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HIGH-GAIN PLANAR RESONANT CAVITY ANTENNAS USING METAMATERIAL SURFACES

By

Shenhong WANG (BS, MS)

A Doctoral Thesis

Submitted in partial fulfilment of the requirements for award of the degree of Doctor of Philosophy by Loughborough University

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Abstract

This thesis studies a new class of high gain planar resonant cavity antennas based on metamaterial surfaces. High-gain planar antennas are becoming increasingly popular due to their significant advantages (e.g. low profile, small weight and low cost). Metamaterial surfaces have emerged over the last few years as artificial structures that provide properties and functionalities not readily available from existing materials. This project addresses novel applications of innovative metamaterial surfaces on the design of high-gain planar antennas.

A ray analysis is initially employed in order to describe the beamforming action of planar resonant cavity antennas. The phase equations of resonance predict the possibility of low-profile/subwavelength resonant cavity antennas and tilted beams. The reduction of the resonant cavity profile can be obtained by virtue of novel metamaterial ground planes. Furthermore, the EBG property of metamaterial ground planes would suppress the surface waves and obtain lower backlobes. By suppressing the TEM mode in a resonant cavity, a novel aperture-type EBG Partially Reflective Surface (PRS) is utilized to get low sidelobes in both planes (E-plane and H-plane) in a relatively finite structure. The periodicity optimization of PRS to obtain a higher maximum directivity is also investigated. Also it is shown that antennas with unique tilted beams are achieved without complex feeding mechanism. Rectangular patch antennas and dipole antennas are employed as excitations of resonant cavity antennas throughout the project. Three commercial electromagnetic simulation packages (Flomerics Microstripes™ ver6.5, Ansoft HFSS™ ver9.2 and Designer™ ver2.0) are utilized during the rigorous numerical computation. Related measurements are presented to validate the analysis and simulations.

Keywords — High-gain Planar Antennas, Subwavelength Resonant Cavity, Metamaterials, Arrays, Electromagnetic Band Gap, Artificial Magnetic Conductor
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Chapter 1 Introduction

1.1 Background

In recent years we have experienced a tremendous growth on the research and development of new wireless communication systems. These include 3G/4G mobile communications, broadband wireless LAN, fixed broadband wireless transmission (e.g. LMDS system) and novel satellite communication systems. All of those systems often require antennas with high directivity, broad bandwidth and high efficiency. Horn antennas and reflector antennas are possible candidates for this type of applications [1-3]. However their weight, profile and complex fabrication may pose problems in terms of cost, deployment and aesthetics. By comparison, planar antennas are becoming increasingly popular due to their significant advantages (e.g. low profile, small weight and low cost) and have found numerous applications to date. A popular example of high-directivity planar antennas is the microstrip array antenna [4][5]. The ability to implement microstrip array antennas on PCB (Printed Circuit Board) makes the integration of antennas with RF circuits possible. This feature would be valuable in designing compact RF transceivers. However the design of the
feeding for microstrip arrays may be very complicated and time-consuming. Moreover, the complex feeding-lines network may degrade the antenna efficiency and radiation performance.

An alternative type of high-directivity planar antenna is the high-gain planar resonant cavity (HPRC) antenna with a simple feeding mechanism proposed by Giswalt Von Tretini in 1956 [6]. Trentini observed that a Partially Reflective Sheet (PRS) array significantly enhances the directivity when placed in front of an antenna. In 1989 J.R. James further studied the topic and more detailed results have been published [7]. Waveguide-slot antennas have been used as the excitations in their antenna configurations. A brief literature review of HPRC antennas will be given in section 1.2.

Over the last few years, there has been a lot of research on metamaterial structures. The terminology ‘meta’ of metamaterials means ‘beyond’ in Greek [8]. It mainly indicates those materials non-existed in nature. Here ‘non-existed’ only means that they cannot be directly found in the natural world, but they can be created by utilizing existing materials. Metamaterials have found several applications in electromagnetics, including RF and photonic engineering [9][10]. Metamaterials are typically composed of passive periodic metallic-dielectric structures. A well-known metamaterial structure is a planar passive periodic array, known as Frequency Selective Surface (FSS). Other metamaterial structures include Photonic Bandgap (PBG) and Electromagnetic Bandgap (EBG). More new terminologies have been invented based on different electromagnetic properties of metamaterials, such as Artificial Magnetic Conductor (AMC), High-impedance Surface, and Left-handed Materials. A brief literature review on metamaterials will be presented in section 1.3.

The aim of this thesis is to incorporate novel metamaterial surfaces into the configuration of HPRC antennas and produce novel antenna designs. The different properties of the metamaterial surfaces are utilized to obtain performance enhancement for HPRC antennas. A microstrip or dipole antenna is used as the new excitation of the resonant cavity to complete the prototype design. A detailed description of the main contributions will be presented in section 1.4.
1.2 High-gain Planar Resonant Cavity Antenna Review

A HPRC antenna consists of two parallel planes that form a planar resonant cavity as illustrated in Figure 1.1 and Figure 1.2. The bottom layer is a fully reflective plane (e.g. a ground plane) and the upper layer is a plane with a high reflection coefficient. The distance between the planes ensures that the propagating waves in the cavity have a desired phase difference when the waves reach the upper plane. The angle of the high-directivity main beam to the normal direction is related with the phase difference. Due to the non-full reflection of the upper plane, the waves in the resonant cavity would have some losses when they propagate in the resonant cavity. The radiation patterns of the antenna are determined by the losses (leaky-waves). Thus the antenna is essentially a kind of leaky-wave antenna. The leaky-waves from the upper plane form a main beam at desired directions when the related resonant conditions are met. The related resonant conditions and the discussion of the phase difference will be presented in section 2.2.1.

The far field pattern of the antenna is determined by the leaky waves. The specific characteristic of the leaky wave is its complex propagation constant $\alpha + j\beta$, where $\alpha$ indicates loss and $\beta$ is the propagation constant. The original leaky-wave antenna is a uniform losses waveguide with a continuous longitudinal slot as shown in Figure 1.3 [11]. The travelling wave in the waveguide propagates with a phase velocity greater than that of light and attenuates as it travels. The attenuation indicates a continuous leakage of energy.
The far-field radiation pattern is characterized by three main features: the angle of the main beam to the normal direction, its 3dB beam-width and the distribution of sidelobes. An approximate expression of the main-beam angle can be defined into \( \cos \phi \approx \frac{\beta}{k_0} \), where \( k_0 \) is the free-space wave number and \( \beta \) is the propagation constant. The 3dB beam-width is found be related with the leakage rate \( \alpha \), a measure of the power leaked away per unit length along the waveguide. If \( \alpha \) is small, it means that the leakage is small and the effective aperture length of the leaky-wave antenna is large. The far-field radiation pattern then has a narrow beam-width. On the other hand, if \( \alpha \) is large, the power rapidly leaks away, then the effective aperture is short and the radiation pattern has a wide beam. The theory of complex waves has been discussed extensively in the literature [12]. To obtain high-directivity, a waveguide-fed parallel plate slot array antenna was tried in recent years [13]

A configuration of HPRC antennas based on dielectric materials is shown in Figure 1.1. It comprises only two dielectric slabs. The bottom layer is a PEC (Perfect Electrical Conductor) grounded dielectric plane. The leaky wave from the upper dielectric plane forms a main directive beam at broadside when the related resonance equations are met, as reported in [14]. A leakywave analysis of the above structure was carried out by D. Jackson et al. [15]. However it may be difficult to obtain high directivity for such a structure. For example, to design a HPRC antenna with a directivity of more than 20 dB, the antenna configuration needs a multi-layer structure or the upper dielectric plane with very high permittivity/ permeability (more than 100) [14]. The multi-layer structure will make the antenna’s profile very large. The high
permittivity/permeability will introduce a high loss in the dielectric plane. The high loss will reduce the antenna's efficiency.

Figure 1.2  HPRC antennas by Giswalt Von Tretini and J.R.James (from [6] and [7])

An alternative configuration of HPRC antennas based on metallic-dielectric planes is shown in Figure 1.2. The bottom layer is a PEC grounded plane and the upper layer is a FSS plane with high reflection coefficient. The FSS is a metallic array printed on a dielectric substrate. When the propagating wave in the resonant cavity reaches the FSS plane, only a small part of electromagnetic waves radiate out as leaky-waves and the rest is reflected back into the resonant cavity. The leaky waves from the FSS plane will form a high-directivity main beam if certain conditions are met. The conditions will be discussed in section 2.2.1. Giswalt Von Tretini has observed the phenomena that the directivity of the antenna can be hugely improved when a FSS plane is placed in the front of the antenna [6]. Subsequently J.R.James further studied similar antennas [7].
Later, with the emerging of PBG/EBG structures, more papers discussed the usage of PBG/EBG planes to obtain high-directivity antennas or wave-focusing [16-21]. All of them are based on the theory of resonance or leaky-wave and some new concepts or terminologies have been used.

1.3 Metamaterials Review

The majority of metamaterials for microwave applications could be implemented by variations of passive periodic metallic-dielectric structures also known as FSS. An example of planar FSS is shown in Figure 1.4, where conducting-type and aperture-type elements are depicted. In the figure the black area represents metal and the grey represents dielectric substrates. A classic paper about FSS has been written by Raj Mittra [22] and there are four special monographs on the topic [23-26].

The reflection magnitudes of conducting-type FSS planes increase with frequency and the FSS plane is totally reflective at resonance. The reflection phases of the conducting-type FSS planes decrease with frequency until it becomes $-\pi$ at resonance. The reflection magnitudes of aperture-type FSS planes decrease with frequency until resonance, where the FSS plane is totally transparent. The reflection phase, in turn, also decreases with frequency until it takes a zero value at resonance.
A variety of geometries can be selected as the unit cell structures, such as dipole, cross dipole, square patch and loops. Different geometries and dimensions have different reflection/transmission responses at different frequencies. A specific geometry can be selected depending on different application or requirements. In some cases double layer or multi layer FSS may be used to get the desired performance.

The terminology FSS has been widely used in microwave community while the term PBG has been widely applied in optical community. The terminology EBG is more widely used recently in the microwave community [27]. The well-known character of EBG structure is its band gaps (stop or pass bands) in the frequency domain. Dispersion diagrams can be utilized to characterize the band gaps of EBG structures. An introduction of dispersion diagrams will be presented in section 2.3.2.

