average phase of the reflection coefficient of the two FSS and the plot for the path length between the two FSS.

Figure 2.14. Maximum transmission of double layer band-pass FSS corresponds closely to the intersection points shown in Figure 2.13.

Figure 2.15. Double layer FSS with tripole element [25].

2.5 Grating Lobe Consideration for FSS [25]

The phenomenon of grating lobe is very critical in the design and performance of the bandpass FSS [25]. Grating lobe appears when the phase difference between a neighbouring element is equal to a multiple of 2π . Referring to Figure 2.16, for a physical inter-element spacing of D_x , the grating lobe direction η_g for a given incidence wave at angle η (with respect to the normal) is given as [25]:

$$\beta D_x (\sin \eta + \sin \eta_g) = 2\pi n \quad (2.84)$$

where $\beta = \frac{2\pi}{\lambda_g}$, $n = 1, 2, \dots$

Hence, the grating lobe frequency, f_{GL_air} , can be determined from:

$$f_{GL_air} = \frac{c_0}{\lambda_g} = \frac{nc_0}{D_x (\sin \eta + \sin \eta_g)} \quad (2.85)$$

The frequency for the onset ($\eta_g = 90^\circ$) of the first grating lobe, f_{GL_air} is given by [25]:

$$f_{GL_air} = \frac{c_0}{D_x (\sin \eta + 1)} \quad (2.86)$$

whereby c_0 is the velocity of light. For arrays loaded with dielectric layers of relative dielectric constant ϵ_r , the onset of grating lobe in the dielectric will occur at a lower frequency given by [25]:

$$f_{\text{GL_trap}} = \frac{c_0}{D_x (\sin \eta + \sqrt{\epsilon_r})} \quad (2.87)$$

It can be seen that although the onset of free space grating lobe can be delayed by reducing the inter-element spacing through dielectric loading, but it will trigger an earlier onset of trapped grating lobe and this might pose problems for finite FSS array [59], a phenomenon that is also observed in the results presented in the subsequent chapters.

Figure 2.16. Top: specular reflection and transmission; bottom: onset of grating lobe [25].

2.6 FEM CAD Approach of Analysis and Design

2.6.1 Introduction

In this section, the design of band-pass FSS using the commercial finite element (FEM) software (Ansoft HFSS) [53] is presented [51], [52]. This package had shown its usefulness in the analysis and design of photonic bandgap structures [60]. For that application, HFSS was used to determine the propagation constant of the surface wave that propagates along the array's planar surface. In this chapter, the transmission coefficient of an incident wave propagating through the FSS structures is determined instead. The creation of the simulation model, interpretation and analysis of the outcome and comparison with experimental results are presented for the square slot and the square loop slot band-pass FSS. The wave propagation through the FSS that are otherwise difficult to quantify can be visualized using this commercial CAD tool.

2.6.2 The Finite Element Method Implementation in Ansoft HFSS [53], [61]-[63]

Ansoft HFSS employs the finite element method to generate the electromagnetic field solution. Unlike the moment method which requires Maxwell's equations to be expressed in integral form, the finite element method expresses Maxwell's equations in partial differential form. The advantages of the partial differential methods stem from their inherent geometric adaptability, low memory demand, and their capability to model heterogeneous (isotropic or anisotropic) media.

In Ansoft HFSS, the geometry is divided into large numbers of tetrahedral and each tetrahedron is basically a four-sided pyramid. An approximation is made to the field or potential in each tetrahedral element in the mesh. Firstly, the vector finite element [62] representation of fields is used to interpolate the value of a vector field inside each tetrahedron to either its tangential or normal components (refer to Figure 2.17).

Figure 2.17. Representation of field quantity in Ansoft HFSS [53].

Secondly, various interpolation methods (i.e. basis functions) are used to interpolate field values from nodal values. A first-order tangential element basis function interpolates field values from both nodal values at vertices and on edges, while a zero-order basis function makes use of nodal values at vertices only and hence assumes the field to vary linearly inside each tetrahedron.

Overall, the value of a vector field quantity (i.e electric and magnetic fields) at points inside each tetrahedral is obtained by a set of local approximation functions that satisfies both physical and mathematical requirements. The Galerkin's method is applied to approximate these analytical operator equations by matrix equations. Individual Galerkin matrix is constructed for each finite element and the matrices for all the elements are then added together to create a single matrix equation that is used to solve for the electromagnetic field across the entire device. Inherent with the finite element methods, the matrix equation is sparse, which implies that there are only a few nonzero values to be stored.

The accuracy of the method depends on both the type of approximation functions used in each element and on the numbers of elements used in the meshing. Ansoft HFSS uses adaptive mesh generation, whereby the mesh is automatically refined by the program via

the comparison of errors in the computed solution. Convergence is reached when the error falls within user-specified accuracy [62].

The wave equation being solved by HFSS takes the form of [53], [63]:

$$\nabla \times \left(\frac{1}{\mu_r} \nabla \times E \right) - k_0^2 \epsilon_r E = 0 \quad (2.88)$$

where $E(x, y, z)$ is a phasor representing an oscillating electric field,

$k_0 = \omega \sqrt{\mu_0 \epsilon_0}$ is the free space wave number,

μ_r is the complex relative permeability and

ϵ_r is the complex relative permittivity.

Preconditioned conjugate gradient algorithm is applied to solve the sparse matrices arising from the finite element method. As a result, the required time to obtain a solution increases linearly with respect to the numbers of unknown. This is much faster when compared to the moment methods, which, in general, produce full-matrix equations. The time taken for full matrices solution increases as the cube of the number of unknowns. However, it should be highlighted that for the finite element method, the mesh and the field solution exist everywhere in the problem domain, including the surrounding air space.

2.6.3 Square Slot Bandpass FSS Design

The application chosen for the first design is to allow electromagnetic (EM) wave to pass through the FSS at 9.25GHz. A square aperture slot is chosen as the FSS element for its simplicity in the design. The FSS elements, square aperture slots with dimensions of 15mm by 15mm shown in Figure 2.18, are printed on a 0.5mils substrate material with relative permittivity of 3.4 and loss tangent of 0.001. The elements are separated by an inter-element periodicity D_x of 30mm in both axes. For practical realization, the infinite theoretical FSS model is being approximated by a finite array with an overall size of 10 cm by 10cm.

2.6.4 Square Slot FSS Simulation Model Setup

A single FSS square slot element is drawn in the centre of the unit cell as shown in Figure 2.18. The unit cell is defined by two orthogonal sets of Master and Slave boundary conditions. The periodicity of the FSS array is represented by this set of so-called “linked boundary conditions”. The medium inside the boundary conditions has been filled with an air box. Two identical Perfectly-Matched-Layers (PML) are defined for the top and bottom of this air box as illustrated in Figure 2.19. This unique HFSS feature provides perfect absorption to the EM incoming wave at large incidence angle. A built-in macro “PMLmaterialsetup” is then used to generate the PML’s material properties, based on the frequency of operation and the impedance matching condition at the boundaries.

To excite the FSS element, an incident wave source having an electric field $\mathbf{E}$, propagating in the direction of the propagation constant $\mathbf{k}$ is being created (Figure 2.19). To investigate the performance of the FSS excited by incoming wave at different incident angle, a phase delay relationship between the Master and Slave boundaries corresponding to the relevant incident angle will be assigned. A total of 45897 tetrahedra meshes with a matrix size of 75546 are generated for the final iteration of the square patch band-pass FSS. To obtain the entire frequency response (2GHz-18GHz) of the FSS, different meshing together with different PML material properties will be created.

Figure 2.18. HFSS uses a unit cell with periodic boundary conditions to represent an infinite FSS array.

Figure 2.19. Simulation model setup.

2.6.5 Analysis of Square Slot FSS Performance

The wave phenomenon and the physics associated with the band-pass FSS can be visualised and explained using the animated field display option. Figure 2.20 illustrates the magnitude of the total electric field propagating through the square slot at the resonance (band-pass) frequency of 9.25GHz. The band-stop performance at other out-of-band frequencies, says 5GHz, is being shown in the scattered field plot as shown in Figure 2.21. Standing wave arises as a result of reflections of the incident wave hitting the FSS.

The induced electric field intensity and current on the surface of the FSS can be visualised in Figure 2.22. It can be seen that due to the discontinuity at the slot edges, charges accumulate resulting in edge singularity (red spot along the slots). From Gauss divergence theorem, $\nabla \cdot \mathbf{D} = \rho$, the high charge density $\mathbf{D}$ at the edge results in high flux density ρ (i.e. high field intensity). The rate of change of the flux density will create the displacement current as indicated by the arrows shown in Figure 2.22.

The frequency for the onset of grating lobes is calculated to be at 10GHz. This phenomenon is marked by the energy being drawn to the grating lobe directions that are at ± 90 degrees from the direction of incidence wave propagation (Figure 2.23). The grating lobes evolve when the electrical length between the adjacent element (along the Y-axis) reaches approximately one free space wavelength, resulting in a 360 degree (or 0 degree) in-phase addition (Figure 2.24) of plane wave grazing along the FSS in the grating lobe directions.

Above this frequency, the grating lobes will slowly lift off from the plane of the FSS array into the free space (Figure 2.25). With the occurrence of grating lobes, the simulated total field solutions for a unit cell will include the grating lobe field penetrating from the neighbouring cells (Figure 2.25). Compared with the actual practical measurement where only the specular field will be captured by the receive antenna, the simulated transmission response above the grating lobe frequency is deemed meaningless

unless there are means to de-embed the specular field component from the grating lobe contributions. As such, the simulated transmission response above this frequency (shown as dotted line in Figure 2.26) deviates from the measured transmission response. A higher simulated S21 is obtained as a result of grating lobe contribution from neighbouring cells. In the measured response, the grating lobe phenomenon is depicted by a dip at around 10.5GHz, indicating energy being drawn to the grating lobes. A measured 3-dB bandwidth of around 14.6% (8.07GHz to 9.42GHz) over the centre frequency of 9.25GHz is obtained. A higher pass-band transmission loss and an overall downward shift of 0.5GHz are observed from the measurement plot as compared to the simulated results. Both phenomena contributed to the otherwise insignificant minor differences observed in the overall results between the simulated FSS model and the inclusion of the thin dielectric backing in the actual FSS prototype.

Figure 2.20. Band-pass performance visualised from the total E-field plot.

Figure 2.21. Reflection of scattered field associated with out-of-band (band-stop) performance.

Figure 2.22. Field intensity and current on the FSS.

Figure 2.23. At 10GHz, the simulated field propagates in the sideward direction signifying grazing of energy along the FSS. This is a result of the first onset of grating lobes at $\pm 90^\circ$ from the broadside of the FSS.

Figure 2.24. In-phase addition of wave front in the grating lobe direction.

Figure 2.25. Total electric field distribution at 17GHz. Electric field propagating outwards into the grating lobes direction and inwards from the neighbouring grating lobes respectively.

Figure 2.26. Simulated and measured transmission response of band-pass FSS.

2.6.6 Square Loop Slot Band-pass FSS

The next structure analysed shows how grating lobes can be shifted to a higher frequency outside the range of interest (i.e. $>18\text{GHz}$). With a similar centre frequency of 9.25GHz , a square loop slot with a dimension of 7.6mm by 7.6mm ($\sim 0.23\lambda$) and a gap width of 0.3mm is formed in between two PEC conductors (Figure 2.27). The extremely small slot gap will ensure a uniform magnetic field distribution as compared to a wider one. For a wider slot, the magnetic current will flow closer to the inner square patch conductor and hence the effective perimeter will be smaller than the mean and this usually resulted in a higher than desired center frequency. The outer conductor is a 0.1mm wide metal strip with an overall dimension of 8mm by 8mm . A 50% reduction in the element length is achieved, compared to the previous design of a 15mm by 15mm square patch aperture. A closer inter-element spacing can thus be realised and this translates to a wider bandwidth with a delayed onset of grating lobes. At the centre frequency, a total of 36819 tetrahedras are generated with a matrix size of 224438 for the converge solution. A scan angle with $\phi=0$ and $\theta=-15$ degrees have been imposed between the linked boundary conditions (Figure 2.27). The simulated transmission performance of the FSS is shown in Figure 2.28. The transmission coefficient at the scan angle is obtained by first defining a line of integration along the scan angle direction and then extracting the magnitude of the field along it.

For practical realization, the FSS elements are printed on a thin dielectric sheet. However, due to the wideband response as compared to the previous narrow band design, the effect of the dielectric will be significant and hence must be included in the simulation model. Using the existing dimension of the square loop slot, it is loaded with a 5mils thick dielectric with a relative permittivity of 2.2 and loss tangent of 0.001. The simulated and measured response is shown in Figure 2.29. The center frequency shifts down to about 7.3GHz . A measured 3-dB bandwidth of around 61.2% (5.73GHz to 10.2GHz) is obtained over the center frequency of 7.3GHz . The simulated results indicate a better S21 performance at the higher frequency range when compared to the measured one. This could be due to the higher loss incurred from the prototype material as compared to a

constant loss figure used throughout the frequencies in the simulation. When compared to the previous design of a square aperture slot, the effect of dielectric is significant. This is due, firstly, to the thicker substrate used in the loop design; and, secondly, to the smaller opening of the slots in the loop as compared to the wider opening of the square aperture slot. For thinner slot opening, larger portion of the field resides inside the dielectric as compared to wider slots whereby the field actually fringes out of the material (Figure 2.30). The higher field concentration in the small openings of the loop resulted the higher dielectric loading effect.

Figure 2.27. Improved slotted-square loop FSS with angle of incidence = 15deg. from the vertical axis.

Figure 2.28. Simulated S21 performance of improved slotted-loop band-pass FSS (without substrate backing) for different angle of incidence.

Figure 2.29. S21 of band-pass FSS with dielectric support.

Figure 2.30. The effect of the slot width on the fringing field and hence the dielectric loading effect.

2.7 Conclusion

The equivalent circuit parameters have been used to approximate the admittance of the FSS sheet seen by an incident wave. The expressions for the reflection coefficient of the single layer FSS sheet and its subsequent material additions are derived from the transmission line model. The expressions are then rearranged to obtain the transparency figure, phase information and bandwidth (i.e. Q factor). The bandwidth is dependent on the angle of incidence. For TM propagation where the magnetic field is normal or transverse to the plane of incidence, the admittance and hence the bandwidth varies by a factor of $1/\cos\theta$, with $\theta = 0$ degree corresponding to the normal incidence. For TE propagation, the admittance and hence bandwidth varies by a factor of $\cos\theta$. The transmission line analogy will be extremely useful in the understanding and visualization of the FSS admittance and its transformation on the smith chart.

By grouping the FSS into pairs, an efficient approximation of its response can be obtained using the FPI analysis approach. A design example has been presented in this chapter whereby the accuracy of the FPI method is compared to the measured result of a

tripole element FSS. Good correlation is observed. The FPI hypothesis will also be useful in determining the thickness of the spacer used to separate the dual layer (Chapter 3.3) and quad-layer (Chapter 3.3) FSS. The field and current display provided by the FEM CAD offers a more comprehensible physical insight to the phenomena occurring at the pass-band, stop-band and onset of grating lobe condition. These information will be useful in the design and analysis of the convoluted FSS element presented in Chapter 3.2. Overall, the theory and physical insight presented in this chapter is useful in the explanation and analysis of the different FSS structures reported in the next chapter.