An ideal EBG would be a 3-D periodic structure that can forbid the propagation of electromagnetic waves in a specified frequency band for all angles and for all polarization states. However a number of EBG structures for microwave applications, such as the suppression of surface waves, belong to 2-D or 2.5-D planar structures. These structures will be studied in this thesis.

A well-known EBG structure is the mushroom-type structure shown in Figure 1.5. The structure was developed by Dan Sievenpiper from UCLA [28]. It contains single or double layer metal FSS and each element is connected to the ground plane by a via. A new terminology “High Impedance Surface” for the structure has also been used in some applications. Due to its high impedance, surface current can not propagate along the structure. Thus it behaves as an Artificial Magnetic Conductor (AMC) according to Maxwell’s equations, where the surface electric current equals to zero. This is a
mathematical idea that is used in certain electromagnetic problems, but does not exist in reality. The key property of AMC is that the reflection phase of an AMC surface is around zero degrees when illuminated with an incident plane wave, while compared with 180 degrees of PEC (Perfect Electrical Conductor). An equivalent parallel L-C resonant circuit model can be utilized to model the electrical character of the structure. The capacitance is decided by the gap between the patches and the inductance is decided by the length of the via. At the resonant frequency $f_0$ of the circuit, the surface impedance would be $\infty$. Thus no surface current is yielded and the reflection phase is zero degree.

![Figure 1.5](image)

(a) Bird's eye view

(b) Cross-section view

Figure 1.5 Mushroom-type EBG surfaces by Dan Sievenpiper (from Ref [28])

1.4 Aim and Overview of the Thesis

The project focuses on the study of novel high-directivity planar resonant cavity antennas using metamaterial surfaces. In particular, the following contributions have been made:

- New excitation techniques: waveguide-slot antennas may be difficult to be fabricated and expensive, particular at microwave frequencies. Rectangular patch antennas and dipole antennas are utilized as the new excitations of low-profile HPRC antennas [29][30].

- Reduced profile: Ray analysis is employed to describe the beam forming function of the planar resonant cavity antenna. The phase equations of resonance predict the possibility of low-profile (including subwavelength)
resonant cavity. Low-profile HPRC antennas are obtained by employing new metamaterial ground planes [29-31].

- Low backlobes: The usage of metamaterial ground planes (AMC) for low-profile HPRC antennas leads to higher backlobes because there is the strongest field with the conducting layer of AMC ground planes. The mushroom-type EBG metamaterial ground plane is utilized for reduction of the backlobes by suppressing the surface waves.

- Low sidelobes in a finite antenna structure: HPRC antennas (in a relatively finite structure) often show higher sidelobes in H-plane than in E-plane. Theoretical analysis deduces that the existence of TEM mode in the resonant cavity contributes to the problem. A novel aperture-type EBG surface (working as PRS) is utilized to suppress the TEM mode and achieves low sidelobes in both planes. At the antenna operating frequency, the aperture-type EBG surface has a high reflection coefficient at normal direction and a TM bandgap along its surface [32].

- Higher efficiency: The periodicity of PRS can be optimized to obtain a higher maximum directivity and the result will help to enhance the HPRC antenna's aperture efficiency.

- Single tilted beams: The ray analysis also predicts the possibility of tilted beams. A novel configuration of HPRC antennas is designed to obtain a unique tilted beam. The beam angle can be steered by changing the distance between the PRS and the ground plane. The HPRC antenna is matched using a rectangular patch antenna as the excitation.

Extensive research has been initially carried out on the theoretical analysis and practical design of new metamaterial surfaces such as AMC ground planes, metamaterials ground planes with negative reflection phases and EBG surfaces. Subsequently they have been applied to the design of HPRC antennas. The physical properties of those metamaterial surfaces and HPRC antennas have been studied using analytical and numerical simulation methods to validate the theoretical results. Three commercial electromagnetic simulation packages (Flomerics Microstripes™ ver6.5, Ansoft HFSS™ ver9.2 and Ansoft Designer™ ver2.0) have been utilized during the
rigorous numerical computation. Related measurements are presented to validate the analysis and simulation results. The thesis is organized as follows:

Chapter 2 introduces the related theory analysis and simulation methods. Ray analysis is utilized to analyze the beam forming function of HPRC antennas. The result predicts the possibility of low-profile (including subwavelength) resonant cavity and tilted beams. Continuing that is the analysis of metamaterials (including AMC and EBG). Subsequently it is found that a novel aperture-type EBG surface can be used to suppress the TEM mode in the resonant cavity. Three commercial simulation packages (Flomerics Microstripes™ ver6.5, Ansoft HFSS™ ver9.2 and Ansoft Designer™ 2.0) are briefly introduced.

Chapter 3 studies the design of low-profile resonant cavity antennas by employing metamaterial ground planes. Firstly a parametric study for an AMC is carried out before it is employed as the ground plane of the HPRC antenna to reduce its original half-wavelength resonant cavity to quarter-wavelength. A rectangular patch is utilized as the excitation of HPRC antennas because it provides more flexibility to match the input impedance. Subsequently a metamaterial ground plane with negative reflection phase is utilized to further reduce the resonant cavity profile up to subwavelength values. In order to exempt the effect of removing the central elements to accommodate the rectangle patch, a dipole is used as the excitation of HPRC antennas. Simulations show that the change yields a higher directivity compared with the result of the HPRC antenna fed by a rectangular patch. The input impedance of the HPRC antenna can be matched by optimizing the length of the dipole and the distance from the dipole to the metamaterial ground plane. A distance $\lambda/6$ between the PRS and the metamaterial ground plane is achieved based on simulations and measurements. Finally it is shown that a PRS with higher reflection coefficients can be utilized to enhance the maximum directivity of subwavelength HPRC antennas.

Chapter 4 is interested in the suppression of related propagation modes and averting grating lobes to get higher efficiency HPRC antennas. Initially low-profile HPRC antennas based on mushroom-type EBG ground planes are investigated after a parametric study on the diameter of the via. The suppression of the AMC surface waves leads to lower backlobes for HPRC antennas. By virtue of the application of a
novel aperture-type EBG surface, HPRC antennas (in a relatively finite structure) show low sidelobes in both planes (E-plane and H-plane). At the operating frequency of antennas, the aperture-type EBG surface has a high reflection coefficient in the normal direction and a TM bandgap along its surface. A conducting-type EBG surface with TE bandgap is utilized as a comparison. Simulation results show that a significant reduction of sidelobes in H-plane is achieved. The periodicity analysis of PRS will help meliorate the aperture efficiency of HPRC antennas.

Chapter 5 focuses on the design of novel HPRC antennas with tilted beams. According to the complex wave analysis in [12], a metallic wall is utilized to block the wave propagation along one side of the resonant cavity. Thus the HPRC antenna is able to obtain a single tilted beam. The beam angle can be steered by changing the distance between the PRS and the ground plane, according to the ray analysis results in section 2.2.1. Subsequently a rectangular patch is used as the excitation of the novel resonant cavity to match the HPRC antenna.

In Chapter 6 a summary of the previous discussions is drawn and future work is discussed.
References


Chapter 2 Theory Analysis and Simulation Methods

2.1 Introduction

This chapter will cover the background theory and simulation methods used for the design of metamaterials-based HPRC antennas. It also includes a brief introduction of simulation software packages used throughout the thesis. The discussion comprises three sections: analysis of HPRC antennas, analysis of metamaterials surfaces and simulation methods.

In section 2.2 ray analysis is used to describe the high-directivity beam forming function of HPRC antennas. Phase equations are formulated to describe the resonance that leads to the beam forming. Then by mathematically transforming the phase equations we can extend them in order to describe several variations of the HPRC antenna: quarter wavelength HPRC antennas, sub-wavelength HPRC antennas and HPRC antennas with tilted beams. The discussion of array periodicity and grating lobes will help to design higher efficiency HPRC antennas.
In section 2.3 two kinds of metamaterials are discussed: Artificial Magnetic Conductor (AMC) (including metamaterial ground planes with negative reflection phases) and Electromagnetic Bandgap (EBG). Field distribution along the normal direction to the AMC surface is plotted out. It is observed that there is the strongest field at the conducting layer of the AMC. To do a comparison, the field distribution for a PEC surface is also plotted. Theoretical analysis shows that a novel aperture-type EBG surface with a high reflection coefficient in the normal direction and TM bandgap along its surface can be obtained.

In section 2.4 three commercial electromagnetic simulation packages (Flometries Microstripes™ ver6.5, Ansoft HFSS™ ver9.2 and Ansoft Designer™ ver2.0) are briefly introduced. Generally Microstripes™ is used to simulate the proposed antennas in its entirety because of its 3-D user interface and efficient computation capabilities. HFSS™ is utilized to get the dispersion diagrams of periodic structures and analyse EBG structures. Designer™ is particularly used for analysing 2D or 2.5 D planar periodic structures.

2.2 Analysis of HPRC Antennas

2.2.1 Ray analysis of beam forming

Geometrical optics is concerned with the analysis and synthesis of optical systems to the approximation that diffraction and interference can be neglected. However, geometrical optics, although certainly an approximation, is still sufficiently accurate to produce meaningful and useful results, even for antennas with equivalent aperture dimensions as small as five wavelengths [1]. In 1965 the geometrical optics (also called ray optics) has been used to mathematically discuss the leaky-wave behaviours of HPRC antennas [2], where it was demonstrated how one Partially Reflective Sheet (PRS) array can significantly improve the directivity of an antenna when parallel to the ground plane as illustrated in Figure 2.1. Here a brief review of the ray analysis for this type of antennas is given and subsequently it is extended for the purposes of the thesis.
In Figure 2.1, $l$ is the distance between the planes and $\alpha$ is the ray angle to the normal direction. Let the reflection coefficient of the PRS be $p \times e^{j\psi}$, the electric field strength of the far-field comprises the vector sum of those leakage rays and can be mathematically expressed as:

$$E = \sum_{n=0}^{\infty} f(a) E_0 p^n \sqrt{1 - p^2} e^{j\varphi_n}$$  \hspace{1cm} (1)$$

where $f(a)$ is the radiation pattern of the source in the resonant cavity. The phase angle $\varphi_n$ is the phase difference between the different rays, e.g. $\varphi_1$ is the phase difference between ray1 and ray2. From the ray propagation path in Figure 2.1, the value of $\varphi_n$ can be calculated:

$$\varphi_n = n\Phi = n\left[\frac{-2\pi}{\lambda} l \cos \alpha + \pi + \psi\right]$$  \hspace{1cm} (2)$$

where $\Phi$ is the phase difference between the neighboured rays.