Chapter 3

Design of Periodic Arrays with Frequency Selective Response for Mobile Communication and GPS

3.1 Introduction

For most applications, in particularly FSS radomes [25], some of the typical requirements include low pass-band insertion loss, rapid roll-off outside the pass-band, delayed onset of grating lobes, stabilised pass-band response with respect to wave incidence angles and polarization. It should be highlighted that the electrical requirements for the FSS can differ for antennas that are similar, but deployed in different environments. In a different environment, the frequencies for the unwanted interference or hostile signal might change and this will result in different sets of out-of-band attenuation requirements for the FSS. In addition to the difference in electrical requirements, different environment present different mechanical constraints for the FSS installation. For applications with either space constraints or one that requires a curved filter surface, the complexity (e.g. fabrication and alignment) and thickness of a multi-layer FSS makes it less attractive compared to a single layer structure.

To allow signal to pass through at the lower frequency band (GPS and DCS1800), low-pass FSS are required and this can be realised either in terms of a low frequency bandpass design or a high frequency bandstop design. However, the low frequency bandpass FSS suffers several shortfalls. Firstly, transmission spikes due to higher order resonance will appear at the higher frequencies and these “transparent windows” will subject the antenna to unwanted interference or hostile signals at the higher frequencies. Secondly, since the element is designed to resonate at a low frequency, its element size will appear relatively larger at the higher frequencies and this will create early onset of grating lobes. The high frequency bandstop design will eliminate such problems and provide various degrees (bandwidth) of shielding depending on the design complexity.

In this chapter, five design variants of a low-pass FSS are presented. These FSS (single and multi-layer design) will produce various degree of roll-off while allowing the GPS and DCS1800 signal to pass through unattenuated. The electrical design, performance and trade-off with design/mechanical complexity will be addressed. The performance of these FSS will be measured using the conventional free space microwave set-up [64] in an anechoic chamber as depicted in Figure 3.1.

Figure 3.1. Conventional free space measurement setup [64].

3.2 Single Layer Bandpass Convolute Loop Design

3.2.1 Introduction and Design Configuration

Variants of the square loop FSS element had shown good performance in the high microwave frequency spectrum [65], [66] and hence an attempt is made to convolute such element for the present application. Figure 3.2 shows the geometry and the dimensions of the convoluted FSS loop element with four folded segments at the corners. Although the design looks like a variant of a cross, but there are subtle differences when compared to the others. For example [67], that element belongs to the “plate” [68] type as opposed to our present “loop” [68] type of design. For that design [67], the cross shape can be obtained by removing copper traces from a square plate while in our design, the cross shape is realised by bending the corners of the square slot loop. As explained earlier [68], the “plate” type suffer from unwanted variation of transmission response over

varying incidence angle and are prone to early onset of grating lobes. For Photonic Bandgap (PBG) application [67] (i.e. suppression of surface wave), the cross elements are made up of conductive patches as compared to slots for our present free space band-pass FSS application. By principle of Babinet’s duality, the magnetic field is responsible for the electric current on the conductive traces of the PBG cross element, while the electric field is responsible for the magnetic currents in our FSS slots.

The present design is etched on a 0.127mm thick copper-cladded substrate with relative permittivity of $\epsilon_r = 3.5$ and a loss tangent of 0.02. This FSS substrate layer is sandwiched between two 5mm thick foam layer with $\epsilon_r = 1.043$ and a loss tangent of 0.001. The structure is simulated and analysed using Ansoft DesignerTM, a full wave EM software package with inclusion of periodic boundary conditions. The simulated model assumes an infinite periodic array. However, for practical implementation, the prototype array has an overall dimension of 406mm by 305mm.

Figure 3.2. Convoluted FSS loop element (slots etched from copper -cladded substrate) arranged in rectangular lattice.

3.2.2 Experimental Results and Analysis

A zoom-in view of the simulated transmission response (Figure 3.3) shows that transmission loss of 0.5dB is observed between 1.25GHz to 2.125GHz, 1.15GHz to 2.05GHz, and 1.38GHz to 1.98GHz for the normal incidence, TM 45° incidence and TE 45° incidence, respectively. The simulated pass-band response centres at 1.7GHz. The measured transmission response depicting the entire pass-band and out-of-band rejection is shown in Figure 3.4a and Figure 3.4b for TE and TM polarizations respectively. Comparing the two plots, a measured 0.5dB transmission loss is obtained from 1.5GHz to 1.93GHz for TM 45° and 1.54GHz to 2.1GHz for TE 45° respectively. With the centre frequency at 1.735GHz, a common 0.5dB-bandwidth of 22.47% is still obtainable for the two extreme incidence angles (TE 45° and TM 45°). Both plots depict a relatively stable transmission response at the pass-band region (transmission loss at 0.5dB level) for the different wave incidence angles. The slight variation between the simulated and measured results might due to the finite size prototype and the measurement tolerances such as the high noise floor and the poor isolation at this relatively low frequency band.

Figure 3.3. A zoom-in view of the simulated pass-band transmission response of the array for different polarization (TE and TM) and different incidence angles.

The relatively wide band performance is achieved by the close packing and the wide slot width. From simulation, a slot width of 7mm and 8mm will yield a bandwidth of 17%

and 19.5% respectively. The slot width may be increased by either extending the outer boundary (L1) or reducing the inner boundary (L2) of the element. It should be noted that reducing the inner boundary increases the upper pass-band frequency while extending the outer boundary decreases the lower pass-band frequency.

Observing the out-of-band rejection performance in both TE (Figure 3.4a) and TM (Figure 3.4b) polarizations, a null is present at around 3.8GHz for the normal incidence (i.e. 0°) response. This can be explained by the fact that the length of the folded segment (i.e. L3 in Figure 3.2) is about a quarter-wavelength at 3.8GHz, analogous to that observed in a tripole element [69]. Considering the coupled transmission line analysis, the folded segment will behave like a stub whereby its effective load impedance approaches infinity at this frequency. This prevents current from flowing along the element. The computed contour of the magnetic current density across the slot (calculated by $\vec{M} = \vec{E} \times \hat{n}$) is shown in Figure 3.5a and Figure 3.5b for normal incidence at 3.8GHz and 1.7GHz respectively. In addition to the prominent null observed from the normal incidence response, various degrees of attenuation are observed for the other wave incidence angles (Figure 3.4). For efficient signal rejection, an overall transmission coefficient of less than -10dB is still obtainable between 3.4GHz to 4.2GHz for all the incidence angles in both polarizations (Figure 3.4).

For the present design, the onset of free space grating lobe for incidence angle of 45° is calculated [70] to be around 2.89GHz, 1.7 times beyond the centre pass-band frequency of 1.7GHz. For normal incidence, the grating lobe occurs at 4.934GHz. Between 2.45GHz and 3.4GHz, there are onsets of trapped [70] and free space grating lobes for the different scanned incidence conditions. These phenomena are associated with energy losses and are reflected from the non-gradual sloping of the transmission response observed between 2.45GHz and 3.4GHz (Figure 3.4). This frequency range is beyond the useful pass-band region and it should have minimal effect for our current application.

The increase in transmission response measured at the lower frequency end (0.7GHz - 0.8GHz) is due to the deteriorating isolation between the receiving and transmitting

antenna of a typical microwave free-space measurement set-up. This is not a physical phenomena associated with the design as can be seen from the simulated results where the element is designed to have its first fundamental resonance at around 1.7GHz.

Figure 3.4a. Measured transmission response (TE polarization) showing pass-band and out-of-band rejection for various incidence angles.

Figure 3.4b. Measured transmission response (TM polarization) showing pass-band and out-of-band rejection for various incidence angles.

3.2.3 Surface Reflection and Transmission Phase Analysis

The underlying physics for both transmission and reflection phenomena described above can be further analysed [71] by comparing the surface reflection phase, array transmission phase and their corresponding amplitude (Figure 3.6). At resonance frequency of 1.75GHz, the transmission magnitude is maximum, indicating minimal mismatch between the FSS and the air. The current along the slotted loop will be in-phase, leading to maximum current flow and hence maximum energy transmitted. In other words, the slot FSS is almost transparent at resonance and the phase of the transmitted wave exiting the FSS will be similar to the phase of the incidence wave impinging the FSS, leading to zero transmission phase difference. The corresponding reflection coefficient will be minimal (approaches zero) and at the point where it is exactly zero, the phase for the reflected wave will encounter a phase transition from -90° to $+90^\circ$. When the transmission coefficient becomes minimal (i.e. reflection is maximal), the array behaves like a metal plate and the phase for the reflected wave is at 180° (i.e. travelling in the opposite direction) with respect to the incident wave while the transmission phase will similarly encounter a phase transition from 90° to 270° as the transmission coefficient approaches zero. It should be noted that the phase information is calculated based on a single FSS surface without the substrate materials. A shift in the phase due to extra delays in the material will be encountered if the materials were added in the simulation model.

In summary, a unique engineering methodology has been applied to obtain selective signal reflection via selecting the length of the convoluted segments to be approximately a quarter-wavelength at the frequency of rejection. Also, stability in the pass-band transmission response is obtained with the convoluted cell size. In order to satisfy a challenging pass-band bandwidth of 22.47% (1.54GHz to 1.93GHz), consideration of grating lobes excitation together with the optimisation of the slot width have been factored in the design of this new convoluted geometry.

Figure.3.5. Computed magnetic surface current density (V mm^{-1}) in the slot of a single element at normal incidence:

- (a) 3.85GHz (stop-band), showing minimal current amplitude
- (b) 1.7GHz (pass-band), showing relatively higher current amplitude

Figure 3.6. Computed (a) surface reflection phase and transmission phase change, (b) reflection and transmission amplitude for a plane wave at normal incidence to the single layer FSS array.

3.3 Double Layer Bandstop Design with Ring Elements

The double layer FSS structure consists of two sheets of FSS separated by a foam spacer with thickness of 25.4mm (Figure 3.7). This thickness is about 0.487λ , where λ is the free space wavelength at 5.75GHz. The foam has a dielectric constant of about 1.043 and a loss tangent of 0.002. The FSS patterns are etched on standard polyimide film with dielectric constant of 3.5 and loss tangent of 0.0026. With an effective dielectric loading of 2.25 and a horizontal inter-element spacing of $D_x= 9.44\text{mm}$ ($D_x= 2 \times S_x$), the lowest frequency for the onset of trapped grating lobe occurs at around 16.56GHz for an incidence wave at 45° . As frequency increases, this grating lobe will propagate as free or trapped surface waves [72]. Following the formula given for a triangular lattice arrangement, there will then be an onset of free space grating lobe at around 21.4GHz for an incidence wave at 45° .

Figure 3.7. Double layer ring FSS with wide spacer.

3.3.1 Design and Analysis Using Smith Chart and Transmission Line Approach

The design and performance of the multi-layer FSS can be studied by considering the admittance of each FSS layer and its transformation along the transmission line when

dielectrics are added to it. It is well known that these multi-layer FSS can be modeled as a transmission line shunted by equivalent circuits in the form of lumped parallel or series capacitance C and inductance L . The L and C behaviour together with its reflection coefficient over a frequency span can be represented on an admittance chart [54]-[57]. The admittance chart in this case depicts a graphical relationship between the normalized free space admittance and the reflection coefficient at the outermost material interface.

Figure 3.8a depicts the simulated admittance locus for the entire frequency sweep of 1GHz to 10GHz. Five frequency points are marked for comparisons and they represent the pass-band at 1GHz (marker #1) and 2GHz (marker #2); the rejection bands at 5GHz (marker #3), 7GHz (marker #4) and 9GHz (marker #5). It is observed the admittance locus lie on the unity conductance circle (i.e. $G = 1$), indicating that the real part of the normalized admittance is equivalent to that of air (i.e. $1/377$ mhos). At the lower frequencies, most of the markers are lying on the bottom half of the admittance chart where the imaginary admittance value is positive (i.e. $+jB$). This indicates a capacitive nature of the FSS sheet when operating at the lower frequencies. It is also observed that the admittance at the lower frequencies lie close to the center of the admittance chart, corresponding to minimal reflection coefficient. The reflection coefficient is merely the distance from the center of the chart to the admittance location. As frequency increases, the admittance moves towards the edge (i.e. high reflection coefficient) and towards the left-hand side of the admittance chart. This corresponds well with the magnitude plot shown in Figure 3.8b where minimal reflectivity is observed between the pass-band of 1GHz to 2GHz and high reflectivity for the rejection band of 7GHz to 11GHz.

With the addition of a 25.4mm thick layer of spacer (Figure 3.9a), minimal changes is observed for the admittance at 1GHz. However, for 2GHz, the admittance is rotated by about 120° from -110° (Figure 3.8a) to 130° (Figure 3.9a) in a clockwise direction and is now spotted with a real value (conductance) of about 1.28 and an imaginary value (suceptance) of $-0.71j$. This is compared to an admittance of $1.02 + 0.661j$ for the same frequency with a single FSS sheet (Figure 3.8a). The addition of the spacer is equivalent to a movement along the transmission line and this causes a variation in the electrical

length. The amount of change in electrical length is equivalent to the angle rotated on the admittance chart. This will change both the real and imaginary part of the admittance. A VSWR of 1.925 and 1.903 is obtained for the admittance at 2GHz in Figure 3.9a and Figure 3.8a respectively. Ideally, the VSWR should be similar, as the additional spacer is equivalent to movement along a lossless transmission line. The phase of the reflection coefficient will change along the line but the magnitude remained constant. Hence, such movement should be represented by a rotation along the circumference of a constant radius circle (i.e. the constant VSWR circle). However, minute losses in the spacer material will result in the deviation from the constant radius circle.

It is also observed that the entire admittance locus is “uncoiled” in a clockwise direction. This is due to the fact that the length of the spacer is relatively longer compared to the wavelength at the higher frequencies and hence the angle of rotation for the admittance at the higher frequency will be larger compared to that at the lower frequencies. This also explains for the smaller variation in the admittance at 1GHz as compared to that at 2GHz with (Figure 3.9a) and without (Figure 3.8a) the foam inclusion. However, at the frequency of around 5.75GHz where the spacer is at approximately half-wavelength, the admittance will rotate by 360° and hence little variation of its position will be noticed with the addition of the spacer. The admittance at 7GHz (marker #4) and 9GHz (marker #5) is located near to the brim of the admittance chart. Since reflection coefficient is the distance from the center of the chart to the admittance location, this would imply that a high reflection coefficient is experienced at these frequencies. This corresponds well with the stop-band seen in the magnitude plot (Figure 3.9a).

A second identical FSS layer is added onto the spacer. This will introduce additional susceptance in terms of L and C to the entire equivalent transmission line. The higher frequencies have now clustered to the left-hand side of the admittance chart (Figure 3.10a). It can be observed that a larger portion (i.e. wider frequency range) of the admittance locus is now located near the brim of the chart (high reflection). This corresponds well with the magnitude plot shown in Figure 3.10b where a wider rejection bandwidth is observed. It can also be observed that marker #1 (1GHz) and marker #2

(2GHz) is now closer to the center of the chart. This implies a better matching (lower reflection coefficient) at the lower frequency. This again corresponds well with the magnitude plot where a much higher transmission coefficient (S_{21}) is observed at the pass-band.

Figure 3.8a. Admittance locus for one layer FSS.

Figure 3.8b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.9a. Admittance locus for one layer FSS and one spacer.

Figure 3.9b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.10a. Admittance locus for a two-layer FSS with one spacer foam.

Figure 3.10b. Magnitude of reflection (S11) and transmission coefficient (S21).

3.3.2 Experimental Results and Analysis

The measured and simulated transmission responses of the FSS at normal incidence are shown in Figure 3.11. Insertion loss averaging at about 0.5dB is observed over a bandwidth of 59.65% from 1GHz to 1.85GHz with the centre pass-band frequency at 1.425GHz. For out-of-band rejection, a measured insertion loss of 10dB is obtained between 4.4GHz to 5.05GHz and 6.02GHz to 12.8GHz, less a narrow transmission window between 9.4GHz to 9.7GHz. A 9.5dB roll-off from the initial -0.5dB is obtained over a frequency span of 2.55GHz (i.e. 1.85GHz to 4.4GHz).

Four transmission peaks are spotted with their mid-band at about 1.7GHz, 5.75GHz, 9.5GHz and 13.6GHz. The thickness of 25.4mm (for the spacer) will correspond to an electrical length of $0.144\lambda_{1.7\text{GHz}}$, $0.487\lambda_{5.75\text{GHz}}$, $0.8\lambda_{9.5\text{GHz}}$ and $1.151\lambda_{13.6\text{GHz}}$ at the respective transmission peak frequency. These peaks are the array interference peaks. When presented in terms of the reflection coefficient [55], they will appear as the array interference nulls. As have been explained earlier in terms of the transmission line analysis [55], the reflection nulls corresponds to the admittance passing through the centre of the smith chart thereby indicating a zero reflection coefficient. Zero reflection coefficient implies that minimal energy is reflected and they will appear as peaks shown in the measured transmission response (Figure 3.11).

Close correlation between measured and simulated results is observed only for frequencies up to 5GHz. A downward shift in the measured frequency response is observed at the frequency range 5GHz-6GHz and 11GHz-14GHz. The measured transmission peak at 5.6GHz occurs at a lower frequency compared to the simulated one at 6.4GHz. The difference between the measured and simulated results could be due to the following factors. Firstly, the actual material properties might have differed from those used in the simulation. At the higher frequency, the transmission peaks are sensitive to such small variations in the material properties [73]. Moreover, the permittivity of the actual material varies with frequency whereas in the simulation model, a constant value is used throughout the entire frequency band of simulation. Secondly, in the prototype

implementation, it is difficult to ensure a tight and uniform bonding of the FSS sheets to the foam material. Possibilities of the presence of air-gaps existed. This will affect the overall electrical length and the effective permittivity. High frequency transmission peaks are again sensitive to minor variations of these parameters [73].

Compared to the transmission peaks at the lower frequency end (i.e. 1GHz - 2GHz), the higher frequency peaks suffers more deviations when compared with the simulated result. This is also true for larger incidence angles, as depicted in Figure 3.12 for the TE case. It can be seen that while the transmission peaks below 2GHz remain relatively stable, the higher frequency peaks shifted upwards progressively at a decreasing magnitude as the scan angle increases (Figure 3.12).

Figure 3.13 compares the measured and simulated TE transmission responses at 45° incidence angle. A maximum insertion loss of 0.5dB is measured over a bandwidth of 32.33% (1.4GHz to 1.94GHz) with the centre frequency at 1.67GHz. For out-of-band rejection, continuous attenuation of more than -10dB is measured from 4.2GHz to 11.7GHz. A 9.5dB roll-off is measured over a frequency range of 2.26GHz (i.e. 1.94GHz to 4.2GHz).

The variation of the TM response measured progressively over different incidence angles is shown in Figure 3.14. The variation of the transmission response with angle of incidence is similar to that observed for the TE case.

Figure 3.15 compares the measured and simulated TM transmission responses at 45° incidence angle. A maximum insertion loss of 0.5dB is measured over a bandwidth of 25% from 1.54GHz to 1.98GHz with the centre frequency at 1.76GHz. For out-of-band rejection, continuous attenuation of more than -10dB is measured from 5.3GHz to 11.2GHz, except for the frequency range 6.5GHz to 7.4GHz where a slight increase in transmission is encountered. A 9.5dB roll-off is measured over a frequency span of 3.32GHz when the transmission power decreases from -0.5dB at 1.98GHz to -10dB at 5.3GHz. The difference between the simulated and the measured responses from 6GHz to

7.6GHz could due to the same imperfections cited for the normal incidence and TE case, except that the variation is more pronounced in this case.

The 0.5dB bandwidth reduces from 59.65% (normal incidence) to 32.33% and 25% for TE and TM respectively. A common 0.5dB bandwidth is still obtainable from 1.54GHz to 1.85GHz for the three extreme cases and from 1.54GHz to 1.94GHz for TE 45° and TM 45°. It can be seen from both TE and TM polarizations that as the angle of incidence increases, the transmission spike at around 6GHz reduces, hence increasing the –10dB stop-band bandwidth. Based on the ability to meet the 0.5dB bandwidth over the required frequency range and the wide rejection bandwidth, the FSS is observed to have a more desired performance at scanned incidence angle rather than at normal incidence. This coincides with most application where sloping of FSS is required [16].

Figure 3.11. Measured and simulated transmission responses for normal incidence.

Figure 3.12. Measured TE transmission response for different incidence angles.

Figure 3.13. Measured and simulated TE transmission responses for 45° incidence angles.

Figure 3.14. Measured TM transmission response for different incidence angles.

Figure 3.15. Measured and simulated TM transmission responses for 45° incidence angles.

3.4 Quad-Layer Bandstop Design with Ring and Circular Patch Element

This design is attempted with the aim of achieving a steeper roll-off and a higher attenuation over a moderate stop-band bandwidth. It is realised by using three foams to provide the separation between four layers of FSS substrates as shown in Figure 3.16. The core foam in the middle has a thickness of 25.4mm while the outer foams have thickness of 5mm each. All the foams have a similar dielectric constant of about 1.043 and a loss tangent of 0.002. The thickness of the core foam is approximately half-wavelength at 5.75GHz.

Figure 3.16. Quad layer FSS

The FSS element consists of a circular patch enclosed within a circular ring element (Figure 3.16). The ring has a mean diameter of 13.66mm and its circumference of 42.91mm is about a wavelength long at 6.99GHz. The ring has a trace width of 1.46mm with the inner circumference responsible for the upper frequency rejection and the outer circumference responsible for the lower frequency rejection. Overall, the ring provides

the necessary stop-band requirement while the patch is added to provide another stop-band resonance at a much higher frequency. The initial intention was to provide a more stabilise response at the higher frequency range (9GHz-11GHz) with respect to incidence angles, together with a steeper roll-off and a deeper attenuation null ($< -25\text{dB}$) over the frequency range of 6.5GHz to 9GHz for shielding effectiveness. This frequency range encompasses the 5th higher order harmonic of the GPS cum DCS 1800 antenna and hence the shielding could reduce interference from antennas operating at this band [74], [75].

The elements are arranged in a lattice with angle $\alpha = 40.89^\circ$ and a horizontal inter-element spacing of $D_x = 15.15\text{mm}$ ($D_x = 2 \times S_x$) as shown in Figure 3.16. Since this is neither a rectangular nor a triangular lattice, the lowest frequency for the onset of trapped grating lobe is estimated to be between 8.9GHz (based on rectangular lattice) and 10.3GHz (based on triangular lattice calculation) for an incidence wave at 45° and with effective dielectric loading of 2.27. The onset of free space grating lobe will occur between 11.6GHz and 13.3GHz for an incidence wave of 45° .

3.4.1 Design and Analysis with Smith Chart and Transmission Line Approach

The design and performance of the quad-layer FSS can be studied by visualising its admittance transformation on the admittance chart. Figure 3.17a depicts the simulated admittance locus for the entire frequency sweep of 1GHz to 10GHz. Five frequency points are marked for comparisons and they represent the pass-band at 1GHz (marker #1) and 2GHz (marker #2); and the rejection bands at 5GHz (marker #3), 7GHz (marker #4) and 9GHz (marker #5). It is observed that the admittance locus follows closely to the unity conductance circle (i.e. $G = 1$). It can be seen that marker #4 is close to the short circuit location and this corresponds to the transmission (S_{21}) null observed in the magnitude plot (Figure 3.17b). This is the stop-band resonance frequency and below this resonance, the FSS behaves like a capacitance sheet whereby most of its admittance locus is located on the bottom half of the admittance chart. Above the resonance frequency, the FSS sheet is inductive. Marker #1 and marker#2 are near to the center of the chart where

a lower reflection coefficient will be encountered. This corresponds to high transmission coefficient (S_{21}) at the lower pass-band frequency (Figure 3.17a).

With the addition of a thin layer of spacer (Figure 3.18a), the admittance for each frequency will rotate in a clockwise direction with the higher frequencies rotating over a larger angular span. The distance between the center of the chart to the location of these admittance points are approximately equivalent to the distance prior the rotation. This is because the addition of the low dielectric spacer is similar to movement along a lossless transmission line and the magnitude of the reflection coefficient along the line should be constant while the phase changes with the variation of the electrical length. The consistency in the reflection magnitude (S_{11}) can be observed by comparing the magnitude plot of Figure 3.17b and Figure 3.18b.

A second identical FSS layer is added onto the spacer. Marker #1 and marker #2 rotated clockwise away from the center of the chart (Figure 3.19a), corresponding to an increase in the reflection coefficient and a reduction of the pass-band transmission. A sharp change in the admittance trajectory is observed for marker #3 where this corresponds to a steep roll-off in the transmission coefficient (S_{21}) shown in the magnitude plot (Figure 3.19b). Comparing the S_{21} of Figure 3.18b and Figure 3.19b, a more gradual transition between the pass-band (<2GHz) and the stop-band is observed in Figure 3.18b (without the second FSS addition). A roll-off of -19.5 dB is observed over a frequency range of 5GHz when the S_{21} decreases gradually from -0.5dB at 1GHz to -20dB at 6GHz. In Figure 3.19b, the S_{21} decreases from about -1dB at 1GHz to -3.4dB at 3GHz and then increases slightly to -2.5dB at 4GHz before decreasing steeply to -30.5dB at 6GHz. This amount to a roll-off of 28dB over a short frequency spans of 2GHz. This is a unique feature where a wide stopband is achievable with two layers compared to other designs where more layers are required. The S_{21} of -10dB is observed from about 5GHz to 9.2GHz.

With the further addition of the thick spacer, the admittance for the different frequencies is again “uncoiled” in a clockwise direction (Figure 3.20a). As explained earlier, this is due to the fact that the admittance at the higher frequencies rotates over a larger angle

compared to those at the lower frequencies. Marker #3, marker #4 and marker #5 are at the brim of the chart and these are associated with high reflection coefficients. These correspond to the reflection (S_{11}) peaks observed at the stop-band between 5GHz and 9.2GHz. As expected, the difference observed between the magnitudes of the transmission and reflection coefficient with (Figure 3.20b) and without the addition of the spacer (Figure 3.19b) is insignificant.

The addition of the third FSS layer saw the “re-coiling” of the higher frequency admittance points to the left-hand side of the admittance chart (Figure 3.21a). As a larger portion of the admittance locus is now situated at the brim of the chart, a corresponding wider band-stop frequency is observed in the S_{11} and S_{21} magnitude plots (Figure 3.21b). The admittance locus between marker #1 and marker #2 are close to the center of the chart. This tally well with the higher transmission coefficient (S_{21}) observed at the pass-band (<2GHz). Although attenuation increases between 3GHz and 5GHz, the transmission null has actually been shifted to a higher frequency at around 7.5GHz. The null has a simulated attenuation of more than -35dB . The addition of the FSS to the present materials stack-up yields an increase in the stop-band bandwidth, a shifting of the attenuation to a higher frequency range and a deeper transmission null.

The addition of another thin foam to the stack-up rotates all the markers in a clockwise direction (Figure 3.22a). Comparing the magnitude plots with (Figure 3.22b) and without (Figure 3.21b) the addition of the present spacer, no significant difference is observed, except for the change in the gradient of the S_{21} plot between 5GHz and 6GHz.

The addition of the forth layer of FSS produces a considerably undesirable effect for the transmission at the lower frequency pass-band region. Marker #1 and Marker#2 are now further away from the center of the chart (Figure 3.23a) and this reduces the transmission coefficient observed at the pass-band of the S_{21} magnitude plot (Figure 3.23b). However, larger attenuation ($S_{21} < -25\text{dB}$) can now be achieved over a wider frequency range, starting from 5.1GHz to 9GHz. Although this design suffers a low transmission coefficient at its pass-band for the simulated case of a normal incidence wave, it will be

shown later that an improved pass-band transmission can be obtained while retaining its deep attenuation feature for scanned incidence conditions.

Figure 3.17a. Admittance locus for one layer FSS.

Figure 3.17b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.18a. Admittance locus for one-layer FSS and one spacer.

Figure 3.18b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.19a. Admittance locus for two-layer FSS with one spacer foam.

Figure 3.19b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.20a. Admittance locus for a two-layer FSS with two spacer.

Figure 3.20b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.21a. Admittance locus for a three-layer FSS with two spacers.

Figure 3.21b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.22a. Admittance locus for a three-layer FSS with three spacers.

Figure 3.22b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.23a. Admittance locus (passband) for a four-layer FSS with three foam spacers.

Figure 3.23b Magnitude of reflection (S11) and transmission coefficient (S21).

3.4.2 Experimental Results and Analysis

Figure 3.24 shows the measured and simulated transmission responses of the FSS at normal incidence. Three significant transmission peaks are measured at the frequency range of 1GHz to 2GHz, 5.5GHz to 6GHz and between 9GHz to 11GHz. These are the array interference peaks and they correspond to the lower, middle and higher frequency peaks respectively. The null depth between the lower and middle frequency peaks is much shallower when compared to the two-layer design. However, a larger attenuation is observed between the middle and upper frequency peaks. Average insertion loss of 0.5dB is measured over a bandwidth of 33.33% from 1.5GHz to 2.1GHz with the centre frequency of 1.8GHz. Although attenuation of more than half-power (i.e. 3dB) is obtainable between 3.8GHz and 4.5GHz, it should be emphasised that the FSS is designed primarily for a more critical damping between 6.5GHz and 9.5GHz. The attenuation bandwidth of -10dB and -25dB is measured from 6.2GHz to 9.08GHz and 6.5GHz to 9GHz respectively. A 9.5dB roll-off is obtained over a frequency range of 0.55GHz when the transmission coefficient drops from -0.5dB at 5.65GHz to -10dB at 6.2GHz.

Both simulated and measured results are in good agreement except for frequencies above 9GHz. This could either due to the difference in material properties between the actual and the simulated model or it could due to the finite FSS prototype. At this frequency, there are onsets of trapped grating lobes. The effect of these trapped grating lobes will not appear in the simulated response as an infinite model is used. On the contrary, a finite FSS is used in the actual implementation and these grating lobes will exit and scattered at the edges of the finite FSS.

The simulated and measured TE 45° transmission responses are shown in Figure 3.25. An average insertion loss of 0.3dB is measured from 1GHz to 2.6GHz. Thus, a 0.3dB bandwidth of 88.89% is obtained over the centre frequency of 1.8GHz. The attenuation bandwidth of -10dB and -25dB is measured from 6.3GHz to 9.2GHz and 6.5GHz to 9GHz respectively. A 9.5dB roll-off is obtained over a frequency range of 0.35GHz when the transmission coefficient drops from -1.5dB at 5.9GHz to -11dB at 6.25GHz.

Figure 3.26 shows the measured and simulated TM 45° transmission responses. Insertion loss of less than 0.4dB is observed from 1GHz to 3GHz. The attenuation bandwidth of -10dB and -25dB is measured from 6.3GHz to 8.8GHz and 6.5GHz to 8.05GHz respectively. As expected for a band-stop design (printed traces), the TM response yields a narrower bandwidth for the -25dB attenuation level when compared to the normal incidence and TE response. A 9.5dB roll-off is obtained over a frequency range of 0.32GHz when the transmission coefficient drops from -2.2dB at 5.98GHz to -11.7dB at 6.3GHz.