Since $p<1$, the absolute value of the above electric field becomes
\[ |E| = |E_0| f(a) \sqrt{\frac{1-p^2}{1+p^2-2p\cos\Phi}} \]  

(3)

To obtain the maximum power strength at the broadside, the above equation needs \( \cos\Phi = 0 \) which means

\[-\frac{4\pi}{k} l \cos\alpha + \pi + \psi = 2N\pi \quad N=0,1,2\ldots \]  

(4)

Let \( \alpha = 0 \) to achieve a high directivity at broadside, hence the resonant cavity distance \( l \) can be derived

\[ l = \left( \frac{\psi}{2\pi} + 0.5 \right) \frac{4}{k} + N \frac{4}{k} \quad N=0,1,2\ldots \]  

(5)

Since \( \psi \) almost equals to \( \pi \) when PRS is nearly fully reflective, the distance \( l \) will be half wavelength or times of half wavelength according to the equation (5).

An alternative approach to model this type of antenna is based on leaky-wave theory. According to the theory of leaky-wave antennas on waveguides (as introduced in section 1.2), it is expected that the leaky waves travelling in the resonant cavity (as shown in Figure 2.1) will generate a main-beam whose angle to the normal direction is \( \alpha \). The angle \( \alpha \) is decided by the propagation constant \( \beta \) of the wave in the resonant cavity: \( \sin \alpha \approx \frac{\beta}{k_0} \), where \( k_0 \) is the wave number of free space [3]. At any time the propagation constant \( \beta \) should not be zero. It means that the angle \( \alpha \) would not be zero at any time as well. When the resonant cavity is fed by a symmetrical source (e.g. a dipole antenna) at the centre, the wave in the resonant cavity will propagate towards all directions and may yield two main-beams at the same time.

In order to make sure that all rays will be at a same phase when they reach the upper layer and the main beam is at the broadside, the propagation path outside of the resonant cavity \( 2l \times t g \alpha \times \sin \alpha \) will not be calculated into the phase difference. Thus the modified mathematical expression of phase equations would be
\[ \varphi_n = n\Phi = n\left[-\frac{4\pi}{\lambda} \times \frac{1}{\cos \alpha} + \pi + \psi\right] \] \hspace{1cm} (6)

To achieve a maximum power strength at broadside, it needs \( \cos \Phi = 0 \) and \( \cos \alpha \to 1 \) which means

\[ -\frac{4\pi}{\lambda} \frac{1}{\cos \alpha} + \pi + \psi = 2N\pi \hspace{0.5cm} N=0,1,2\ldots \] \hspace{1cm} (7)

Hence the modified resonant cavity distance can be derived

\[ l = \cos \alpha \left[\left(\frac{\psi}{2\pi} + 0.5\right) \frac{\lambda}{2} + N \frac{\lambda}{2}\right] \hspace{0.5cm} N=0,1,2\ldots \] \hspace{1cm} (8)

When a high-directivity main beam is achieved at broadside, the angle \( \alpha \) would be very small and \( \cos \alpha \to 1 \). For example, the angle \( \alpha \) is only about 8° when the high directivity is about 20dB. Therefore the above equation (8) has no difference in essence from the previous equation (5).

By mathematically transferring the above phase equations (8) and (4) into different expressions, several theoretical results can be obtained with regard to the design of novel HPRC antennas:

\textit{A. Quarter-wavelength HPRC Antennas}

For a highly reflective PRS (\( p = 1 \) and \( \psi(0) = -\pi \)), the resonant distance \( l \) is about half wavelength (or times of half wavelength) to steer the high-directivity main beam at broadside according to the equation (4) or (8). If a metamaterial ground plane with reflection phase \( \phi(\alpha) \) is introduced in the antenna configuration, then the above phase equation (8) would be changed into

\[ l = \cos \alpha \left[\left(\frac{\psi + \phi(\alpha)}{2\pi}\right) \frac{\lambda}{2} + N \frac{\lambda}{2}\right] \hspace{0.5cm} N=0,1,2\ldots \] \hspace{1cm} (9)
If the metamaterial ground plane is an AMC surface whose reflection phase is zero, then the distance would become \( l \approx \frac{\lambda}{4} \). So the whole antenna profile might be approximately reduced from half wavelength to quarter wavelength. This theoretically predicts the possibility of a quarter-wavelength HPRC antenna. Detailed simulations and measurements will be presented in section 3.2.

**B. Sub-wavelength HPRC Antennas**

By changing the reflection phase \( \phi(\alpha) \) (e.g. \(-10^9\)) of the metamaterial ground plane and keeping all other parameters constant, we obtain a smaller distance \( l \) according to the phase equation (8). Therefore it is expected that the whole profile of the resonant cavity could be further reduced and a sub-wavelength HPRC antenna might be possible. By further reducing the reflection phase of the metamaterial ground plane, the distance \( l \) would become very small. At last the whole antenna configuration may comprise only two layered dielectric slabs incorporating metallic arrays. The detailed research will be presented in section 3.3.

**C. Tilted Beam HPRC Antennas**

When the HPRC antenna is fed by symmetrical source (e.g. a dipole antenna) and achieves a maximum directivity at broadside, the main beam actually consists of two separate beams and the angle between them is very small [5]. The separate beams are respectively formed by the leaky waves propagating in the resonant cavity (one side for one beam). The beam angle \( \alpha \) to the normal direction can be calculated by writing the equation (8) as follows:

\[
\cos \alpha = \frac{l}{\left[\left(\frac{\nu}{2\pi} + 0.5\right)\frac{\lambda}{2} + N\frac{\lambda}{2}\right]}
\]

(10)

So to obtain a maximum directivity at broadside, the actual distance \( l \) should be slightly less than the wave propagation path and the beam angle between the separate tilted beams:

\[
\alpha = \arccos \left\{ \frac{l}{\left[\left(\frac{\nu}{2\pi} + 0.5\right)\frac{\lambda}{2} + N\frac{\lambda}{2}\right]} \right\}
\]

(11)
To obtain symmetrical tilted beams at angle $\alpha$ to the normal direction, there is phase difference between the neighboured rays when they reach the upper layer. And we can transfer the previous phase equation (2) into

$$\cos \alpha = \lambda \times [(2N + 1)\pi + \psi]/(4\pi l)$$

(12)

and the beam angle

$$\alpha = \arccos \left\{ \lambda \times [(2N + 1)\pi + \psi]/(4\pi l) \right\}$$

(13)

Thus it is predicted that HPRC antennas with symmetrical tilted beams are possible when the distance between PRS and ground plane is larger than the desired resonant distance to achieve a main-beam at broadside. In order to obtain a unique tilted beam, a novel solution is needed since all excitations (e.g. dipoles and rectangular patches) in the thesis would be symmetric and belong to ‘bi-excitation’ which means the radiated waves can propagate along both sides of the resonant cavity [5]. If the excitation is omni-directional, then a HPRC antenna with bow-tie radiation pattern could be obtained [7]. Related investigation will be done in Chapter 5.

2.2.2 Periodic aperture array and grating lobes

In the previous section an approximate equation $\frac{1+p}{1-p}$ of directivity enhancement is derived, where $p$ is the reflection coefficient of PRS, when the phase equation (4) or (8) is met. This section will discuss the optimisation of PRS periodicity to design a higher directivity HPRC antenna for a fixed $p$.

A. Periodic aperture arrays

Since the above PRS plane is a periodic aperture array, the whole antenna might be thought as a periodic aperture array antenna. When the array antenna is fed by plane waves, its mathematic model can be expressed by a discrete-time rectangular pulse:

$$E[n] = 1 \text{ when } n < |N_1| \text{ and } 0 \text{ for others, where } N_1=1,2,3\ldots$$
as illustrated in Figure 2.2(a) with 5 elements \( N_i = 2 \). Each unit cell indicates an element of periodic aperture array antenna. Thus its far-field radiation pattern can be derived from Huygens principle by using the Fourier Transform and are plotted in Figure 2.2(b) when \( N_1 = 1, 2, 3 \).

\[
F(\theta) = \frac{\sin(\theta \times (N_i + 0.5))}{\sin(\theta / 2)}, \text{ where } \theta \text{ is angle and } N_1 = 1, 2, 3, \ldots
\]

The results clearly show that more unit cells will yield a sharper main-beam and a higher directivity could be achieved.

Figure 2.2 (a) Field distribution
When the HPRC antenna is fed by a single source (e.g., a dipole antenna), the field in the resonant cavity may be expressed into $E = e^{-\alpha|z|}$ [8], where $\alpha$ is the loss, $\beta$ is the propagation constant and $z$ is the propagation distance from the source. The related mathematic model of the field distribution would be:

$$E[n] = e^{-\alpha|n|}, \text{ where } n = 0, 1, 2, 3... \text{ and } \alpha \text{ is the leakage rate}$$

as shown in Figure 2.3(a) with $n=10$ and $\alpha = 0.2$. Each unit cell indicates an element of the periodic aperture array antenna.

Following the similar way above, the far-field radiation pattern in the Fourier Transform may be derived

$$F(\theta) = \frac{1-e^{-2\alpha}}{1-2e^{-\alpha} \cos \theta + e^{-2\alpha}}$$

and plotted in Figure 2.3(b) when $\alpha = 0.1/0.15/0.2$ respectively.
Figure 2.3 HPRC antennas fed by a single source
Figure 2.3 shows that the far field radiation patterns are decided by the leakage rate $\alpha$, not the number of elements (unit cell) $n$ when the HPRC antenna profile is infinite. However, by using a small periodicity and hence more unit cells in a limited area, we may have a more regular field distribution that is expected to improve the directivity as well as the sidelobe level for HPRC antennas. Further simulations will be presented in section 4.4.

B. Grating lobes

Meanwhile a smaller periodicity would help to avert grating lobes in a periodic array. Here only the simple one dimensional case will be discussed. Figure 2.4 shows a plane wave incident upon a one-dimensional periodic structure with a periodicity $D$. The incident angle is $\theta_1$. Then the phase delay for each element is $D \sin \theta_1$ with respect to its neighbour (to the left). All reflected waves from the elements should be in phase so that we can obtain plane waves that propagate in the directions as shown in Figure 2.4. However there are possibly other directions where propagation can take place as illustrated in Figure 2.5.

![Figure 2.4 Reflection of periodic arrays](image1)

![Figure 2.5 Grating lobes of periodic arrays](image2)
Here the incident wave is the same as before and a grating lobe direction is denoted by $\theta_g$. The figure shows that the total phase delay for an element with respect to its neighbour to the left is about $\beta D(\sin \theta_i + \sin \theta_g)$, where $\beta$ is the propagation constant of wave. If the delay equals to a multiple of $2n\pi$ ($n$ is integer number), all waves from the elements will be in phase in the direction $\theta_g$. Following the ray propagation path we can derive the condition of generating grating lobes:

$$
\beta D(\sin \theta_i + \sin \theta_g) = 2n\pi \quad \text{where } ... \quad \beta = \frac{2\pi}{\lambda}, D \text{ is periodicity}
$$

and the frequency of generating grating lobes

$$
\frac{1}{f_g} = \frac{c}{\lambda_g} = \frac{nc}{D(\sin \theta_i + \sin \theta_g)}
$$

In the configuration of HPRC antennas, there are two factors to be considered in calculating the periodicity to avert grating lobes. One is the dielectric slab. When the wave propagates in a dielectric slab with permittivity $\varepsilon_r$, then the wavelength in the dielectric slab would be $\frac{\lambda}{\sqrt{\varepsilon_r}}$. In the configuration of HPRC antennas, the periodic EBG surface (FSS) is printed on one side of the dielectric slab, another side of the FSS is free space. A simple calculation of the approximate permittivity is $\frac{1+\varepsilon_r}{2}$ [9]. Thus the equation of yielding grating lobes would become

$$
\frac{1}{f_g} = \sqrt{\frac{1+\varepsilon_r}{2}} \frac{c}{\lambda_g} = \sqrt{\frac{1+\varepsilon_r}{2}} \frac{nc}{D(\sin \theta_i + \sin \theta_g)}
$$

The other is that the above derivation is based on ray analysis of plane waves. However the radiated waves by a practical source (e.g. a patch antenna) are not plane waves. Therefore an approximate incident angle should be used. The value is equal to the beam angle of the main beam formed by the leaky-waves ($\alpha = \arcsin \frac{\beta}{k_0}$, where $\beta$ is the propagation constant and $k_0$ is the wave number of free space) in the resonant
cavity. When a high directivity is achieved, the beam angle is very small (e.g. 7 – 8 degrees for a directivity about 22dB) from the paper [5].

The analysis results of periodic aperture array and grating lobes indicate that a smaller periodicity is expected to result in a better performance for HPRC antennas. However, if the periodicity is too small, it will introduce some new problems. For example, it would be very sensitive to the fabrication errors.

It must be noticed that the periodicity and the reflection coefficient of PRS are not all the factors that decide the maximum directivity of HPRC antennas. An example on that will be presented in section 4.4.

2.3 Analysis of Metamaterial Surfaces

Electromagnetic metamaterials can be generally divided into three classes: Artificial Magnetic Conductor, Electromagnetic bandgap and Negative Refraction Index structures. AMC and EBG surfaces will be discussed here and utilised in the HPRC antenna design. Another kind of metamaterial ground planes with negative reflection phases, which is an extension of AMC surfaces, will also be discussed and applied in the antenna.

2.3.1 Artificial Magnetic Conductor (AMC)

The term AMC stems from the zero reflection phase property of metamaterial planes when illuminated with a normal incident plane wave. That makes it show magnetic character although there is no real magnetic material. This is compared with the character of Perfect Electrical Conductor (PEC) that has 180 degrees phase shift when illuminated with a normal incident plane wave. Generally the reflection phase from −90 degrees to +90 degrees is considered as the working frequency band of AMC. It has been demonstrated that a mushroom-type structure could exhibit EBG and AMC at the same frequency band [10]. This is a 3-D structure and needs a via in the centre of each unit cell. Vias are used to connect the conducting patches to the ground plane (see section 1.3). The existence of via makes it much difficulty to fabricate the structure. However it has been found that via is not definitely required to make an
AMC surface [11]. Investigation results show that any FSS planes can show AMC character when printed on a grounded dielectric substrate.

To design an AMC surface, all of the three simulation software packages (Microstripes™ ver6.5, HFSS™ ver9.2 and Designer™ ver2.0) can be used respectively. The brief introduction of them will be presented in section 2.4. Here Flormerics Microstripes™ ver6.5 is used to obtain the field distribution of the unit cell structure as shown in Figure 2.8. To do a comparison, the dimension of the unit cell structure is completely the same as the one used by [11]. The periodicity of the unit cell structure is 10mm × 10mm. The square patch size is 9.5mm × 9.5mm. The thickness of dielectric substrates is 1.57mm with the permittivity 3.5. The problem space is 10mm × 10mm × 100mm as shown in Figure 2.6. The unit cell of the AMC is also presented in the figure. The grey area is dielectric and the black area is copper. The periodic boundary conditions, namely ‘Wrap boundary’ in Microstripes™ is selected in order to simulate only a single unit cell of the periodic structure. The bottom boundary is ‘Electrical Wall’ and the top boundary is ‘Absorbing’.

The simulated reflection phase as shown in Figure 2.7 has a good agreement with the measured result in [11], whose 1st frequency mode with zero reflection phases is around 5.85GHz. The paper demonstrates that the reflection coefficient has much loss at the 2nd mode which has the zero reflection phase property. Our simulation result shows that the reflection coefficient has some loss at the 1st mode as shown in Figure 2.7. However the simulated loss at 1st mode is only 0.21dB in Microstripes v6.5™ and 0.4dB in Designer v2.0™ while the simulated loss at 2nd mode (around 19GHz) in Designer™ is less than −6dB and the measured loss at 2nd mode is about 14dB. Here we use Ansoft Designer™ to compare with Microstripes™ and get the simulation at 2nd mode since Microstripes™ doesn’t work well on the frequency band. Simulations indicate that Microstripes™ and Designer™ can have good agreements with measurement on 1st mode. The reflection coefficient of a ground plane is also simulated to do a comparison and its value is −0.001dB as shown in Figure 2.7 (a). Section 3.3 will demonstrate that the loss doesn’t affect the application of AMC on low-profile HPRC antennas since the value is very small (about 0.2dB).
Subsequently field distributions along the normal direction to the AMC and PEC surfaces are plotted out as shown in Figure 2.8. Four frequency points are selected. At 5GHz the AMC surface has a reflection phase 135 degrees. At 5.831 GHz the AMC surface has a maximum loss (about 0.21dB). At 5.930GHz the AMC surface has zero reflection phases. At 7GHz the AMC surface has a negative reflection phase about −110 degrees. The clear boundary for different wavelengths along the z axis can be observed in the figure. It is also observed that the strongest field strength is at the conducting layer of AMC surfaces. This is completely different from the result of PEC surface.

Figure 2.7 shows that reflection phase of AMC surfaces is similar to a cosine function. It shows a negative reflection phase after the zero reflection phases. Thus a metamaterial ground plane with negative reflection phase as requested by section 2.2.1 can be obtained. The metamaterial ground plane is an extension of AMC surfaces. Application of AMC surfaces (including metamaterial ground planes with negative reflection phases) on HPRC antennas will be discussed in Chapter 3.

Figure 2.6 Geometry model of AMC unit cell and problem space
Figure 2.7 Reflection coefficients of AMC and ground plane

(a) Simulation in Microstripes v6.5™

(b) Simulation in Designer™ ver 2.0
<table>
<thead>
<tr>
<th></th>
<th>5GHz</th>
<th>5.83GHz</th>
<th>5.93GHz</th>
<th>7GHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMC</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>PEC</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

$E(V/m)$

-40dB 0dB

Figure 2.8 Field distributions of AMC and PEC
2.3.2 Electromagnetic Bandgap (EBG)

Electromagnetic Bandgap structures have been studied over the last few years for microwave applications. The accurate way to analyse the bandgaps of periodic structures is to produce the dispersion diagrams for the unit cell in a full-wave simulation. An efficient and automated way to do this is in Ansoft HFSST™ using the Eigenmode solver and Optimetrics modules.

A. Dispersion diagrams of periodic structures

The speed of light in a medium is the velocity at which a plane wave would propagate in medium, while the phase velocity is the speed at which a constant phase point travels. For a TEM plane wave, these two velocities are identical, but for other types of guided wave propagation the phase velocity may be greater or less than the speed of light. If the phase velocity and the attenuation of a guide are constants, then the phase of the wave that contains more than one frequency component will not be distorted. If the phase velocity is different for different frequencies, then the individual frequency components will not be able to keep their original phase relationships as they propagate along the guide and a wave distortion will occur. The phenomena is called dispersion, since different phase velocities allow the 'faster' waves to lead in phase relative to the 'slower' waves, and the original phase relationships will gradually be dispersed as the wave propagates along the guide [12].

It is useful to plot the propagation constant $\beta$ versus the free space propagation constant $k$ (or $\omega$), in order to clearly define the passband and stopband characteristics of a periodic structure. Such a graph is called a $\kappa - \beta$ diagram, or Brillouin diagram (after L. Brillouin, a physicist who studied wave propagation in periodic crystal structures). For instance, consider the dispersion relation for a waveguide mode:

$$\beta = \sqrt{\kappa^2 - \kappa_e^2},$$

where $\kappa_e$ is the cut-off wave number of the mode, $\kappa$ is the free space wave number, and $\beta$ is the propagation constant of the mode. The relation is plotted in a $\kappa - \beta$ diagram of Figure 2.9 [12]. For values of $\kappa < \kappa_e$, there is no real solution for $\beta$, so the mode cannot propagate. For $\kappa > \kappa_e$, the mode can propagate, and $\kappa$ approaches $\beta$ for large values of $\beta$ (TEM propagation).
The $\kappa - \beta$ diagram is very useful in interpreting the various wave velocities associated with a dispersion structure. The phase velocity is $V_p = \frac{\omega}{\beta} = c \frac{\kappa}{\beta}$, which is seen to be equal to $c$ (speed of light) times the slope of the line from the origin to the operating point on the $\kappa - \beta$ diagram. The group velocity is $V_g = \frac{d\omega}{d\beta} = c \frac{d\kappa}{d\beta}$, which is the slope of the $\kappa - \beta$ curve at the operating point. Thus, from the figure, it can be observed that the phase velocity for a propagating waveguide mode is infinite at cut-off and approaches $c$ (from above) as $\kappa$ increases. The group velocity, however, is zero at cut-off and approaches $c$ (from below) as $\kappa$ increases. Actually a $\kappa - \beta$ diagram can be used to study the dispersion characteristics of many types of microwave components and transmission lines.