Compared to the two-layer design, the present configuration offers better passband insertion-loss performance, a steeper roll-off and larger attenuation at the stop-band. However, the bandwidth for the stop-band is narrower. Its performance is comparatively less sensitive to the variation in incidence angles, materials properties and bonding tolerances.

Figure 3.24. Measured and simulated transmission responses for normal incidence.

Figure 3.25. Transmission response (TE case) of FSS2 for 45° incidence angles.

Figure 3.26. Transmission response (TM case) of FSS2 for 45° incidence angles.

3.5 Quad-Layer Bandstop Design with Ring Element

This quad-layer design incorporates four FSS sheets with slotted-ring as its array elements. As shown in Figure 3.27, the materials that are used for the FSS are similar to those used in the previous quad-layered design. The ring has a trace width of 0.65mm and a mean diameter of 15.05mm. Its mean circumference of 47.28mm corresponds to one free space wavelength at 6.34GHz. Compared to the other two designs reported earlier, the larger ring dimension corresponds to a lower stopband resonance frequency. Overall, this helps to ensure that a -10dB attenuation is achievable at a lower frequency (i.e. 3.5GHz). The elements are arranged in a triangular lattice with the angle $\alpha = 60^\circ$ as shown in Figure 3.27. With an effective dielectric loading of 2.27 and a horizontal inter-element spacing of $D_x = 15.85\text{mm}$ ($D_x = 2 \times S_x$), the frequency for onset of trapped grating lobes occurs at about 9.83GHz. As frequency increases, surface wave will appear and this is followed by the onset of free space grating lobes at 12.75GHz.

Figure 3.27. Quad layer FSS.

3.5.1 Design and Analysis Using Smith Chart and Transmission Line Approach

Figure 3.28a depicts the simulated admittance locus for the entire frequency sweep of 1GHz to 10GHz. Five frequency points are marked for comparisons and they represent the passband at 1GHz (marker #1) and 2GHz (marker #2); and the rejection bands at 5GHz (marker #3), 7GHz (marker #4) and 9GHz (marker #5). It is observed that the admittance locus follows closely to the unity conductance circle. The admittance value at around 4GHz, which lies in between marker #2 and marker #3, is closest to the short circuit location (i.e. at left-hand side of the admittance chart with a marked angle of 180°). Compared to the other designs, the single sheet FSS is designed with a lower stop-band resonance frequency ($<5\text{GHz}$, marker #3). Below this resonance, the FSS behaves like a capacitance sheet whereby most of its admittance locus is located on the bottom half of the admittance chart. Above the resonance frequency, the FSS sheet is inductive. It can be seen that at frequencies away from the stop-band region, the FSS tends towards the center of the smith chart where it sees a normalized admittance of air. This implies that when the admittance gets closer to the center of the circle, the reflection coefficient reduces and more energy is allowed to pass through. This relatively transparent behaviour is evident at the lower pass-band frequency range (1GHz to 2GHz) of the corresponding magnitude plot as shown in Figure 3.28a.

With the addition of a thin layer of spacer (Figure 3.29a), the entire admittance locus is “uncoiled” and rotates in a clockwise direction. The angle of rotation for each frequency points on the admittance chart corresponds to the additional electrical length incurred by the spacer. The length of the spacer will be relatively longer compared to the wavelength at the higher frequencies and hence the angle of rotation for the admittance at the higher frequency will be larger compared to that at the lower frequencies. Comparing Figure 3.28a and Figure 3.29a, the admittance at 1GHz is rotated closed to a constant radius circle (marked in red). This implies that the magnitude of the reflection coefficient with and without the foam is similar. In terms of transmission line analogy, it implies that movement along the line will only change the angle of the reflection coefficient while the magnitude remains constant. This consistency is observed in the comparison of the magnitude plots in Figure 3.29b and Figure 3.28b. However, losses in the foam material will introduce additional resistive

values in the transmission line model and this affects the resistance (impedance analysis) or the conductance (admittance analysis) value.

A second identical FSS layer is added onto the spacer. This will introduce additional susceptance in terms of L and C to the entire equivalent transmission line. From Figure 3.30a, the admittance at the lower frequencies has moved further away from the center of the chart, indicating an increase in the reflection coefficient. This corresponds to the low transmission coefficient observed for the lower pass-band frequencies in Figure 3.30b. From Figure 3.30c, a larger admittance locus for the higher frequencies has now transformed closer to the brim of the admittance chart. This corresponds to the wider stop-band bandwidth (S_{21} of -10dB from 3.5GHz to 7.2GHz) that is observed in the magnitude plot of Figure 3.30b when compared to stop-band (3.4GHz to 5.6GHz) shown in Figure 3.29b.

With the further addition of the thick spacer, the admittance for the different frequencies is again “uncoiled” in a clockwise direction (Figure 3.31a and Figure 3.31b). As explained earlier, this is due to the fact that the admittance at the higher frequencies rotates over a larger angle compared to those at the lower frequencies. Marker #5 is observed to be near to the center of the chart. This implies a lower reflection coefficient and it corresponds to the transmission peak at 3GHz shown in the magnitude plot of Figure 3.31c. As expected, no significant difference is observed between Figure 3.31c (addition of spacer) and Figure 3.30b, except for the peak at 3GHz.

The addition of the third FSS layer saw the “re-coiling” of the higher frequency admittance points to the left-hand side of the admittance chart (Figure 3.32a). Marker #4 (7GHz) is now closer to the brim of the chart when compared to its location in Figure 3.31a. This implies that the higher frequency now experiences a higher reflection coefficient, indicating an upward expansion of the stop-band bandwidth. This phenomenon can also be observed by comparing the magnitude plots of Figure 3.31c and Figure 3.32b (with inclusion of the third FSS). Marker #3 is now transformed to the short circuit location of the admittance chart. The admittance locus for the lower frequencies is still considerably far away from the center of the smith chart (i.e. high reflection coefficient) and it can be seen from the magnitude plot

(Figure 3.32b) that the addition of the third layer FSS does not improve the pass-band transmission performance.

The addition of another thin foam to the stack-up rotates all the markers in a clockwise direction (Figure 3.33a). However, the distance from the center of the smith to each marker remains quite similar when compared to that in Figure 3.32a. In between marker #1 and marker #2, the admittance locus is seen to pass by close to the center of the smith chart. This portion will have a low reflection coefficient and it corresponds to the sharp transmission (S21) peak seen at 1.75GHz in the magnitude plot of Figure 3.33b. This spacer has the effect of extending the band-stop frequency to a lower frequency range when comparing Figure 3.33b with Figure 3.32b.

The addition of the forth layer of FSS saw the admittance at 1.3GHz crossing close to the center of the admittance chart. However, the admittance at 2GHz (marker #5) has moved further away from the center and has now incurred a higher reflection coefficient. The addition of this FSS layer has caused the pass-band to be shifted to a lower frequency region. From the simulated magnitude plot (Figure 3.34b), maximum transmission (S21) is now spotted for frequency less than 1.5GHz. For the stop-band (Figure 3.34c), a wider bandwidth is observed with a larger trace of the admittance locus residing near the brim of the admittance chart. A simulated S21 of -10dB is obtained continuously from 3.2GHz to 8.5GHz for the case of normal incidence (Figure 3.34c).

Figure 3.28a. Admittance locus for one layer FSS.

Figure 3.28b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.29a. Admittance locus for one-layer FSS and one spacer.

Figure 3.29b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.30a. Admittance locus (pass-band frequency) for two-layer FSS with one spacer foam.

Figure 3.30b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.30c. Admittance locus (band-stop frequency) for a two-layer FSS with one spacer.

Figure 3.31a. Admittance locus (pass-band frequency) for a two-layer FSS with two spacers.

Figure 3.31b. Admittance locus (band-stop frequency) for a two-layer FSS with two spacers.

Figure 3.31c. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.32a Admittance locus for a three-layer FSS with two spacers.

Figure 3.32b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.33a. Admittance locus for a three-layer FSS with three spacers.

Figure 3.33b Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.34a. Admittance locus (passband) for a four-layer FSS with three foam spacers.

Figure 3.34b Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.34c Admittance locus (band-stop frequency) for a four-layer FSS with three foam spacers.

3.5.2 Experimental Results and Analysis

Figure 3.35 shows the measured transmission response of the FSS at normal incidence. A maximum insertion loss of 0.5dB is measured over a bandwidth of 25.50% (1.3GHz to 1.68GHz) with the centre frequency of 14.9GHz. At 1.8GHz (i.e. DCS1800 band), the insertion loss deteriorates to about -2dB. The high insertion loss renders its inapplicability as a radome for the antenna's [24] DCS1800 operation when the antenna surface and the FSS are in parallel (i.e. the wave from antenna is at normal incidence to the FSS). For out-of-band rejection, the -10dB attenuation level is measured from 3.15GHz to 8.2GHz. Close correlation between the measured and simulated results is observed except for the increase in transmission coefficient measured at around 5.5GHz. The simulated transmission coefficient at that frequency is relatively low (< -20dB level) compared to the measured transmission coefficient of -12.5dB. This could be due to similar reasons highlighted for the double-layer design (refer to Section 3.3.2). Four transmission peaks are observed at the frequency range 1GHz-2GHz, 2.9GHz, 5.4GHz and above 9GHz. These are the array interference peaks. It can be seen that except for the lowest frequency peak, all the other higher frequency peaks have been shifted to a lower frequency range when compared to the previous designs.

Figure 3.36 shows the measured and simulated transmission responses for a TE polarised wave at incident angle of 45°. A pass-band bandwidth of 42.11% is measured with insertion loss averaging at 0.5dB from 1.2GHz to 1.84GHz (centre frequency at 1.52GHz). For out-of-band rejection, the -10dB attenuation is measured from 2.2GHz to 2.86GHz and 3.4GHz to 9.15GHz. Both simulated and measured results are in good agreement except for frequencies above 9.5GHz. At this frequency range, there is onset of trapped grating lobes within the materials layer. Due to the finiteness of the prototypes, these trapped grating lobes and surface waves could be launched into the free space at the edges of these finite samples compared to an infinite periodic surface in the simulation model [59]). In addition, the higher frequency transmission peaks are very sensitive to variation of material properties and fabrication tolerances. Hence, a more significant deviation between the simulated and measured results is observed at the higher frequency. A 9.5dB roll-off is measured over a frequency range of 0.36GHz when the S21 drops from -0.5dB at 1.84GHz to -

10dB at 2.2GHz.

Figure 3.37 shows the measured and simulated transmission responses for a TM polarised wave at incident angle of 45° . A pass-band bandwidth of 29.79% with an insertion loss of 0.5dB is measured from 1.4GHz to 1.89GHz with the centre frequency at 1.645GHz. For out-of-band rejection, the -10 dB attenuation is measured from 3.2GHz to 8.6GHz. Both simulated and measured results agree well except for frequencies above 9.5GHz, where they suffer the same problem as that encountered in the TE case. A 9.5dB roll-off is measured over a frequency range of 0.71GHz when the S_{21} drops from -0.5dB at 1.89GHz to -10 dB at 2.6GHz.

Overall, although the -10 dB attenuation can be obtained at a lower frequency range compared to previous designs, however the roll-off is not steep enough to ensure that the pass-band is not affected. With the attenuation extended to the lower frequency range, a higher insertion loss is incurred at the DCS1800 frequency band for the normal incidence case. As such, this design is only recommended for applications whereby the FSS radome is mounted at an angle with respect to the antenna surface.

Figure 3.35 Measured and simulated transmission responses for normal incidence.

Figure 3.36 Transmission response (TE) of FSS2 for 45° incidence angles.

Figure 3.37 Transmission response (TM case) of FSS2 for 45° incidence angles.

3.6 Quad-Layer Bandstop Design with Staggered Tuning

Instead of having four sheets of identical FSS elements, staggered tuning is employed for the design of this FSS whereby the two outer FSS layers are tuned to resonate at a different frequency compared to the two inner-most layers. Referring to Figure 3.38, the two inner sheets will have larger FSS ring elements compared to the two outer layers. The trace of the larger ring element is bounded by diameter d_1 and d_2 while d_3 and d_4 are the two diameters for the smaller ring element. With a mean diameter of 14.41mm and a copper trace width of 2.02 mm, the larger ring will have a mean circumference of 45.27mm corresponding to one free space wavelength at 6.63GHz. The smaller element has a mean diameter of 12.6 mm and a copper trace width of 2mm. Its mean circumference of 39.58mm corresponds to a free space wavelength at 7.58GHz. The core foam in the middle has a thickness of 25.4mm while the outer foams have thickness of 5mm each. All the foams have a similar dielectric constant of about 1.043 and a loss tangent of 0.002. The FSS patterns are etched on a 0.127mm thick layer of polyimide film with dielectric constant of 3.5 and loss tangent of 0.0026. The two inner FSS layers and the core foam will provide the basic band-stop attenuation bandwidth while the two outer FSS and foams are designed to suppress the leakage of energy appearing as transmission peaks over the stop-band.

With an effective dielectric loading of 2.27 and a horizontal inter-element spacing of $D_x = 15.904\text{mm}$ ($D_x = 2 \times S_x$), the lowest frequency for the onset of trapped grating lobe occurs at around 9.78GHz for an incidence wave at 45° . It should be noted that with a wider incidence angle ($> 45^\circ$), this frequency of onset of trapped grating lobe will be lower. As frequency increases, this grating lobe will propagate as free or trapped surface waves [72]. Following the formula given for a triangular lattice arrangement, there will then be an onset of free space grating lobe at around 12.7GHz for an incidence wave at 45° .

Figure 3.38. Quad layer FSS.

3.6.1 Design and Analysis using Smith Chart and Transmission Line Approach

Figure 3.39a depicts the simulated admittance locus for the entire frequency sweep of 1GHz to 10GHz. Five frequency points are marked for comparisons and they represent the pass-band at 1GHz (marker #1) and 2GHz (marker #2); and the rejection bands at 5GHz (marker #3), 7GHz (marker #4) and 9GHz (marker #5). It is observed that the majority part of the admittance locus lie on the bottom half of the admittance chart, where the imaginary admittance value is positive (i.e. $+jB$), indicating a capacitive characteristic. It is also observed that the admittance at the lower frequencies lie close to the center of the admittance chart, corresponding to minimal reflection coefficient. As frequency increases, the admittance moves towards the short circuit location (i.e. highly reflective) on the left-hand side of the admittance chart. This corresponds well with the magnitude plot shown in Figure 3.39b where minimal reflectivity is observed between at the pass-band between 1GHz to 2GHz and high reflectivity is observed at the rejection band of 6GHz to 9GHz.

With the addition of a thin layer of spacer (Figure 3.40a), minimal variations is observed for the admittance at 1GHz. However, for 2GHz, the admittance is rotated in a clockwise direction and is now spotted with an imaginary value (suceptance) of $0.5j$ and a real value (conductance) of about 1.4. The addition of the spacer is equivalent to a movement along the transmission line and this can ideally be represented by movement along the circumference of a constant radius circle (the red circle in the center of the chart in Figure 3.39a). This circle is usually known as the constant SWR (Standing Wave Ratio) circle. In order to track this transformation for the admittance at 2GHz (i.e. marker #2), the constant SWR circle in Figure 3.39a is copied to Figure 3.40a. The addition of the spacer should cause marker #2 in Figure 3.40a to move along the same constant SWR circle. But due to the fact that there is a small amount of losses in the foam, it can be seen that marker #2 does not move along the same circumference of the constant radius circle. It is also observed that the entire admittance locus is “uncoiled” and rotates in a clockwise direction. The angle of rotation for each frequency points on the admittance chart corresponds to the additional electrical length incurred by the spacer. The length of the spacer will be relatively longer when compared to the wavelength at the higher frequencies. Consequently, the angle of rotation for the admittance at the higher frequency will be larger compared to that at the lower frequencies. The admittance at 8GHz, which lies in between marker #4 and marker #5, is located on the purely reactance circle (i.e. at the brim of the admittance chart). Since reflection coefficient is the distance from the center of the chart to the admittance location, this would imply that a large reflection coefficient is experienced at 8GHz. This corresponds well with the magnitude plot shown in Figure 3.40a where maximal reflection is observed at 8GHz.

A second FSS layer (resonating at a lower frequency) is added onto the spacer. This will introduce additional suceptance in terms of L and C to the entire equivalent transmission line model. The higher frequencies now have clustered to the left-hand side of the admittance chart (Figure 3.41a). It can be observed that a larger portion (i.e. wider frequency range) of the admittance locus is now located near the brim of the chart (high reflection) with marker #4 (7GHz) at mid way. This corresponds well with the magnitude plot shown in Figure 3.41b where a wider rejection bandwidth is observed with the mid-frequency at 7GHz. It can also be observed that marker #1 (1GHz) and marker #2 (2GHz) is now further away from the center of the chart while

the other markers remain close to the brim. This implies that there is an increase in the reflection coefficient (i.e. higher mismatch) at the lower frequency. This again corresponds well with the magnitude plot where a much lower transmission coefficient (S_{21}) is observed at the pass-band.

With the further addition of the thick spacer ($\sim 0.487\lambda$ where λ is the free space at 5.75GHz), the admittance for the different frequencies is again “uncoiled” in a clockwise direction (refer Figure 3.42a). It can be seen that the admittance at 1GHz (marker #1) is rotated by 60° in an anti-clockwise direction from the initial -174° (in Figure 3.41a) to -114° as demonstrated in Figure 3.42a. Furthermore, the rotation follows the locus of the constant SWR circle, indicating that the loss of the spacer is negligible at 1GHz and its addition is equivalent to movement along a lossless transmission line. Similar transmission performance at 1GHz is observed in the magnitude plots shown in Figure 3.41b and Figure 3.42b (with the thick foam). In Figure 3.41a, the admittance at 5.75GHz will lie somewhere in between the angles marked 165° to 175° . With the addition of the thick foam, it can be estimated from Figure 3.42a that the admittance will again lie within this range of angle, with relatively minor deviation. Although finer frequency steps are required to re-calculate the exact admittance value, however the first-cut approximation was given based on the fact that the half-wavelength foam will transform the admittance by 360° back to the original location. This assumption will be valid only for a particular frequency and it can be seen that admittance at the higher frequencies will be rotated over a larger angle compared to those at the lower frequencies.

The addition of the third FSS layer saw the “re-coiling” of the admittance points to the left-hand side of the admittance chart (Figure 3.43a). An increased segment of the high frequency admittance is now lying on the brim of the chart where reflection coefficient is maximum. This implies an increase in the band-stop bandwidth as shown with a S_{21} of -10dB in the magnitude plot (Figure 3.43b). The admittance chart will only plot the result for frequencies up to 9GHz while other higher frequency results are avoided to prevent confusion. The admittance at the lower pass-band frequencies is near to the center of the chart. This relatively lower reflection coefficient performance is reflected in the magnitude plot shown in Figure 3.43b. A steeper roll-off and wider stop-band bandwidth is observed with the addition of this

FSS layer as illustrated by comparing Figure 3.43b and Figure 3.42b.

The addition of another thin foam to the stack-up is shown to rotate most of the admittance locus to the upper half of the admittance chart (Figure 3.44a). This effect is analogous to that shown in Figure 3.40a. Since the addition of the spacer is similar to merely moving along a transmission line, the distance between the center of the chart to the admittance at the various frequencies should have minimal variation. This consistency is reflected in the reflection (S_{11}) and transmission (S_{21}) responses depicted in Figure 3.44b and Figure 3.43b.

With the addition of the forth layer of FSS and a finer frequency resolution for the simulation at the pass-band (step of 0.25GHz from 1GHz to 2GHz), it can be observed that a majority portion of the pass-band admittance locus is residing near to the center of the admittance chart (Figure 3.45a). In particular, the admittance at ~ 1.7 GHz is passing through the center with zero reflection coefficient. For out-of-band performance (Figure 3.45b), a larger stretch of the higher frequency admittance locus is now transformed to the brim of the chart (i.e. large reflection coefficient), indicating a wider stop-band bandwidth. Good correlation is observed in the simulated magnitude plot (Figure 3.45c) where high transmission coefficient is obtained in the pass-band (1.5GHz to 2GHz) and attenuation of more than -10 dB is observed from 3GHz to 11GHz, except for a narrow spike appearing between 9.5GHz to 10GHz. A red arrow in the admittance chart of Figure 3.45a indicates this peak.

Figure 3.39a. Admittance locus for one layer FSS.

Figure 3.39b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.40a. Admittance locus for one layer FSS and one spacer.

Figure 3.40b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.41a. Admittance locus for two layers FSS and one spacer foam.

Figure 3.41b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.42a. Admittance locus for a two-layer FSS with two spacers.

Figure 3.42b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.43a. Admittance locus for a three-layer FSS with two spacers.

Figure 3.43b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.44a. Admittance locus for a three-layer FSS with three spacers.

Figure 3.44b. Magnitude of reflection (S11) and transmission coefficient (S21).

Figure 3.45a. Admittance locus (pass-band) for a four-layer FSS with three foam spacers.

Figure 3.45b. Admittance locus (band-stop frequency range) for a four-layer FSS with three foam spacers.

Figure 3.45c. Magnitude of reflection (S11) and transmission coefficient (S21).

3.6.2 Experimental Results and Analysis

The FSS sample is again measured using the conventional two-port free space method [64]. Figure 3.46 shows the measured transmission response for a normal incidence wave on the FSS. An average insertion loss of 0.6dB is measured over a bandwidth of 28.91% from 1.45GHz to 1.94GHz with a centre frequency at 1.695GHz. The 10dB attenuation is measured from 3GHz to 11GHz, except for a sharp transmission peak measured at around 8.9GHz. A 9.5dB roll-off is obtained over a frequency span of 1.11GHz when the transmission coefficient decreases from -0.6dB at 1.94GHz to -10.1dB at 3.05GHz. The simulated result reveals a sharp peak as well, but it occurs at a higher frequency. Again, the difference could be due to deviation in the material properties and the fabrication tolerances encountered in the actual implementation. Compared to the previous designs, there are only two significant array interference peaks at both lower and upper frequencies. The usual transmission peak at the mid-frequency (~6-7GHz) has been shifted to a higher frequency (8.9GHz) with a much reduced amplitude.

Figure 3.47 shows the measured and simulated TE transmission responses for a wave at incidence angle of 45°. A bandwidth of 62.07% with an average insertion loss of 0.6dB is measured from 1GHz to 1.9GHz with the centre frequency at 1.45GHz. For out-of-band rejection, attenuation of more than -10dB is measured from 3.05GHz to 10.3GHz, corresponding to a stop-band bandwidth of 108.61% with centre frequency at 7.25GHz. A 9.5dB roll-off is obtained over a frequency range of 1.2GHz when the transmission coefficient drops from 0.6dB at 1.9GHz to -10.1dB at 3.1GHz. Simulated and measured results correlate well except for frequencies higher than 10.5GHz. Again, these spurious resonance at the higher frequencies are very sensitive to small losses and angle of incidence [73] and hence slight variation between the simulated model and the prototype will result in a large deviation in the results at the high frequencies.

The TM transmission response at incidence angle of 45° is shown in Figure 3.48. A bandwidth of 42.42% with average insertion loss of 0.6dB is measured from 1.3GHz to 2GHz with centre frequency at 1.65GHz. For out-of-band rejection, a -10dB

attenuation is measured from 3.4GHz to 9.6GHz, corresponding to a bandwidth of 95.38% with the centre frequency at 6.5GHz. A 9.5dB roll-off is obtained over a frequency range of 1.45GHz. Deviation between simulated and measured results are observed only for frequencies higher than 9GHz.

Overall, this design presents a low insertion loss over the GPS and DCS1800 band. Average insertion loss of 0.6dB is obtained for a wide angle of incidence (0° to 45°) for both TE and TM polarizations. In this case, an increase in attenuation bandwidth is obtained without sacrificing on the pass-band insertion loss. Unlike previous design where transmission peaks appear at mid frequencies, continuous attenuation of at least 10dB is obtained across a wider stop-band bandwidth for TE and TM polarizations in the present design.

Figure 3.46. Measured and simulated transmission responses for normal incidence.

Figure 3.47. Measured and simulated TE transmission responses for 45° incidence angles.

Figure 3.48. Measured and simulated TM transmission responses for 45° incidence angles.

3.7 Conclusions

A few variants of low pass FSS have been presented in this chapter. They are designed to enable the low frequency communication signal (i.e. GPS and DCS 1800 frequency band) to pass through while reflecting higher frequency components. For the convoluted band-pass FSS, a pass-band bandwidth of 22.47% (1.54GHz to 1.93GHz) is obtained with only a single layer FSS. Selective signal reflection can be achieved by selecting the length of the convoluted segments to be approximately a quarter-wavelength at the frequency of rejection. However, the relatively longer element at

the higher frequencies will create transmission ‘windows’ and early onset of grating lobes. Hence this design will most probably be suitable for suppression of antenna’s higher order interference rather than for wideband RCS reduction. For this design, the transmission and reflection phenomena are analysed by studying the surface reflection phase, array transmission phase, their corresponding amplitude and the surface current.

To avoid the early occurrence of grating lobe and higher order resonance problem, high frequency band-stop FSS are implemented. The band-stop FSS are analysed by visualising the transformation of the admittance locus (on the Smith chart) upon subsequent additions of material layers to the multi-layer stack-up. Their performances are summarised in Table 3.1. It can be observed that the different design provides different degree of roll-off, band-pass and band-stop bandwidth, and level of attenuation. Several unique performance pertaining to some of the designs are observed. For the quad-layer design with ring and circular patch element, a steeper roll-off with a larger attenuation ($<-25\text{dB}$) is observed for all the extreme scanned conditions. As for the quad layer with ring element only, the pass-band for the normal incidence case is much lower than that required for GPS and DCS1800. It can only meet the pass-band requirement when the FSS is tilted. The quad-layer with staggered tuning has the best out-of-band performance whereby continuous attenuation of at least 10dB is obtained for almost all the scanned conditions, less a very minute spike of 0.2GHz for the normal incidence case.

Table 3.1 Summary of all the band-stop FSS performance.

	Double Layer	Quad Layer (ring & circular patch)	Quad Layer (ring)	Quad Layer (staggered tuning)
Normal Incidence				
Pass-band Bandwidth (with S21 ≥ -0.5dB)	59.65% (1GHz - 1.85GHz)	33.33% (1.5GHz - 2.1GHz)	25.50% (1.3GHz - 1.68GHz)	(S21 ≥ -0.6dB) ¹ 28.91% (1.45GHz - 1.94GHz)
Stop-band Frequency Range (with S21 < -10dB)	4.4GHz - 5.05GHz, 6.02GHz - 12.8GHz.	(-10dB) 6.2GHz - 9.08GHz (-25dB) 6.5GHz - 9GHz.	3.15GHz - 8.15GHz	3GHz - 8.8GHz, 9GHz - 11GHz
9.5dB Roll-off	2.55GHz	0.55GHz	N.A.	1.11GHz
TE 45°				
Pass-band Bandwidth (with S21 ≥ -0.5dB)	32.33% (1.4GHz - 1.94GHz)	88.89% (1GHz - 2.6GHz)	42.11% (1.2GHz - 1.84GHz)	(S21 ≥ -0.6dB) 62.07% (1GHz - 1.9GHz)
Stop-band Frequency Range (with S21 < -10dB)	4.2GHz - 11.7GHz	(-10dB) 6.3GHz - 9.2GHz (≤ -25dB) 6.5GHz - 9GHz	2.2GHz - 2.86GHz and 3.4GHz - 9.15GHz	3.05GHz - 10.3GHz,
9.5dB Roll-off	2.26GHz	0.35GHz	0.36GHz	1.2GHz
TM 45°				
Pass-band Bandwidth (with S21 ≥ -0.5dB)	25% (1.54GHz - 1.98GHz)	200% (1GHz - 3GHz)	29.79% (1.4GHz - 1.89GHz)	(S21 ≥ -0.6dB) 42.42% (1.3GHz - 2GHz)
Stop-band Frequency Range (with S21 < -10dB)	5.3GHz - 6.5GHz 7.4GHz - 11.2GHz	(-10dB) 6.3GHz - 8.8GHz, (≤ -25dB) 6.5GHz - 8.05GHz	3.2GHz - 8.6GHz	3.4GHz - 9.6GHz,
9.5dB Roll-off	3.32GHz	0.32GHz	0.71GHz	1.45GHz

¹ A larger portion of the S21 is averaging between -0.5dB and -0.6dB. Hence a S21 of ≥ -0.6dB is adopted.

Chapter 4

Design and Analysis of Balun-Fed Circularly Polarised Antenna

4.1 Introduction

This chapter describes the development of a coaxial balun-fed antenna that provides a circular polarization (CP) over a wide bandwidth compared to conventional CP microstrip antennas where certain tradeoffs are required between the impedance and CP axial ratio bandwidth [24], [11], [76]. This antenna, which consists of two orthogonal pairs of bow-tie antennas and with a shape resembling that of a Maltese Cross, will be useful for GPS, MSAT and DCS1800 mobile communication following an increasing demand to integrate mobile communications networks with tracking and location-determination technology [24]. While GPS requires CP operation, mobile communication (GSM) operates with linear polarization. A CP antenna can be used to receive a linearly polarized signal, with a maximum 3-dB loss for receiving a perfect horizontal or vertical polarized signal [9]. Hence, the antenna is designed for optimal GPS operation and then extended for GSM coverage rather than vice versa. Moreover, in most cases, a much more stringent voltage standing-wave ratio (VSWR) performance is required for the GPS operation, while for mobile communication, a VSWR of up to 3 (i.e. higher mismatch) or the equivalent of a return loss of 6 dB is tolerable [10].

4.2 Equivalent Circuit Parameters of a Balun [77]-[79]

The balun structure originally proposed by Roberts [77] is shown in Figure 4.1. The device has two lengths of coaxial transmission line, a and b . The center coaxial terminal $C1$ is the connection to the external RF source (or load), while the terminals $E1$ and $E2$ are connected to the balanced load (or antenna). Center conductors of lines a and b are connected at $C2$, while outer conductors of a and b are connected at $E3$ (short-circuit). The center conductor of line b ends at $C3$ (open-circuit).

The characteristic impedances of coaxial lines a and b are denoted as Z_a and Z_b respectively. Z_{ab} is the characteristic impedance of the balanced two-wire transmission lines comprising the outer conductors of line a and b . Z_S and Z_L are source and load impedances respectively, which may be connected to the balun. Z_M is the impedance looking into coaxial line b with an open circuit at $C3$ and Z_N is the impedance looking from $E1$ - $E2$ along the open transmission line ab with a short-circuit at $E3$.

Figure 4.1. Roberts balun structure and its equivalent circuit [78]-[79].