B. Mushroom-type EBG surfaces

Mushroom-type EBG structures have been investigated extensively over the last few years. Figure 2.10 shows the dispersion diagram of a mushroom-type EBG structure [13]. In the left part of the light line, it is the leaky-wave area where the phase velocity is much higher than the speed of light and the group velocity is less than the speed of light. In the right part of the light line, it is the bounded wave area. The first mode is a TM mode and the second mode is a TE mode. The grey area is a bandgap of
the mushtoom-type EBG surface. We must pay more attention to the leaky-wave part in the dispersion diagram when we use a mushroom-type EBG surface. When a mushroom-type EBG surface is integrated into an antenna, the leaky-wave would affect the radiation pattern of the antenna. The beam angle formed by the leaky wave can be obtained in an approximate equation \( \cos \theta = \frac{\lambda}{L} \) as introduced in section 1.2 [3]. Application of the mushroom-type EBG surfaces on HPRC antennas will be discussed in section 4.2.

![Dispersion diagram of the mushroom-type EBG structure](image)

Figure 2.10 Dispersion diagram of the mushroom-type EBG structure by Dan Sievenpiper (from Ref [13])

**C. Single layer EBG surfaces (FSS)**

It is known that a single layer conducting-type FSS shows TE bandgaps along its surface [14]. Later it is also found that an aperture-type FSS shows TM bandgap along its surface. Therefore the property of aperture-type FSS surfaces might be used to suppress the TEM mode in the resonant cavity of HPRC antennas. The existence of the TEM mode in the resonant cavity may cause higher sidelobes in H-plane for a HPRC antenna in a relatively finite structure. The phenomena of higher sidelobes in H-plane will be discussed in section 4.3.1.
According to the theory of parallel plate waveguides [12], we know that the propagation constant of the TEM mode in the resonant cavity is \( k_0 \) (the wave number of free space) and the propagation constant of TM modes \( \beta = k_0 \cos \theta \) (the value is very small because \( \theta \) is close to 90 degrees for an antenna with high-directivity at broadside) in the resonant cavity. Meanwhile according to the just discussed theory of dispersion, along the propagation direction of the TEM mode, a dispersion phenomena for TM modes in the resonant cavity will not appear since the propagation constant \( \beta \) is very small and the dimension \( d \) of planes is several wavelengths (so \( \beta \times d \) is very small). Thus a TM bandgap along the propagation direction of the TEM mode will only affect the TEM mode in the resonant cavity. The TM bandgap hardly affects the TM modes in the resonant cavity.

Therefore at the antenna’s operating frequency, a novel aperture-type PRS plane with a high reflection coefficient in the normal direction and a TM bandgap along its surface can be utilized in the configuration of HPRC antennas. The HPRC antenna in a relatively finite structure is expected to obtain low sidelobes in both planes. Detailed discussions will be presented in section 4.3.

### 2.4 Simulation Methods

Although a theoretical analysis is very important, a rigorous full-wave computation for the electromagnetic analysis and design is needed in order to obtain an accurate result. As for numerical models, every modelling activity may include more or less the followings steps [15][16]:

- Conceptualisations
- Formulation
- Numerical implementation
- Computation
- Validation

In the thesis, three commercial simulation software packages (Flometrics Mircostripes™ ver6.5, Ansoft HFSS™ ver9.2 and Ansoft Designer™ ver2.0) will be utilized to run all simulations and they follow similar steps:
- Definitions of geometry models
- Specifications of materials and boundaries
- Meshing
- Setup of output variables
- Running of simulations
- Output of results

2.4.1 Microstripes™ ver6.5

In the thesis Microstripes™ (Transmission Line Modelling Method) is mainly used to run the numerical simulations for the antennas in its entirety because of its friendly interface and fast computation capabilities. The 3-D electromagnetic simulation software package is distributed by Flomerics Electromagnetic Division Ltd. [17]. The recent version 6.5 runs on a PC and offers many simulation electromagnetic features. There are many good visualisation tools, such as build models, surface current, near to far field and 3-D radiation pattern.

Time domain-based transmission line modelling is not a new idea. A lossless transmission line contains a distributed inductance and capacitance. TLM method was first presented by Johns and Beurle in 1971 [18]. Johns and Beurle used the earlier proposed principles of equivalent transmission line circuits and their application to the solution of electromagnetic problems, along with the theory of pulse propagation on transmission line to develop a numerical technique compatible with computational methods for the solution of two-dimensional electromagnetic problems. This method has been used for many applications and it has been further developed [17].

The design process follows the several steps as the introduction of this section:
- Define the geometry of the objects by using the versatile and intuitive modeller based on the ACIS kernel or import the model from another CAD package
- Specify the frequency range and pulse time
- Assign the material properties to the each body of the geometry
• Define the boundary of computation. There are four kinds of boundaries in Microstripes™: Electrical wall, Magnetic wall, Absorbing and Wrap boundary. Warp boundary is used for those simulations of periodic structures.

• Choose the type of excitation (e.g. plane wave or coaxial line)

• Define the results to be obtained (radiation pattern, fields, surface current, loss, etc.)

• Check and mesh the computation manually after the automation meshing. The meshing is very important for simulations. The minimum value of the meshing should be small enough to ensure that a mesh doesn’t contain more than one material. To get a more accurate result, the gap between conducting components should have several meshing cells even the gap is very small. The character ‘Lumped cell’ in Microstripes™ can be utilized to save simulation time and memory.

• Generate the TLM file and let the efficient time domain solver based on the TLM (Transmission Line Matrix)

2.4.2 Ansoft HFSS™ ver9.2

HFSS™ ver9.2, which is distributed by Ansoft [19], employs frequency domain-based Finite Element Method (FEM) to generate an electromagnetic field solution. FEM generally divides the full problem space into thousands of smaller regions and represents the field in each sub-region (element) with a local function. In HFSS™, a geometric model is automatically divided into a large number of tetrahedra, where a single tetrahedron is a four-sided pyramid. This collection of tetrahedra is referred to as the finite element mesh.

The software includes the following functionalities that are related with the research project:

• Basic electromagnetic field quantities and for open boundary problems.

• Characteristic port impedances and propagation constants.

• Generalized S-parameters and S-parameters renormalized to specific port impedances.
The eigenmodes or resonances of a structure. An eigenmode solution solver need not specify sources for the problem. The calculation of resonance for the model is based on the geometry, materials and boundaries.

To run eigenmode solver solutions for unit cells in obtaining their dispersion diagrams, there are several important parameters that need to be defined: PML, Master/Slave Boundary, Adaptive pass and Minimum frequency

**A. PML**

A Perfectly Matched Layer (PML) boundary is used to simulate materials that absorb outgoing waves. Setting up a PML boundary is similar to setting up a radiation boundary: draw a virtual object around the radiating structure; however, instead of placing a radiation boundary on its surfaces, add fictitious PMLs to fully absorb the electromagnetic field impinging upon them. These materials are complex anisotropic.

There are two types of PML applications: free space termination and reflection-free termination. Free space termination is utilized to analyze EBG structures in the thesis. The surface that PMLs are associated with radiates into free space equally in every direction. PMLs are more appropriate than radiation boundaries in this case because PMLs enable radiation surfaces to be located closer to radiating objects, reducing the problem domain.

In the thesis we can create PML automatically to analyze EBG structures since the base object touching the PML is planar and its material is homogenous. The following guidelines need be taken care when assigning PML boundaries:

- When automatically creating PMLs, HFSS™ creates a new relative coordinate system for each PML object. This results in the z direction of the PML object coinciding with the normal direction of the base object’s face.
- HFSS™ treats PMLs uniformly with regard to thickness. If the PMLs in our design vary in thickness, create a separate PML group for each thickness.

**B. Master/Slave Boundary**

The paired Master and Slave boundaries enable us to model planes of periodicity where the E-field on one surface matches the E-field on another to within a phase difference. They force the E-field at each point on the slave boundary match the E-
field to within a phase difference at each corresponding point on the master boundary. This is useful for simulating devices such as infinite arrays. We must pay attention to the angle when we define the coordination system - the angle between the axes defined by the $u$ point and $v$ point must be identical for the master and slave boundary.

C. Minimum Frequency and Adaptive Pass

Subsequently it is the parameter definition of analysis setup, such as minimum frequency that is related to the initial meshing. To produce the optimal mesh, HFSS™ uses an iterative process, called an adaptive analysis, in which the mesh is automatically refined in critical regions. First, it generates a solution based on a coarse initial mesh. Then it refines the mesh in areas of high error density and generates a new solution. When selected parameters converge to within a desired limit, HFSS™ breaks out of the loop. If an adaptive solution is required, then it needs define the parameter: the maximum number of passes. An adaptive analysis is a solution process in which the mesh is refined iteratively in regions where the error is high. Thus it can increase the solution’s precision. A criteria is set up to control mesh refinement during an adaptive field solution. Many problems can be solved by using only adaptive refinement.

Even though a computation is not converged, HFSS™ still stops the computation and gives out a result when the number of the adaptive pass is met. To characterize the truth of the result, the quality factor (Q) of the mode can be used. As we know, the quality factor (Q) of a passive resonant cavity would be more than 1. So if the quality factor (Q) of the mode is less than 1, we must repeat the computation by setting a new minimum frequency that is generally larger than the previous one.

There is a trade-off among the size of the mesh, the desired level of accuracy, and the amount of available computing resources. The accuracy of the solution depends on the size of the each individual element (tetrahedra). Generally speaking, solutions based on meshes using thousands of elements are more accurate than solutions based on coarse meshes using relatively few elements. To generate a precise description of a field quantity, each element must occupy a region that is small enough for the field to be adequately interpolated from the nodal values. However, generating a field
solution involves inverting a matrix with approximately as many elements as there are tetrahedra nodes. For meshes with a large number of elements, such an inversion requires a significant amount of computing power and memory. Therefore, it is desirable to use a mesh fine enough to obtain an accurate field solution but not so fine that it overwhelms the available computer memory and processing power.