If the transmission line losses are neglected, the impedances Z_M and Z_N , which are open-circuited and short-circuited at $C3$ and $E3$ respectively, can be expressed as [78]:

$$Z_M = -jZ_b \cot \theta_b \quad (4.1)$$

$$Z_N = jZ_{ab} \tan \theta_{ab} \quad (4.2)$$

where θ_b and θ_{ab} are, respectively, the electrical lengths of transmission lines b and ab , taking into account their respective physical lengths and propagation constants. From the equivalent circuit, the impedance Z_{in} , of the balun structure can be expressed as [77], [78]:

$$\begin{aligned} Z_{in} &= \frac{Z_N Z_L}{Z_N + Z_L} + Z_M \\ &= \frac{Z_L Z_{ab}^2 \tan^2 \theta_{ab}}{Z_L^2 + Z_{ab}^2 \tan^2 \theta_{ab}} + j \frac{Z_L^2 Z_{ab} \tan \theta_{ab}}{Z_L^2 + Z_{ab}^2 \tan^2 \theta_{ab}} - j Z_b \cot \theta_b \end{aligned} \quad (4.3)$$

By letting $\theta_b = \theta_{ab} = \theta$ and rationalising, the following expression is obtained

$$Z_{in} = \frac{Z_L Z_{ab}^2 + j \cot \theta [Z_L^2 (Z_{ab} - Z_b \cot^2 \theta) - Z_b Z_{ab}^2]}{Z_{ab}^2 + Z_L^2 \cot^2 \theta} \quad (4.4)$$

It can be seen that the reactive component of the impedance Z_{in} is zero when $\cot \theta = 0$ (i.e. $\theta = \pi/2$), for which Z_{in} equals to Z_L . This occurred when the transmission lines are quarter wavelength at the center frequency (i.e. mid-band).

From Eqn. (4.4) with $\theta_{ab} = \theta_b = \theta$ and $\theta = 90^\circ$ at mid-band, a considerably wider bandwidth can be obtained by minimizing the imaginary term in Eqn. (4.4). For this purpose, Z_b is chosen according to:

$$Z_b = Z_L^2 / Z_{ab} \quad (4.5)$$

and Z_{ab} is made as large as possible [78].

4.3 Implementation of Coaxial Roberts Balun

Any twin wire transmission line can support both odd and even mode excitations. Figure 4.2 shows the E field of both modes in the cross-sectional view of the transmission line.

Figure 4.2. Even and odd modes.

In the case of a balanced bow-tie antenna, the arms (connecting to the antenna) are to be excited only by the odd mode. The odd mode excitation will ensure that the current entering the antenna are 180° out of phase. The relationship between the spacing D and characteristic impedance Z_{ab} between the two conductors is given as (referring to Figure 4.3a)[80]:

$$Z_{ab} = \frac{120}{\sqrt{\epsilon}} \cosh^{-1} \frac{D}{d} \quad (4.6)$$

where d is the diameter of the outer conductor and ϵ is the permittivity of the medium between the two conductors. The characteristic impedance for the normal coaxial transmission line is given as (referring to Fig.4.3b):

$$Z_0 = \frac{138}{\sqrt{\epsilon}} \log \frac{b}{a} \quad (4.7)$$

where b and a are the radius for the outer conductor and inner probe respectively.

Figure 4.3(a). Twin-wire transmission line; (b). Coaxial transmission line.

4.4 Design and Development [24]

The antenna consists of four truncated triangular copper plates of 1.95mm thickness is shown in Figure 4.4. Each truncated triangular element has a flare angle of 63° and a length of 22.4mm for each of the arms (refer Figure 4.5). The radiating elements are connected to four cylindrical conductors. These conductors form two orthogonal pair of baluns to feed the respective bow-tie in the two axes. Each pair of balun has been designed following the odd and even mode analysis. In our case of a balanced bow-tie pair, the coaxial balun are excited only by the odd mode. The odd mode excitation will ensure that the current entering the bow-tie pair are 180° out of phase. Critical dimensions for the balun in Figure 4.5 are presented in Table 4.1. Both orthogonal pairs of balun have the same dimension except for the inner conductor residing inside the cylinders. To prevent the inner conductors of the orthogonal baluns from touching each other, the length for one of them is increased to crossover at a higher height. This will cause some differences in the S11 phase as seen from the input port of the two baluns.

4.5 Measurements and Results

The simulated impedance looking across the gap of either pair of bow-tie is $49.52 + j5.5$ ohms at the design center frequency of 1.575GHz. Upon integrating with the balun, the simulated impedance (using Ansoft HFSS) of the entire antenna structure is $50.8 + j3.5$ ohms. The measured and simulated VSWR at the input port of the high and the low balun are shown in Figure 4.6a. This antenna has a measured bandwidth of 12.06% (1.48GHz

to 1.67GHz) with VSWR less than 2. For receive mode operation, a VSWR of less than 3 is obtained from 1.445GHz to 1.87GHz, for a 26.98% bandwidth. A small variation in the VSWR is observed between the two input ports, resulting in the narrower common bandwidth. The non-symmetrical performance is due to different fabrication and tolerance at the joint of the balun to the bow-tie elements.

It should be highlighted that for the present design, tradeoff in the VSWR performance is adopted for better matching of the phase (minimal phase difference) between the two input ports. The small phase difference will help to ensure a good CP axial ratio over the desired band. The consequence for this is that a good VSWR can only be obtained over a small optimized bandwidth as can be seen from the expanded VSWR plot (Figure 4.6b). Without the CP requirement, the bandwidth (VSWR<2) of this balun-fed antenna is expected to cover up to 80% of the frequency range presented in Figure 4.6b. The limiting factor for the VSWR in this case is not the balun but the radiating element. This is because the design of the radiating element can actually be treated as that of a half-wave, single band dipole design.

To obtain circular polarization, a 1-4 GHz Anaren10025-3 90° hybrid is used to provide the $\pi/2$ phase difference between the two input ports. This will eventually produce a 90° sequential phase difference between the four triangular radiating elements attached to the baluns. But, prior to attaching the hybrid, the difference in S_{11} phase measured between the two input ports should be minimal. Figure 4.7 shows the measured S_{11} phase performance at the two input ports. A maximum measured phase difference of 5° is obtained from 1.4GHz to 1.7GHz and a maximum of 10° is measured from 1.7GHz to 2GHz. This small phase difference is achieved by using a sliding metal clamped to the exterior of the balun.

The transmission coefficient S_{12} and S_{21} is lower than -22 dB across the desired frequency band, indicating good isolation between the two pair of bow-ties fed by the dual-axis baluns. Mutual coupling is expected to be minimal, as the E fields for two pairs are orthogonal to each other.

With the attachment of the 90° hybrid, the measured axial ratio (AR) at the boresight is shown in Figure 4.8. In comparison with the VSWR bandwidth, it shows a relatively wideband axial ratio performance with $AR < 2\text{dB}$ over the frequency band 1.4GHz to 2GHz. Figure 4.9 show the measured axial ratio performance versus the angle of antenna rotation (angle ϕ in Figure 4.5). As expected, the axial ratio deteriorates as the antenna rotates away from the boresight. Figure 4.10 shows the measured circular polarization gain pattern at different frequencies. A maximum gain of 4dBi at 6° off the boresight and a 3dB beamwidth of at least 58° are obtained at 1.575GHz. At 1.8GHz, a maximum gain of 3.58dBi is obtained at -2.1° off the boresight.

4.6 Conclusions

A novel antenna in the shape of a Maltese Cross has been proposed for GPS and GSM/DCS1800 operation. Two pairs of coaxial balun, with one slightly taller than the other, are designed to feed the bow-tie elements in an orthogonal manner. The isolation between the two bow-tie antennas is found to be minimal in such an arrangement. The phase difference between the two input ports of the two baluns is minimized with a modified balun and a sliding metal clamp on the exterior of the balun. Axial ratio performance ($AR < 2\text{dB}$) is obtained over the frequency band of 1.4–2 GHz. This antenna exhibits a relatively better axial ratio performance that is easily obtained as compared to microstrip antennas, where higher tolerance is required in order to excite the appropriate modes for circular polarization. The antenna will be integrated with the quad-layer FSS reported in Chapter 3 and its performance in both the radiation and scattering mode would be evaluated and reported in the next chapter.

Figure 4.4. Top view of the four truncated triangular radiating elements attached to cylindrical baluns.

Figure 4.5. Structure of a coaxial balun

Table 4.1 Dimensions for a pair of crossed coaxial balun.

Parameters	Balun No. 1	Balun No. 2
D_a	1.95mm	1.95mm
D_{a2}	2.4mm	2.4mm
D_b	2.56mm	2.56mm
D_x	1mm	1mm
D_y	4.5mm	4.5mm
L_a	57.65mm	60.65mm
L_b	44.91mm	47.91mm
H_a	2mm	2mm
H_b	46.4mm	46.4mm
H_c	2mm	2mm
S_1	7mm	7mm

Figure 4.6a. VSWR at the input of the high balun and the low balun.

Figure 4.6b. VSWR (over an extended frequency range) at the input of the high balun and the low balun.

Figure 4.7. Measured S_{11} phase at the input of the low and the high balun.

Figure 4.8. Measured axial ratio at the boresight (antenna rotating in a fixed axis).

Figure 4.9. Measured axial ratio with the antenna rotating on the azimuth.

Figure 4.10. Measured circular polarisation gain pattern.

Chapter 5

Integration of Periodic Array with Antenna

5.1 Introduction

It is well known that with proper design of the antenna and its radome [16], [81]-[83], it is possible to reduce the antenna scattering so that it becomes less visible to an incident electromagnetic wave [13]-[17]. Frequency selective surfaces (FSS) have been shown to reduce the out-of-band scattering of antenna over a wide frequency band for specified directions via proper shaping [20]. Examples include the hybrid radome [21], conical [22] and tilted flat panel [23] FSS geometry. Besides application in RCS (radar cross section) reduction, FSS can be used to reject unwanted interference from the external sources that is affecting the antenna at its higher harmonics or to reduce coupling and unwanted scattering to the surrounding electronics systems [82], [83].

In this chapter, the effect of the FSS on the gain and radiation pattern of the GPS-cum-DCS1800 Maltese cross-shaped antenna [24] will be evaluated for the in-band operating frequency range and its effect on the antenna backscattering reduction will be evaluated at the antenna's out-of-band frequency range. The quad layer FSS (with staggered tuning) presented in the previous chapter will be used for the integration. It should be noted that the Maltese cross look-alike antenna is designed with a finite ground plane. The ground plane is relatively larger compared to the radiating element and will be expected to scatter most of the incident energy. A comparison of the total scattering of the antenna with that calculated for the finite ground plate (using the formula for square plate RCS) will also be presented.

5.2 Experimental Setup and Measurement

For the present application, the FSS is inclined at an angle ϕ of 23.5° from the vertical axis such that the main specular return will be reflected towards other direction as shown in Figure 5.1. The tilting also helps to ensure a clearance of 5cm between the FSS and the

center of the radiating element. The 5cm clearance is about $\lambda/4$ away from the radiating element. The evanescent waves will have practically diminished and the effect of the FSS on the radiating element would be minimized [84]. The effect of the FSS on the antenna gain and radiation pattern performance will be evaluated for the in-band antenna operating frequency range from 1.5GHz to 1.9GHz. At out-of-band frequency, for example at 8.5GHz and 9.7GHz, the effect of the FSS on the antenna monostatic RCS, i.e. backscattering, will also be evaluated. The two measurement setups will be slightly different. For the RCS measurement, an additional antenna (next to the transmit antenna) is required to receive the power scattered back from the Maltese cross-shaped antenna.

Figure 5.1. (a) Schematic diagram of antenna and FSS set-up; (b) actual prototype setup for the measurement of the monostatic back scattering.

5.3 Antenna Out-of-Band Scattering

The measured $RCS^\dagger$ of the antenna with and without a termination is compared as shown in Figure 5.2 for an azimuth sweep of $\pm 90^\circ$ at 8.3GHz. A 50 ohms load is used for the termination as this is the impedance value of the feed network that is connected to the antenna for its in-band operation. It should be noted that beyond its in-band operating frequency range, for example at 8.3GHz, the 50ohms load might not be the perfect

termination. Nevertheless, it can be seen from Figure 5.2 that with a 50 ohms termination, the peak of the RCS is reduced by about 0.9dB for both TE and TM cases. Due to preceding reasons and the desire to achieve a closer resemblance to the actual implementation whereby the antenna is connected to a feed network, the terminated antenna will be used for all following investigations. This is also analogous to the rationale developed for the determination of antenna scattering [85]. For angles away from the broadside, the reduction in the backscatter is more significant with the open-circuited antenna. Similar characteristics can be observed for the measured RCS as plotted in Figure 5.3 for 9.7GHz. Although there is a less significant reduction in the main peak of the RCS. It can be observed that in each of the frequency and polarization plots (Figure 5.2 – Figure 5.3), the nulls and peaks of the RCS pattern for the terminated and unterminated cases follow relatively closely to each other up to an angular span of $\pm 26^\circ$. At the broadside, a peak RCS of about 12dBm^2 and 14dBm^2 is obtained for both polarizations at 8.3GHz and 9.7GHz respectively. The measured RCS is contributed by both the radiating element and the ground plane. The RCS of the standalone finite ground plane (23cm by 23cm) is estimated to be 14.3dBm^2 and 15.65dBm^2 at 8.3GHz and 9.7GHz respectively. The estimation uses the formula for the RCS of a square plate (i.e. $\sigma = 4\pi\text{Area}^2/\lambda^2$). It can be deduced that the RCS of the entire antenna, including the radiating element and the ground plane, is about 2.3dB and 1.65dB lower than that of the standalone square plate respectively. This could perhaps due to the cancellation effect of the scattered field between the antenna element and the ground plane or the positioning error incurred while trying to maintain the antenna in an upright position. It can also be due to the fact that the antenna has a smaller effective aperture compared to the square ground plane.

5.4 Effect of Periodic Array on Antenna Out-of-Band Scattering

With the addition of the FSS, the peak of the RCS is reduced by about 18dB (12dBm^2 to -6dBm^2) and 23dB (12dBm^2 to -11dBm^2) for TE (Figure 5.4a) and TM (Figure 5.4b) cases at 8.3GHz respectively. Within the azimuth angle of $\pm 30^\circ$, the RCS of the antenna

[†] The unit for RCS is dBm^2 . A similar notation, dBsm , is used in all the graphical plots.

with FSS inclusion is lower than that of the antenna without the FSS shield. However, with careful observation, the RCS of the antenna with the FSS inclusion begins to increase at an azimuth angle of about $\pm 26^\circ$. It eventually reaches to an almost equivalent level of that without the FSS inclusion at the azimuth angle of $\pm 40^\circ$ for the TE case (Figure 5.4a). This can be explained by the following. At broadside (i.e. azimuth angle of 0°), the antenna and ground plane is completely shielded behind the tilted FSS and hence the RCS reduction is most effective. As the antenna and FSS rotates away from the broadside, the antenna's cylindrical balun and the ground plane behind the FSS will be increasingly exposed to the normal incident wave. This is due mainly to the finite size FSS which when placed at a distance away from the ground plane, could not provide a complete shielding for the ground plane and antenna behind it. In the extreme case (i.e. at azimuth angle of $\pm 90^\circ$), the antenna's balun will be completely exposed to the transmitting antenna (as appeared in Figure 5.1). Hence the RCS value should be equivalent to that without a FSS shield.

The above explains for the phenomena whereby the RCS of the antenna (after installing the FSS), reaches to a comparable level as the original antenna without the FSS. However, it should be highlighted that at the extreme azimuth angle ($\pm 90^\circ$), the RCS of the antenna (with FSS) increases to a level beyond the RCS of the original antenna (refer to Figure 5.4a). This is possibly due to the strong scattering at the edges of the finite FSS with the scattering that is due to the edges of a finite array [86]. This phenomenon is more significant for the TE polarization rather than the TM polarization. This is due to the fact that the FSS, when seen from the side at azimuth angles of $\pm 90^\circ$ (Figure 5.1), is tilted at 23.5° away from the vertical axis (as compared to 66.5° from the horizontal axis) and hence the FSS will scatter more along the direction of the vertical axis (i.e. direction of the TE polarization). For the TM case (Figure 5.4b), at angles beyond $\pm 68^\circ$, the RCS of the antenna with the FSS shield is observed to be higher than that of the antenna solely.

At 9.7GHz, the peak of the RCS at broadside is reduced by 16.5dB (13.5dBm^2 to -3dBm^2) and 14.5dB (13.5dBm^2 to -1dBm^2) for TE (Figure 5.5a) and TM (Figure 5.5b)

polarizations respectively. Within an azimuth angle of $\pm 34^\circ$, a lower RCS is observed with the FSS inclusion for both TE and TM cases. Again, beyond this angle, there are certain RCS peaks that are comparable to the RCS of the antenna solely. At around $\pm 90^\circ$, the RCS of the antenna with FSS shield increases to a level much higher than that at 8.3GHz. This is due to the fact that with the increase in frequency and wide incidence angle, there are onsets on trapped surface waves associated with the FSS inter-element spacing and its dielectric loading [86], [87]. These surface waves will exit and scattered from the finite edges of the FSS in the end-fire direction [87]. The combination of this together with the scattering that existed at the edges of a finite array (as had been observed at 8.3GHz) resulted in a higher RCS at 9.7GHz. Again, this phenomenon is more significant for the TE polarization compared with the TM case (Figure 5.5b).

Despite the slight increment of the RCS at angles far away from the broadside, the FSS is effective in reducing the RCS of the antenna for angles within $\pm 30^\circ$, which could be sufficient for most applications [20]. Due to the size and symmetry limitation imposed by the tilted FSS plate, the RCS reduction is for the azimuth axis only rather than in both the azimuth and elevation axis. To extend the RCS reduction for a different axis, a more symmetrical FSS geometry such as those of a cone are preferred [16]. In addition, the antenna can be placed in an inset cavity and shaped FSS radomes can be used to cover the opening of the inset cavity [88].

5.5 Scattering of Antenna with Tilted Periodic Array versus Scattering of Tilted Metallic Plate

At out-of-band frequency, the tilted FSS behaves like a metallic plate and hence it is worthwhile to compare its performance with that of a titled metallic plate. For TE polarization (Figure 5.6a), closer correlation between the measured RCS of the antenna with FSS and the simulated RCS of a tilted plate is observed within an azimuth angle of $\pm 10^\circ$ from the broadside. The measured and simulated results show a RCS value of about -5.8 dBm^2 and -6.25 dBm^2 at the broadside respectively. This deviation could due to the error in ensuring the precise titling angle of the FSS. At angles away from the broadside,

the measured RCS of the titled FSS is higher than that of a simulated titled metallic plate. Although the size of both the tilted FSS and metallic plate is similar, there is a vertical ground plane behind the finite FSS. As mentioned earlier, at certain angles, the FSS will not be able to provide a complete shielding to the ground plane behind it. Hence, additional scattering that is due to the ground plane will be captured in the measured result as compared to the simulated result of a standalone tilted metallic plate. Figure 5.6b compares the measured and simulated results for the TM polarization. The measured RCS broadside peak is about 4dB lower than the simulated one. At 9.7GHz, both measured TE (Figure 5.7a) and TM (Figure 5.7b) broadside peak is higher than the simulated one. Furthermore, within the azimuth angle of $\pm 10^\circ$ from the broadside, the measured RCS is also much higher than the simulated metallic plate. At higher frequency, due to the shorter wavelength, slight variation in the implementation could lead to a large deviation in the measured response when compared with the simulated one. Overall, a more stringent control of the tilt angle as well as a larger FSS shielding is required for a more accurate comparison with the simulated result of a titled metallic plate. The implementation requires a more precise jig fixture coupled with larger FSS substrate materials. Unfortunately, due to budget constraints in the project, this is not implemented for the current proof-of-concept study.

Figure 5.2a. Measured monostatic RCS (TE) of antenna at 8.3GHz with and without a 50ohms termination.

Figure 5.2b. Measured monostatic RCS (TM) of antenna at 8.3GHz with and without a 50ohms termination.

Figure 5.3a. Measured monostatic RCS (TE) of antenna at 9.7GHz with and without a 50ohms termination.

Figure 5.3b. Measured monostatic RCS (TM) of antenna at 9.7GHz with and without a 50ohms termination.

Figure 5.4a. Measured monostatic RCS (TE) of antenna at 8.3GHz with and without FSS.

Figure 5.4b. Measured monostatic RCS (TM) of antenna at 8.3GHz with and without FSS.

Figure 5.5a. Measured monostatic RCS (TE) of antenna at 9.7GHz with and without FSS.

Figure 5.5b. Measured monostatic RCS (TM) of antenna at 9.7GHz with and without FSS.

Figure 5.6a. Comparison between measured RCS (TE) of antenna with FSS and simulated RCS of titled metal plate at 8.3GHz.

Figure 5.6b. Comparison between measured RCS (TM) of antenna with FSS and simulated RCS of titled metal plate at 8.3GHz.

Figure 5.7a. Comparison between measured RCS (TE) of antenna with FSS and simulated RCS of titled metal plate at 9.7GHz.

Figure 5.7b. Comparison between measured RCS (TM) of antenna with FSS and simulated RCS of titled metal plate at 9.7GHz.

5.6 Effect of Periodic Array on Antenna In-Band Performance

It should be noted that the antenna prototype and the measurement set-up for the present study (with FSS integration) differs from that reported in Chapter 4 as well as in our earlier publication [24]. Consequently, slight deviations in the result are inevitable. The radiation performance of the antenna is measured for the azimuth axis only. The pattern measured has a slight variation compared to that reported earlier [24]. This is due to the different setup implemented for the two different measurement scenario. With the attachment of the 90° hybrid, the measured axial ratio (AR) performance versus the angle (azimuth) of antenna rotation is shown in Figure 5.8. There is a degradation of the axial ratio performance with the addition of the FSS shield. However, a 3-dB bandwidth is still obtainable for an angle of 110°. For actual circular polarization (CP) operation, the axial ratio performance in both elevation and azimuth axis should be evaluated with a symmetrical FSS shielding. The measured normalized radiation pattern is shown in Figure 5.9. With the inclusion of the FSS, a lower gain is observed across the azimuth sweep. The antenna gain at broadside is about 0.5dB lower than the gain of the antenna without the FSS. A 3dB-beamwidth of 75° (-30° to 45°) is obtainable from the antenna solely as compared to a beamwidth of 73° (-28° to 45°) with the FSS inclusion. A boresight shift of about 0.9° is observed. The VSWR of the antenna measured at the input port of the two baluns (one crossing at a higher location compared to the other [24]), is compared for the case with and without FSS inclusion. Figure 5.10 shows the VSWR over the frequency range where the antenna is optimized for circular polarization while Figure 5.11 shows the VSWR at frequencies beyond the intended CP operation. From Figure 5.10, the VSWR is generally higher with the FSS inclusion, except for frequencies below 1.55GHz. For frequencies higher than its intended usage (Figure 5.11), a mixture of minor improvement and degradation of VSWR is observed with the FSS inclusion. The finite FSS has a relatively minor impact on the antenna's in-band performance compared with its out-of-band RCS performance. This is due to the fact that the band-pass FSS is supposedly transparent (with low insertion loss) at the antenna's in-band frequency range.

Figure 5.8. Measured axial ratio over the azimuth axis at 1.7GHz.

Figure 5.9. Measured antenna radiation pattern at 1.7GHz for an azimuth sweep.

Figure 5.10. Measured antenna VSWR over the frequency band intended for CP operation.

Figure 5.11. Measured antenna VSWR over an extended frequency range (beyond the CP operation).

5.7 Relationship between Antenna Range Equation and Periodic Array

The performance of the FSS is closely associated to the various parameters of the radar range equation. With a transmit and receive antenna gain of G_t and G_r respectively, the detection range r is given as [89]:

$$r = \left[\frac{P_t G_t G_r \lambda^2 \sigma}{P_r (4\pi)^3} \right]^{\frac{1}{4}} \quad (5.1)$$

where P_t and P_r is the transmit and receive power respectively and σ is the RCS of the antenna. With all the parameters of Eqn. (5.1) remaining unchanged, a reduction of the antenna gain (G_t and G_r) as a result of the loss incurred in the FSS radome will reduce the antenna's detection range r . This is often undesirable, as it implies a reduction of the antenna's efficiency. Assuming that the gain of the transmit and receive antenna (with the FSS inclusion) is given as $G_{t_{fss}}$ and $G_{r_{fss}}$ respectively, then the ratio between the maximum detectable range r and the new detection range r_{fss} can be written as:

$$\frac{r_{fss}}{r} = \left[\frac{G_{tfss} G_{rfss}}{G_t G_r} \right]^{\frac{1}{4}} \quad (5.2)$$

But $G_{tfss} = (L_{fss})(G_t)$ and $G_{rfss} = (L_{fss})(G_r)$, where L_{fss} is the insertion loss of the FSS. Hence Eqn. (5.2) can be written as:

$$\frac{r_{fss}}{r} = \left[L_{fss}^2 \right]^{\frac{1}{4}} \quad (5.3)$$

The range reduction will be given by:

$$\text{range_reduction}(\%) = \left(1 - \frac{r_{fss}}{r} \right) \times 100\% \quad (5.4)$$

When expressed in decibels, the ratio between the maximum detectable range r and the new detection range r_{fss} can be written as:

$$\frac{r_{fss}}{r} = 10^{\frac{1}{40} \{ [G_{tfss}(\text{dB}) + G_{rfss}(\text{dB})] - [G_t(\text{dB}) + G_r(\text{dB})] \}} \quad (5.5)$$

For a loss of 0.5dB (i.e. linear value of 0.891) incurred by the FSS, the range would be reduced by 0.56% with the inclusion of the FSS shield. It should be noted that further losses due to beamwidth and polarization deviation caused by the FSS will further reduce the detection range.

The above considers the antenna operating in the in-band frequency range. At out-of-band, the antenna is a potential scatterer. Referring to Eqn. (5.1) again, with all parameters remained unchanged, a reduction of the RCS σ would reduce the maximum detectable range of the antenna. This is desirable from the perspective of the antenna (and hence the target) detectability and survivability. The tilted FSS will provide the shaping required to reduce the monostatic RCS peak, as depicted in Figure 5.4 and Figure 5.5. Similarly, the ratio between the maximum detectable range r and the new detection range r_{fss} in terms of RCS can be expressed as:

$$\frac{r_{fss}}{r} = [\Delta\sigma]^{\frac{1}{4}} \quad (5.6)$$

where $\Delta\sigma$ is a ratio of the linear value of the antenna RCS with (σ_{fss}) and without (σ) the FSS inclusion. When expressed in decibels, the ratio for the range can be expressed as:

$$\frac{r_{fss}}{r} = 10^{\frac{1}{40}[\sigma_{fss}(\text{dB}) - \sigma(\text{dB})]} \quad (5.7)$$

Referring to Figure 5.4 and Figure 5.5, reduction of 18dB, 23dB 16.5dB and 14.5dB are observed for the TE and TM polarizations at 8.3GHz and 9.7GHz respectively. This will translate to a reduction of the detection range by a factor of 64.52%, 73.39%, 61.32%, and 56.60% respectively. It should be highlighted that this range reduction is borne by the antenna of the external parties.

5.8 Conclusions

This is an initial attempt to investigate the effect of a tilted FSS on the monostatic RCS reduction of an antenna mounted with a finite ground plane. This planar titled FSS have proven its effectiveness in reducing the RCS over the broadside specular region. However, it has been discovered that at angles away form the broadside, the RCS reduction degrades as a result of the insufficient shielding provided by the finite FSS. At the extreme azimuth angles, the RCS for the antenna with FSS inclusion is in fact higher than the antenna itself (i.e. without FSS). This is deemed to be associated to the strong scattering at the edges of a finite array. It has also been shown that this increase in RCS is more prominent for the TE polarization rather than the TM case. The presence of trapped grating lobes (in the FSS material) at the higher frequency end is also reflected in the measured RCS result. These trapped grating lobes will exit and scatter at the finite edges of the FSS, giving rise to the high RCS observed at the extreme azimuth angles. These are some of the unique phenomena observed and reported for this relatively novel attempt to integrate a tilted FSS with the circularly polarized Maltese-cross antenna. For effective RCS reduction in both elevation and azimuth axis, a more symmetrical FSS shielding, such as those of a cone or pyramid are suggested. The impact of the FSS on the antenna's

in-band radiation performance have shown to be minimal and this is as expected since the FSS is supposedly transparent at the in-band frequency. The seemingly contradictory impact of the FSS performance on the radar range equation has been revealed. For in-band operations, the reduction of the detection range as a result of the loss in the FSS is undesirable. However, for out-of-band operations, the reduction in antenna scattering as a result of the FSS inclusion is desirable. Simplified expressions relating the detection range with 1) the loss of the FSS and 2) the RCS reduction as a result of the inclusion of FSS have been derived. The detection range will only be reduced by 0.56% as a result of an insertion loss of 0.5dB incurred by the FSS. On the contrary, greater benefits of the FSS inclusion is observed when the detection range is reduced by at least 56.60% as a result of a 14.5dB reduction in the monostatic RCS.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

Several analytical and design techniques for FSS had been reviewed. The accuracy of these methods has been verified with several test cases. The techniques include the transmission line cum equivalent circuit method and the FPI analogy. For the first method, a separate set of equations had been presented for the band-stop and band-pass case of which the expression for opaqueness and transparency were obtained. The equations have been presented in a systematic manner such that it depicts clearly the transformation of the admittance and the phase of the reflection coefficient with respect to the sequential addition of materials. Besides, approximation for the bandwidth and resonant frequency have been derived for the bandstop FSS based on this formulation. For the test case, the phase of the reflection coefficient of a FSS resonating at 590GHz with $Q=5.55263$ agrees well with published results. For the FPI analysis, a thorough understanding of the formulation allows us to extend the prediction to the design a tri-pole FSS where the result corresponds well with measurement.

The commercial FEM CAD tool Ansoft HFSS [53] was evaluated for its appropriateness in the design of FSS and its accuracy verified with measured results. Several unique findings were discovered during the usage of this software for the design of FSS. Firstly, the measured frequency response of the FSS is in good agreement with the simulated results prior to the onset of grating lobes. Upon the onset of grating lobes, a higher transmission coefficient is observed from the simulated result. This finding indicates that there is a need to de-embed the grating lobe field from the neighbouring unit cell when the Master-and-Slave boundary condition is used to approximate the infinite array configuration. On the other hand, we realised that this software possess the capability of visualising and differentiating between the onset of grating lobe and the propagating grating lobe field phenomenon that are otherwise difficult to illustrate. A physical understanding to the edge singularity effect was achieved by correlating the theory with

the result presented in the form of a surface current plot. It has been shown that the mastery of the fundamentals developed in Chapter 2 is useful. This is essential in the analysis and design of all subsequent bandpass and bandstop FSS presented in this thesis.