2.4.3 Ansoft Designer™ ver2.0

Ansoft Designer™ ver2.0 [20] uses the Method of Moments (MoM) to solve the Mixed-Potential Integral Equation (MPIE) and calculate surface current $J$ everywhere on the mesh to generate a solution from which S-parameters can be computed. The MoM can be broken down into two sections: the basis functions and the testing functions.

- **Basis Function**
  Ansoft Designer uses a zero-order normal element basis function to interpolate the interior current values from values on the edges. Zero-order normal elements have one unknown for each edge in the mesh.

- **Testing Function**
  Testing functions are applied to the MPIE to obtain a matrix equation. The matrix equation is used to find $J$. From $J$, Designer™ can calculate the S-parameters and the radiated fields. The functions used for the basis functions are also used for the testing functions.

In Designer™, the surface of the geometric model is automatically divided into triangles and rectangles. This collection of triangles and rectangles is referred to as the mesh. A volume mesh is not generated inside the model because at high frequencies the skin depth is so small that a surface mesh is sufficient.

Designer™ is mainly used to simulate the periodic structures in this research project, especially for parametric study of AMC, since it provides more flexible solutions while compared with Microstripes™ and HFSS™. A significant feature is to support oblique incidence.
The aperture part of structures is usually represented by magnetic surface in Designer™. Our simulation results show that that is not very accurate, especially at high frequency. We must pay attention to that when simulating aperture-type structures (e.g. aperture-type FSS).
2.5 Conclusions

The ray analysis of the beam forming function of the resonant cavity antennas predicts the possibility of low-profile (up to subwavelength) HPRC antennas and tilted beams. The analysis of periodic aperture arrays and grating lobes demonstrates that the periodicity of the PRS could be optimised to obtain higher efficiency HPRC antennas.

Two kinds of metamaterial planes are investigated: AMC (including metamaterial surface with negative reflection phases) and EBG. Field distributions at normal direction to AMC/PEC surfaces are plotted out. It is observed that there is the strongest field at the conducting layer of AMC surfaces. A novel aperture-type EBG surface with a high reflection coefficient in the normal direction and a TM bandgap along its surface can be obtained to suppress the TEM mode in the resonant cavity.

Three commercial electromagnetic simulation packages are used in the thesis to finish all designs and analysis: Flometrics Microstripes™ ver6.5, Ansoft HFSS™ ver9.2 and Designer™ ver2.0. Based on their special characters, they will be employed at the different simulations. Microstripes™ ver6.5 is mainly used to simulate HPRC antennas in its entirety. Ansoft HFSS™ ver9.2 is used to get the stopband or passband of periodic structures by obtaining their dispersion diagrams of unit cells. Ansoft Designer™ ver2.0 is particular to calculate the complex reflection coefficients of the periodic structures.
References


[17] Flomerics Electromagnetic Division, Microstripes™ ver 6.5, 2004


Chapter 3 Low-profile HPRC Antennas based on Metamaterial Ground Planes

3.1 Introduction

In this chapter HPRC antennas based on metamaterial ground planes will be investigated. Section 2.2.1 has demonstrated the feasibility of low-profile HPRC antennas. Different AMC surfaces including metamaterials with negative reflection phases are designed and utilized in the configuration of low-profile HPRC antennas [1][2].

In section 3.2 a parametric study on AMC surfaces is carried out. An AMC surface is then fabricated and tested experimentally. Subsequently the AMC surface is employed as the ground plane of HPRC antennas to obtain a quarter-wavelength resonant cavity [3]. A low-profile HPRC antenna fed by a rectangular patch is fabricated and measured in an anechoic chambers to validate the analysis.
In section 3.3 a metamaterial ground plane with negative reflection phases is utilized in the configuration of HPRC antennas to further reduce the profile of the resonant cavity. The reduction of the antenna profile leads to a lower directivity. The directivity is determined by the leak-mode loss of the resonant cavity [4]. Ray analysis is employed to estimate the leaky-mode loss of a subwavelength resonant cavity. PRS planes with higher reflection coefficients are employed to achieve higher directivities. To exempt the effect of removing the central conducting patches to accommodate the rectangular patch antenna as done in section 3.2, a dipole is used as the excitation of subwavelength HPRC antennas.

3.2 Low-profile HPRC Antennas with AMC Ground Planes

3.2.1 AMC design and measurement

Until present several structures showing artificial magnetic conductance have been investigated, such as mushroom-type EBG, UC-PBG and split-rings [5-7]. Actually any planar periodic passive structures, also known as FSS, can be used in order to realise an AMC surface when printed over a ground plane. In general, the AMC bandwidth is defined as the bandwidth of $-90^\circ$ to $90^\circ$ reflection phase values. A square patch is selected as the geometry of unit cell because it yields a broadband response when compared with other elements (e.g. square loops). The structure is shown in Figure 3.1. The major contribution to the overall capacitance in this resonant cavity comes from the gap discontinuity at both ends of the patch. The patch itself with the current flow produces an inductance. A detailed discussion on the transmission-line model and the related calculation of the parameters may refer to [8].

Initially a parametric study of the planar AMC structure is carried out for the normal incidence of plane waves. The initial dimension of square patches is 4.1mm×4.1mm, the periodicity is 4.5mm, the thickness of the dielectric substrate is 1.15mm and the permittivity is 2.2. The varying parameters are the dimension of square patches (Figure 3.2), the gap width of square patches (Figure 3.3), the thickness of dielectric substrates (Figure 3.4) and the permittivity of dielectric substrates (Figure 3.5). When
one parameter is investigated, the rest parameters are not changed. Simulations are carried out in Ansoft Designer™ ver2.0.

(a) planar AMC structure
(b) unit cell of AMC

Figure 3.1 Geometry model of planar AMC surfaces

Figure 3.2 Reflection phase of AMC for different size of square patches (normal incidence)
Figure 3.3  Reflection phase of AMC for different gaps between square patches (normal incidence)

Figure 3.4  Reflection phase of AMC for different thickness of dielectric substrates (normal incidence)
Simulations of the mushroom structure and published results have shown that a similar response is obtained in the presence of vias for the varying parameters: dimensions of square patches, gap width of square patches, thickness of dielectric substrates and permittivity of dielectric substrates [9].

The reflection phase of AMC with the incidence angle of plane waves is also investigated and the result is shown in Figure 3.6. It is observed that the frequency with zero reflection phases goes up when the incidence angle increases. This is unlike the mushroom-type structure whose frequency with zero reflection phases goes down with the incidence angle increasing [10].
Figure 3.6 Reflection phase of AMC with the incidence angles

Figure 3.7 Reflection phase of AMC with the diameter of via
To better understand the function of AMC surfaces and access the effect of via when present, a parametric study on the diameter of via for mushroom-type EBG structures is also carried out in Ansoft Design™ ver2.0 and the result is shown in Figure 3.7. The via is cylindrical. Its dimensions are 0mm x 0mm x 1.15mm (no via), 0.1mm x 0.1mm x 1.15mm, 0.5mm x 0.5mm x 1.15mm, and 1mm x 1mm x 1.15mm respectively. The rest parameters are kept as the previous ones. Figure 3.7 proves that the dimension of a thin via slightly affects the reflection phase of AMC for normal incidence since different diameters will yield different values of the inductance [11]. This result will be useful for the application of the mushroom-type EBG structure in Chapter 4.

An AMC surface without vias (as shown in Figure 3.1) has been fabricated and measured in an anechoic chamber. The permittivity of the dielectric substrate is 2.2 and the thickness is 1.15mm. The size of square patches is 4.1mm x 4.1mm. The periodicity is 4.5mm. The ground plane and the patch array are made of copper. The setup is shown in Figure 3.8: two horn antennas, illuminating a flat surface, and measure their reflection phases.

The measurement steps are as following:

- Put a PEC ground plane on the wall and then measure the complex reflection coefficient of the PEC ground plane
- Replace the PEC ground plane with the AMC plane and measure the complex reflection coefficient of the AMC plane
- Calculate the reflection phase of the AMC plane by comparing the above measured results

The measured result is shown in Figure 3.9. There is a good agreement between simulation and measurement. The simulation result in the figure is for a normally incident plane wave and obtained in Microstripes™ ver6.5. The measured range between -90° ~ +90° is from 12.3GHz to 17GHz with the zero reflection phase frequency at 14.3GHz and a bandwidth is about 33%.
Figure 3.8  Environment setup for AMC measurement

Figure 3.9  Reflection phase measurement of the AMC surface
3.2.2 Rectangular patch antennas as excitation

A waveguide slot antenna has been used as the excitation of HPRC antennas in [12-14]. It is known that there are some problems associated with the usage of the waveguide, such as its weight, profile and cost. A good alternative for the waveguide is microstrip antennas that have found wide application [15][16].

Microstrip antennas can be designed into many geometrical shapes. All of them can be divided into four basic categories: microstrip patch antennas, microstrip dipoles, printed slot antennas, and microstrip travelling-wave antennas. After some comparison, a rectangular patch antenna is selected as the excitation of low-profile HPRC antennas because it provides more parameters to match the input impedance.

A microstrip rectangular patch antenna comprises a conducting rectangular patch on one side of a dielectric substrate with a ground plane on the other side as shown in Figure 3.10. A coax-line is typically used as the feeding of the patch antenna due to its simplicity.

![Figure 3.10 Models of a rectangular patch antenna](image)

The dimension of the rectangular patch can be approximated calculated based on the following formulas [16]

The width \( W = \frac{\varepsilon_r}{2f_0 \sqrt{\varepsilon_{r0} + \varepsilon^2}} \), the length \( L = \frac{\varepsilon_r}{2f_0 \sqrt{\varepsilon_{r0}}} - 2\Delta l \) and the fringing length \( \Delta l = h \xi_1 \xi_2 \xi_3 \xi_4 \),

where \( \xi_1 = 0.434907 \frac{2.881 + 0.25 (W/h)^{0.8544 + 0.226}}{2.881 - 0.189 (W/h)^{0.8544 + 0.87}} \),

\[ \xi_2 = 1 + \frac{(W/h)^{2.71}}{2.3586 + 1} \]
\[ \xi_3 = 1 + \frac{0.5274 \text{arctan}[0.084(W/h)^{1.9413}]}{\varepsilon_r^{0.9256}} \]

\[ \xi_4 = 1 + 0.0377 \text{arctan}[0.067(W/h)^{1.456}] \{6 - 5 \exp[0.036(1 - \varepsilon_r)]\} \]

\[ \xi_5 = 1 - 0.218 \exp(-7.5W/h) \]

\[ \varepsilon_r = \frac{(\varepsilon_r + 1)}{2} + \frac{(\varepsilon_r - 1)}{2} \left[ 1 + \frac{10h}{\lambda} \right]^{-\frac{1}{2}} \]

where \( \varepsilon_r \) is the relative permittivity of the dielectric substrate and \( h \) is the thickness of the dielectric substrate.