To cater for the need of a bandpass radome for the GPS and DCS1800 antenna, low-pass FSS in the form of low frequency band-pass and high frequency band-stop FSS were investigated. Several novel design configurations arise from this work includes: 1) a single layer convoluted square loop (Maltese cross shaped) band-pass FSS with a significant passband bandwidth of 22.47%; 2) a quad- layer FSS design with ring and circular patch element for high level attenuation; and 3) a quad- layer staggered-tuned FSS with wide band-stop bandwidth. The admittance chart is used to design, visualize and interpret the changes of the multi-layer band-stop FSS. It has been shown that due to the relatively large element size at the higher frequencies, the low frequency band-pass FSS are prone to early onset of grating lobes and undesired transmission spikes at the higher harmonics. It is believe that this is the first time that low-pass FSS, both single and multi-layer as well as bandpass and bandstop, have been optimized and compared for this low frequency GPS and DCS1800 usage.

In terms of antenna design and analysis, a novel antenna (Maltese Cross-shaped) has been developed for circular polarization. The feed network requires the understanding of the coaxial balun network in terms of its equivalent circuit parameters and the coplanar waveguide theory, of which both have been presented in this thesis. It has been highlighted that a tradeoff in the VSWR performance was adopted to achieve a better matching of the phase (minimal phase difference) between the two input ports. The small phase difference helped to ensure a good CP axial ratio over the desired band. The consequence for this is that a good VSWR can only be obtained over a small bandwidth. Nevertheless, it has been demonstrated that the antenna gain, CP axial ratio bandwidth, and radiation pattern consistency of the Maltese Cross-shaped antenna are sufficiently attractive for GPS and GSM/DCS1800 band operations, although the VSWR can be further improved.

The integration of the antenna with the FSS is a very unique experience. Several novel phenomena discovered after the integration of the FSS and the antenna were being reported. Firstly, the trapped grating lobes within the FSS will exit and scatter at the finite edges of the FSS, giving rise to the high RCS observed at the extreme angles. Secondly, this phenomenon is more prominent for the TE polarization than the TM case. Similar to previous publications, the backscattering is reduced at out-of-band while the antenna performance is affected marginally in the in-band. It has also been brought to our attention through measurement and theoretical derivation that the FSS performance has a two-fold seemingly contradictory impact on the radar equations. For in-band operations, the reduction of the detection range as a result of the loss in the FSS is undesirable but for out-of-band operations, the reduction in antenna scattering as a result of the FSS inclusion is desirable. Two simplified expressions relating the detection range (r_{fss}) directly with 1) the loss of the FSS (L_{fss}) and 2) the RCS reduction ($\Delta\sigma$) as a result of the inclusion of FSS have been derived. The reductions pertaining to the Maltese Cross-shaped antenna with the quad layer high frequency bandstop FSS have been calculated to be 0.56% and 56.60% respectively. It is evident that the advantages of FSS inclusion outweighs the minute distance reduction due to its losses.

6.2 Further Work and Recommendations

For practical applications, the multi-layer FSS can be coated with a thin layer of structural laminates for protection against adverse environmental conditions. The thin layer laminate will most likely affect the electrical performance of the FSS and hence characterization of the loss and permittivity of the laminates might be required. A simple mutual coupling measurement and analysis can be performed on the entire antenna-cum-FSS structure to evaluate 1) the FSS shielding effectiveness against electromagnetic interference from a transmitting antenna placed within close proximity and 2) the reduction of coupling from the Maltese Cross-shaped antenna to a neighbouring antenna as a result of the reduced backscattering from the Maltese Cross-shaped antenna with the FSS inclusion. The analysis can be simplified to a two port S-parameters analysis

between the transmitting antenna and the receiving Maltese Cross-shaped antenna and vice versa.

For analytical work, from the admittance locus presented for the various band-stop FSS, the equivalent circuit parameters can be derived and input into an optimization or neural algorithm, of which it will serve as a priori information for the synthesis of FSS. It is then possible to obtain the desired parameters (size and spacing) of the FSS based on a desired transmission response requirement, instead of the time-consuming twiddling and trail and-error approach of optimizing the FSS parameters.

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APPENDIX A

DERIVATION FOR ANTENNA SCATTERING

Antennas can be the dominant scatterers for an integrated platform. It has been shown that the radar cross section (RCS) of an antenna due to its reradiation of an incident wave is comparable to the RCS of a flat plate with an area equal to the antenna effective aperture [16]. There is hence an increasing need for the scattering of the antenna to be controlled over specified frequency band, aspect angles and polarization without significant degradation of the antenna radiation performance. This forms the basis for the investigation into the relationship between the radiation performance and the scattering characteristics of the Maltese-cross antenna.

The relationship between the gain and RCS of an antenna has been derived [16]-[17], [90]-[91] by considering the antenna's reradiation properties. Referring to Figure A1, a plane wave having a power density of S_i is assumed to be illuminating an antenna with an effective aperture of A_e , the power captured by the antenna is given as:

$$P_r = S_i A_e \quad (\text{A.1})$$

For an isotropic antenna, the re-radiated power (from the antenna) at a distance r would be uniformly distributed over a sphere with surface area $4\pi r^2$. However, for antenna with gain G_r in a specified direction, the power density S_r in the specified direction exceeds the isotropic value by the gain factor G_r , i.e.

$$S_r = \frac{P_r}{4\pi r^2} G_r = \frac{S_i A_e}{4\pi r^2} G_r \quad (\text{A.2})$$

Rearranging (A.2),

$$A_e G_r = 4\pi r^2 \frac{S_r}{S_i} \quad (\text{A.3})$$

On the other hand, the radar cross section σ can be written as [16]:

$$\sigma = \lim_{r \rightarrow \infty} (4\pi r^2) \frac{|E_s|^2}{|E_i|^2} = \lim_{r \rightarrow \infty} (4\pi r^2) \frac{|S_r|}{|S_i|} \quad (\text{A.4})$$

Comparing (A.3) and (A.4) provides the backscattering due to the reflecting antenna with gain G_r , i.e.:

$$\sigma_{\text{ant}} = A_e G_r = (4\pi r^2) \frac{|S_r|}{|S_i|} \quad (\text{A.5})$$

In associating the incident waves (and hence the gain G_i of the incident antenna) with the effective aperture of the illuminated (reflecting) antenna, the effective aperture can be written as:

$$A_e = \frac{\lambda^2 G_i}{4\pi} \quad (\text{A.6})$$

Substituting (A.6) into (A.5), the antenna mode RCS σ_{ant} and thus be written as

$$\sigma_{\text{ant}} = \frac{\lambda^2}{4\pi} G_i G_r \quad (\text{A.7})$$

By incorporating the polarization mismatch factor p_i and p_r for the incident and reflecting antenna respectively, and the reflection Γ from the antenna termination, the antenna mode RCS can be further expressed as [90]:

$$\sigma_{\text{ant}} = \frac{\lambda^2}{4\pi} G_i G_r p_i p_r |\Gamma|^2 \quad (\text{A.8})$$

However, the gain G_r of the illuminated (reflecting) antenna can be expressed in terms of its own effective aperture and is given as:

$$G_r = \frac{4\pi A_e}{\lambda^2} \quad (\text{A.9})$$

By substituting (A.9) into (A.5), the resulting RCS can hence also be expressed as:

$$\sigma_{\text{ant}} = \frac{4\pi A_e^2}{\lambda^2} \quad (\text{A.10})$$

This corresponds to the RCS formula for a flat plate with normal incidence. It coincides with the assumption that the antenna is terminated by short circuit and all received power is reradiated. Taking into consideration of the absorption provided by a matched load, the RCS is given as.

$$\sigma = \frac{4\pi A_e^2}{\lambda^2} |\Gamma|^2 \quad (\text{A.11})$$

The structural [16] or residual [90] mode RCS of the antenna is then defined as the remaining scattering components (i.e. $|C|^2$) that must be added to the antenna mode RCS in order to obtain the total antenna RCS and it is given as [90]:

$$\sigma_{\text{res}} = \frac{\lambda^2}{4\pi} G_i G_r |C|^2 \quad (\text{A.13})$$

The total antenna RCS is then given as:

$$\sigma_{\text{tot}} = \frac{\lambda^2}{4\pi} G_i G_r |\Gamma + C|^2 \quad (\text{A.14})$$

In most cases, the RCS value is taken with respect to a target having an area of 1m^2 , i.e.

$$\sigma(\text{dBm}^2) = 10 \log \frac{\sigma(\text{m}^2)}{\sigma_{\text{ref}}} = 10 \log \frac{\sigma(\text{m}^2)}{1} \quad (\text{A.15})$$

Figure A1. RF Link Budget with Antenna Scattering.

APPENDIX B

DERIVATION FOR RANGE REDUCTION AND ITS RELATIONSHIP WITH FREQUENCY SELECTIVE ARRAY SHIELDING

The performance of the FSS is closely associated to the various parameters of the radar range equation. Although the derivation of the equation has well been studied [89], [92], it will be consolidated and presented here in a manner that highlights its relevance with FSS performance.

With a transmitter emitting power P_t , the transmitted power density at a distance r is given as:

$$S_t = \frac{P_t}{4\pi r^2} \quad (\text{W/m}^2) \quad (\text{B.1})$$

When the transmitter is replaced with an antenna having gain G_t in a specified direction, the transmitted power density is given as:

$$S_t = \frac{P_t G_t}{4\pi r^2} \quad (\text{W/m}^2) \quad (\text{B.2})$$

The total power intercepted by a target with area or RCS σ is given as:

$$P_t = S_t \sigma = \frac{P_t G_t}{4\pi r^2} \sigma \quad (\text{W}) \quad (\text{B.3})$$

When this power hits a target it will reradiate. The reflected power density detected back at the receiver (after undergoing another spreading loss of $1/4\pi r^2$) is given as:

$$S_r = \left(\frac{P_t G_t}{4\pi r^2} \right) \left(\frac{\sigma}{4\pi r^2} \right) \quad (\text{W/m}^2) \quad (\text{B.4})$$

The power received at an antenna with effective aperture A_e is given as:

$$P_r = S_r A_e = \left(\frac{P_t G_t}{4\pi r^2} \right) \left(\frac{\sigma}{4\pi r^2} \right) A_e \quad (\text{W}) \quad (\text{B.5})$$

The effective aperture of the received antenna can be expressed in terms of gain, i.e.:

$$A_e = \frac{\lambda^2 G_r}{4\pi} \quad (\text{m}^2) \quad (\text{B.6})$$

Thus the power received can be expressed as:

$$P_r = \left(\frac{P_t G_t}{4\pi r^2} \right) \left(\frac{\sigma}{4\pi r^2} \right) \left(\frac{\lambda^2 G_r}{4\pi} \right) = \frac{P_t G_t G_r \sigma \lambda^2}{(4\pi)^3 r^4} \quad (\text{B.7})$$

For monostatic RCS, $G_t = G_r$ and equation (B.7) can be expressed as:

$$P_r = \frac{P_t G_t^2 \lambda^2 \sigma}{(4\pi)^3 r^4} \quad (\text{W}) \quad (\text{B.8})$$

Rearranging (B.7), the detection range r can be expressed as:

$$r = \left[\frac{P_t G_t^2 \lambda^2 \sigma}{P_r (4\pi)^3} \right]^{\frac{1}{4}} \quad (\text{m}) \quad (\text{B.9})$$

The maximum detectable range r is determined by the minimal received signal strength P_r . When converted to decibels, equation (B.9) becomes:

$$r = 10^{\frac{1}{40} [P_t (\text{dBm}) + G_t (\text{dB}) + G_r (\text{dB}) + 20 \log(\lambda) + \sigma (\text{dBm}^2) - 30 \log(4\pi) - P_r]} \quad (\text{B.10})$$

In most applications, the power received is affected by the following parameters and this includes [93], [94]:

- the thermal noise power N ($N = kTB_n$, where k is Boltzmann's Constant 1.38×10^{-23} , T is the temperature and B_n is the noise bandwidth of the receiver),
- receiver's noise figure, F ($F = (S_{in}/N_{in})/(S_{out}/N_{out})$,
- additional losses L , due to pulse integration, scanning and beam pattern distortion and
- pulses-per-scan PPS ($\text{PPS} = \theta_s f_r / \omega$, where θ_s is the beamwidth of the scanning antenna, f_r is the radar pulse repetition frequency

The power received P_r (i.e. Eqn.B.8) can thus be expressed as signal-to-noise-ratio [93]:

$$\text{SNR} = \frac{P_r}{N} = \frac{P_t G_t^2 \lambda^2 \sigma \theta_s f_r}{(4\pi)^3 r^4 k T B_n F L \omega} \quad (\text{W}) \quad (\text{B.11})$$

When converted to decibels, equation (B.11) can be written as:

$$\begin{aligned} \text{SNR(dBm)} = & P_t(\text{dBm}) + G_t(\text{dB}) + G_r(\text{dB}) + 20 \log(\lambda) + \sigma(\text{dBm}^2) \\ & + 10 \log(\theta_s f_r) - 30 \log(4\pi) - 40 \log(r) - 10 \log(k T B_n) \\ & - F(\text{dB}) - L(\text{dB}) - 10 \log(\omega) \end{aligned} \quad (\text{B.12})$$

With all the parameters of Eqn. (B.9) remaining unchanged, a reduction of the antenna gain (G_t and G_r) as a result of the loss incurred in the FSS radome will reduce the antenna's detection range r . This is often undesirable, as it implies a reduction of the antenna's efficiency. For transmission, a one-way loss is incurred, while for the reception of the backscattered signal, the gain will be reduced by twice the loss encountered in the transmit path. Assuming that the gain of the transmit and receive antenna with the FSS inclusion is given as G_{tfss} and G_{rfss} respectively, then the ratio between the maximum detectable range r and the new detection range r_{fss} is given as:

$$\frac{r_{\text{fss}}}{r} = \left[\frac{G_{\text{tfss}} G_{\text{rfss}}}{G_t G_r} \right]^{\frac{1}{4}} \quad (\text{B.13})$$

But $G_{\text{tfss}} = (L_{\text{fss}})(G_t)$ and $G_{\text{rfss}} = (L_{\text{fss}})(G_r)$, where L_{fss} is the loss incurred in the FSS. Hence Eqn. (B.10) can be written as:

$$\frac{r_{\text{fss}}}{r} = \left[L_{\text{fss}}^2 \right]^{\frac{1}{4}} \quad (\text{B.14})$$

The range reduction will be given by:

$$\text{range_reduction}(\%) = \left(1 - \frac{r_{\text{fss}}}{r} \right) \times 100\% \quad (\text{B.15})$$

When expressed in decibels, the ratio between the maximum detectable range r and the new detection range r_{fss} can be written as:

$$\frac{r_{fss}}{r} = 10^{\frac{1}{40}[(G_{ifss}(dB) + G_{rfss}(dB)) - (G_i(dB) + G_r(dB))]} \quad (B.16)$$

For a loss of 0.5dB (i.e. linear value of 0.891) incurred by the FSS, the range would be reduced by 5.6%.

However, referring to equation (B.7) and (B.8) again, with all parameters remained unchanged, a reduction of the RCS σ would reduce the maximum detectable range of the target. This is desirable from the perspective of the targets detectability and survivability. Tilting of the FSS will provide the shaping required to reduce the backscattering or RCS σ in a specified direction [89], [92]-[94]. It should be highlighted that the detection range varies with the forth root of the RCS σ and with all parameters being the same, a reduction of RCS σ by a factor of 16 is required to halve the detection range. Similarly, the ratio between the maximum detectable range r and the new detection range r_{fss} in terms of RCS can be written as:

$$\frac{r_{fss}}{r} = [\Delta\sigma]^{\frac{1}{4}} \quad (B.17)$$

where $\Delta\sigma$ is a ratio of the linear value of the antenna RCS with and without the FSS inclusion. When expressed in decibels, the ratio for the range can be expressed as:

$$\frac{r_{fss}}{r} = 10^{\frac{1}{40}[\sigma_{fss}(dB) - \sigma(dB)]} \quad (B.18)$$