The offset from the centre is \( 0.5 \frac{L}{\sqrt{\varepsilon_r}} \) when using a coax-line as a feeder. Assuming the antenna’s operating frequency is 14GHz and the input impedance is 50Ω, the related parameters can be calculated according to the above formulas: \( L = 6.34\text{mm}; W = 6.24\text{mm}; Offset = 2.77\text{mm} \). The thickness of the dielectric substrate is 1.15mm and the permittivity is 2.2.

Subsequently the rectangular patch is used as the excitation of a HPRC antenna. The configuration of HPRC antennas is similar to the one introduced by Figure 1.2. The dimension of the ground plane is 150mm × 150mm. The thickness of the PRS dielectric substrate is 1.55mm with the permittivity \( \varepsilon_r = 2.55 \). The PRS contains 9×9 square patch arrays. The square patch element is 10mm×10mm and the unit cell periodicity is 11mm. The conducting layer is copper. The distance between the PRS and the ground plane is 11.2mm. The rectangular patch antenna with a ground plane 40mm×40mm (no PRS) is also simulated to do a comparison. The simulations are carried out in Microstripes™ ver6.5.

Figure 3.11 shows the simulated return losses. It is observed that both of the antennas can be matched after optimisation. Figure 3.12 shows that a significant directivity improvement about 14.5 dB is achieved by the planar resonant cavity. The radiation pattern at the maximum directivity is shown in Figure 3.13 and the SLL is less than -15dB. The cross polarization level is less than -35dB in both planes. So a rectangular patch can be used as the excitation of HPRC antennas.
Figure 3.11  Return losses of the rectangular patch antenna and the related HPRC antenna

Figure 3.12  Directivities of the rectangular patch antenna and the related HPRC antenna
Figure 3.13 Radiation patterns of the rectangular patch antenna and the related HPRC antenna at 14GHz
3.2.3 Antenna performance

Section 3.2.2 demonstrates the feasibility of using a rectangular patch as the excitation of HPRC antennas. When the author was working on the antenna, the version of Flomerics Microstripes™ was 6.0 (the current version is 6.5). By using the version 6.0, the author designed a 6mm×4.5mm rectangular patch antenna and the simulated working frequency is 14GHz. After fabrication, its practical dimension was about 6.2mm×4.6mm. The soldering of the connection between the inner conductor of the coax-line and the rectangular patch generates a small lump in the feeding point. The lump will lengthen and widen the equivalent dimension of the rectangular patch. That will change the surface current distribution of the rectangular patch and shift the resonant frequency of the rectangular patch to a slightly lower one. When the rectangular patch with an AMC surface is integrated into the low-profile HPRC antenna, the measured operating frequency of the low-profile HPRC antenna is around 14GHz which will be shown in section 3.2.4. So it is expected that a slightly larger rectangular patch is required to match the return loss at 14GHz.

After some investigation it is found that a 6.3mm × 4.7mm rectangular patch shows an operating frequency at 14GHz in Microstripes™ ver6.5. The simulated result is shown in Figure 3.14. The feeding point is 0.5mm away from the centre of the rectangular patch. The radius of the feeding via is 0.1mm. The dimension of the ground plane is 50mm×50mm. The thickness of the dielectric substrate is 1.15mm and the permittivity is 2.2.

Subsequently the highly reflective PRS from [14] has been designed using a square patch array printed on a similar dielectric substrate as the AMC surface (without the ground plane). The dimension of square patches is \( L_{prs} = 10 \text{ mm} \) and the periodicity \( D_{prs} = 11 \text{ mm} \). The thickness of the dielectric substrate is 1.5mm and the permittivity is 2.55. High reflectivity value about 0.955 is obtained for a broad range of frequencies on either side of 14 GHz [14].
The microstrip patch antenna (6.2 mm × 4.6 mm) has been designed together with an AMC ground plane. The patches of the AMC surface are surrounding the antenna that is printed on the same dielectric substrate. The square patch of the AMC surface is 4.1mm × 4.1mm and the periodicity is 4.5mm. The thickness of the dielectric substrate is 1.15mm and the permittivity is 2.2. In order to accommodate the rectangular patch antenna, the central four elements are removed. The rectangular patch is fed by a via that is 0.8mm far away from the centre of the patch in order to achieve a good matching. The radius of the via is set to 0.1mm in the simulations.

The resonant cavity antenna is formed by placing the PRS at the resonant distance over the rectangular patch antenna with the AMC ground plane. The resulting structure is similar to the schematic diagram of Figure 3.15(b). The resonant distance for an operating frequency of 14GHz is simulated at 7.1mm from the PRS to the ground plane (approximately 5.95mm from the PRS to the conducting layer of AMC surface). This corresponds to an approximately quarter wavelength resonant cavity at
14 GHz since the quarter wavelength of 14GHz is about 5.7mm. The corresponding resonant distance for the same antenna configuration but with a PEC (instead of the AMC) ground plane is about 11.3 mm. Therefore, a reduction of the antenna profile by approximately 40% has been achieved by virtue of employing the AMC ground plane.

In the case of a rectangular patch antenna as the excitation, the low-profile HPRC antennas can be matched by changing the location of the feed point as shown in Figure 3.16. The feeding offset can be changed to match the input impedance of low-profile HPRC antennas and there is not much effect on the antennas' other performances (e.g. directivities and radiation patterns) from the simulation results as shown in Figure 3.16, Figure 3.17, Figure 3.18 and Figure 3.19.

Figure 3.17 shows that the maximum directivity is slightly different for different feeding points. Figure 3.18 indicates that the backlobes of low-profile HPRC antennas fed by a rectangular patch antenna are higher than the backlobes of the PEC grounded HPRC antenna fed by a rectangular patch antenna as illustrated in Figure 3.13 (from –35dB to –25dB, about 10dB difference). Different feeding points only slightly affect the radiation patterns in E-plane.

The higher backlobes in the above antenna configuration are caused by the introduction of the AMC surface. The patch antenna excites a strong surface wave. The surface wave propagates along the conducting layer of AMC and is diffracted out when it reaches the edges of the ground layer. In the antenna configuration with a PEC ground plane, less surface waves can reach the boundary of the ground layer. So the backlobes is up to –35 to –40 dB while compared with the peak value of mainlobes.

Figure 3.19 demonstrates that there is strong field with the conducting layer of AMC. However the surface waves can be suppressed by employing mushroom-type EBG ground planes. The related investigation will be presented in section 4.2.
Figure 3.15 Models of a low-profile HPRC antenna with AMC ground planes

Figure 3.16 Return loss of low-profile HPRC antennas for different offset position
Figure 3.17 Directivities of low-profile HPRC antennas for different offset position

Figure 3.18 (a) E-plane
Figure 3.18 Radiation patterns of low-profile HPRC antennas for different offset position at 14.1GHz
Subsequently three different PRS planes containing $7 \times 7$ unit cells or $9 \times 9$ unit cells or $11 \times 11$ unit cells were tried to check the effect of PRS dimension on the performance of HPRC antenna, especially the backlobes [16]. The result will help to design higher aperture efficiency HPRC antennas.

Figure 3.20 shows that the operating frequency of return loss slightly shifts to a lower value by virtue of the application of a larger PRS. In addition, a larger PRS plane will reduce edge diffraction and allow more leakage waves. Therefore the bigger PRS plane will slightly enhance the maximum directivity as shown in Figure 3.21.

Figure 3.22 shows that at the frequency of maximum directivity the sidelobes of the low-profile HPRC antennas are reduced, but the backlobes remain at $-25\text{dB}$ to $-30\text{dB}$. This fact confirms that the backlobes are mainly caused by the edge diffraction of AMC surface. Figure 3.23 illustrates the related field distributions and agrees with
Figure 3.22 very well, a bigger PRS plane has less diffraction from the resonant cavity and lower sidelobes are observed. However, strong fields always exist with the metamaterial ground planes.

Therefore the dimension of PRS will mainly affect the radiation patterns of low-profile HPRC antennas, not the backlobes. The backlobes can be reduced by introducing mushroom-type EBG structures. The detailed discussion will be presented in section 4.2.
Figure 3.21 Directivities of low-profile HPRC antennas for different PRS planes

Figure 3.22 (a) E-plane
Figure 3.22 Radiation patterns of low-profile HPRC antennas for different PRS planes at the frequency with maximum directivity (14GHz or 14.1GHz)
3.2.4 Measurements

To validate the above simulation results, an AMC ground plane with a rectangular patch is fabricated as shown in Figure 3.24(a). The permittivity of the dielectric substrate is 2.2 and the thickness is 1.15mm. The dimension of the dielectric substrate is 150mm x 150mm. The periodicity is 4.5mm and the size of square patches is 4.1mm x 4.1mm. The number of square patches is 32 x 32. The space by removing the central four elements is used to accommodate a 6.2mm x 4.6mm rectangular patch.

The PRS is the one used in a previous measurement [14]. The dimension of the dielectric substrate is 150mm x 150mm with permittivity 2.55 and thickness 1.5mm. The PRS consists of 9 x 9 unit cells. The periodicity is 11mm with a square patch 10mm x 10mm. To suspend the PRS plane, four holes are drilled at the four corners of the planes as shown in Figure 3.24. Four plastic screws with related plastic columns are used to separate the planes. The length of the plastic columns is 5.9mm.

The measured and simulated return losses are shown in Figure 3.25. The simulation is carried out in Microstripes™ ver6.0. In the simulation, the dimension of the rectangular patch is 6.0mm x 4.5mm. Default meshing (max 1.5mm and min 0.15mm) is used the maximum frequency is 20G Hz. The measured -10dB return bandwidth
is more than 1.5GHz. The gain of the low-profile antenna is shown in Figure 3.26. A maximum gain of 19 dBi has been obtained at about 14 GHz and a good matching has been achieved. The antenna bandwidth as defined from the -3 dB gain level and 10 dB return loss is about 2%. Compared with the directivity of the HPRC antenna with a PEC ground plane as shown in Figure 3.12, the gain loss for the low-profile HPRC antennas is due to the several reasons:

- The removing of the central elements to accommodate the rectangular patch antenna. Thus the phase conditions of resonance in the area cannot be met and much loss would be generated. This is the main reason of the lower directivity.
- As discussed in the section 2.5.1, AMC surfaces don't fully reflect back the incident waves even though they are backed by fully reflective ground planes. The loss is about 0.2dB. It is not major according to the investigation result in Figure 3.34, where the low-profile HPRC antenna shows a similar maximum directivity with a PEC grounded HPRC antenna when both of them are fed by a dipole.

Figure 3.24 (a) AMC with rectangular patch
Radiation patterns of E-plane and H-plane are shown in Figure 3.27. In both planes highly directive pattern is obtained and the sidelobe level is below $-15$ dB. The cross polarisation level is below $-25$ dB in both planes.

To summarize, the theoretical prediction of a quarter wavelength resonant cavity antenna using an AMC ground plane has been presented. Measured results have demonstrated the profile reduction of planar resonant cavity antennas to approximately quarter-wavelength.
Figure 3.25  Return loss of the low-profile HPRC antenna

Figure 3.26  Gain of the low-profile HPRC antenna
Figure 3.27 Radiation patterns at the frequency (14GHz) of maximum directivity
3.3 Subwavelength HPRC Antennas based on Metamaterial Ground Planes

Section 2.2.1 theoretically predicts the possibility of a lower profile (subwavelength) resonant cavity by virtue of the application of metamaterial ground planes with negative reflection phases. Section 2.3.1 demonstrated that metamaterial ground planes with negative reflection phases could be achieved based on the design of AMC surfaces. This section will focus on subwavelength HPRC antennas based on the metamaterial ground planes. In addition, a new excitation (dipole antennas) is utilized.

3.3.1 Dipole antennas as excitation

Section 3.2 demonstrates that a rectangular patch antenna can be used as the excitation of low-profile HPRC antennas. However removing the central elements to accommodate the rectangular patch will locally disturb the cavity structure and slightly reduce the antenna’s directivity. Thus a dipole antenna is investigated as the excitation of the following low-profile HPRC antennas.

Flomerics Microstripes™ allows a much simpler definition of a ‘dipole’ while compared with its practical implementation [18]. Three points of a wire need to be defined and then its physical character specification, e.g. thickness and impedance. A real dipole antenna could be made using coaxial lines. One side of a conducting pole grows out from the inner conductor while the other side comes out from the ground layer of coaxial lines. A detailed discussion on this will be presented in section 3.3.4.

The whole HPRC antenna configuration is the same as the previous one as shown in Figure 3.15. A wire working as a dipole is defined in the configuration. To compare with the results of the HPRC antenna fed by a dipole antenna, the dipole antenna over the ground plane without PRS is also simulated. The length of a centre-fed dipole is 10mm and the radius is 0.25mm. The thickness of the dielectric substrate is 1.15mm with permittivity 2.2.

Figure 3.28 shows the return loss of a dipole antenna over a ground plane with and without the PRS. The directivity improvement by the ground plane is about 5 dB
when the dipole is placed 3mm far away from the ground whose dimension is 40mm×40mm. The PRS increases the directivity further. The distance between the PRS and the ground plane is 11.2mm. The substrate dimension of the ground plane and the PRS plane is 150mm×150mm. The remaining parameters are the same as the previous ones. Figure 3.29 shows that the directivity has a significant improvement of about 14.5dB by virtue of the application of PRS. Figure 3.30 shows that the SLL of the HPRC antenna at the maximum directivity is lower than -15dB. The cross polarization level is less than -30dB in both planes. So a dipole antenna can be used as the excitation of HPRC antennas.

Figure 3.28 Returns loss of a dipole antenna, over a ground plane and the HPRC antenna fed by the dipole
Figure 3.29  Directivities of a dipole antenna, over a ground plane and the HPRC antenna fed by the dipole

Figure 3.30  (a) E-plane
3.3.2 Metamaterial ground plane design and application

The phase conditions of resonance in section 2.2.1 predicts the possibility of subwavelength resonant cavity when a metamaterial ground plane (MGP) with negative reflection phase is employed as illustrated in Figure 3.31. Section 2.3.1 demonstrates that the MGP at a specific frequency can be obtained starting from an AMC response at that frequency and then increasing the square patch size of the AMC structure. A detailed discussion on subwavelength HPRC antennas based on MGP will be presented in the followings.

Simulations of the metamaterial ground planes, the PRS surface and the planar antenna in its entirety were carried out in Microstripes™ ver6.5. Closely packed square patches (Figure 3.32(b)) are utilized for the metamaterial ground plane because they yield a broad bandwidth. Here the central operating frequency of the antenna is 14GHz. Metamaterial ground planes with an array periodicity $D_{mgp} = 4.5$ mm and...
different square patch length $L_{mgp}$ have been used in the antenna and their reflection responses are shown in Figure 3.32(a).

A square patch PRS array has been designed according to [14] in order to produce an antenna with high-directivity and sufficient bandwidth. The PRS contains $9 \times 9$ square patches with each patch $10\text{mm} \times 10\text{mm}$ and periodicity $11\text{mm}$. The dielectric permittivity $\varepsilon_r$ is 2.55 and thickness is $1.5\text{mm}$. Its complex reflection coefficient is shown in Figure 3.33. The planar antenna is formed as shown in Figure 3.31(b). To present the principle, a centre-fed $10\text{mm}$-long wire dipole is used in Microstripes™ ver6.5 as the excitation of the resonant cavity. The radius of the wire is $0.25\text{mm}$ and the wire is $3\text{mm}$ far away from the PEC ground plane. The dimension of the dielectric substrates is $150 \times 150\text{mm}^2$. The permittivity $\varepsilon_r$ is 2.2 and the thickness is $1.15\text{mm}$.

The directivities of the resonant cavity antennas as illustrated in Figure 3.34 are obtained from the 3D simulations for the corresponding metamaterial ground planes. By changing the distance $h$ between the PRS and the ground copper plane, we achieved the maximum peak directivity at $14\text{GHz}$ for those HPRC antennas. The distance $h$ is $11.2\text{mm}$ ($L_{mgp}=0\text{mm}$), $7.1\text{mm}$ ($L_{mgp}=4.1\text{mm}$), $5.7\text{mm}$ ($L_{mgp}=4.2\text{mm}$) and $4.5\text{mm}$ ($L_{mgp}=4.3\text{mm}$) respectively.
Figure 3.32 Reflection phases of MGP with the size of square patch

Figure 3.33 (a) Reflection coefficient
Figure 3.33 Reflection response of PRS for different dimensions of square patches

Figure 3.34 Simulated directivities of HPRC antennas with MGP
The simulation results in Figure 3.34 show that when the metamaterial ground planes' reflection phases become negative at 14GHz the whole profile of HPRC antennas is further reduced at the expense of the maximum directivity and the bandwidth. The bandwidth reduction is caused by the reflection phase curves of the metamaterial ground planes. If the negative reflection phase of a metamaterial ground plane is constant, the antenna bandwidth would become broader when the distance is reduced, according to the theory result in [14]. A wideband HPRC antenna would be achieved if the reflection phase of a metamaterial ground plane meets the gradient \( \frac{4\pi}{c} \frac{1}{\cos a} \) [14].

A simple explanation on the directivity degradation can be attributed to the strong coupling between the PRS array, the dipole and the ground plane, which occurs due to their close proximity. To understand that, we can plot the field distribution and PRS surface currents in Figure 3.35. With the reduction of the resonant distance, the surface current becomes weaker as the wave moves toward the edges. Energy from the excitation cannot propagate away from the central source point. Thus the radiation aperture becomes smaller and the directivity is reduced [4].

In the Figure 3.36, the radiation patterns at 14GHz with the maximum directivity are presented. They agree with the field distribution results of Figure 3.35 very well. As for \( L_{mgp}=0 \)mm MGP, the backlobes of the HPRC antenna is the lowest. While the \( L_{mgp}=4.3 \)mm's lower directivity causes a lower ratio between the mainlobe and the sidelobes. The HPRC antenna with \( L_{mgp}=4.2 \)mm MGP achieves the lowest sidelobes given in the dimension of the planes. This shows that an optimised dimension PRS exists for a specialized HPRC antenna. A full wave rigorous numerical method for the whole HPRC antenna configure would be accurate in obtaining the dimension of PRS.
Figure 3.35  Field distribution of HPRC antennas with MGP at 14GHz
Figure 3.36 Radiation patterns at the frequency (14GHz) of maximum peak directivity
3.3.3 Directivity analysis

To understand the reduction of directivity in the above antenna configurations, the best way is to find the field distribution in the resonant cavity and then derive the far-field radiation based on (Discrete) Fourier Transform. According to the wave propagation in the parallel-plate waveguide, the field distribution in the resonant cavity can be expressed in an approximate form $e^{-\alpha z + j\beta z}$ [19], where $\alpha$ is the loss, $\beta$ is the propagation constant and $z$ is the distance of wave propagation. The loss $\alpha$ is mainly introduced by the leaky-wave loss. The most efficient way to compute the leaky-wave loss is based on full wave rigorous numeral analysis of the unit cell in periodic structures. However in the above resonant cavity antennas the periodicity of PRS and the metamaterial ground planes are non-commensurate (they are not multiple integers of each other). This case may be very common since the two planes are designed independently. Thus, a numerical analysis of the unit cell is not possible.

![Wave propagation in the resonant cavity antenna](image)

Here, by further extending the ray analysis, the leakymode loss $\alpha$ can be calculated in a simple way based on the previous parallel plate waveguide model. The diagram in Figure 3.37 shows the ray propagation and wave front direction of a uniform plane-wave component of TM or TE waves in the resonant cavity and it is assumed that the phase conditions of resonance as discussed in section 2.2.1 are met.

The wave loss in the waveguide can be expressed as $e^{-\alpha z}$ after travelling a length $z$ in field expression. If the reflection coefficient of the upper layer plane is $p$ and the wave propagates along a distance $Z$ in the resonant cavity, the wave is reflected
\[
\frac{z}{2d_z} (d_z = h*\tan \theta) \text{ times by the upper layer PRS. Thus the leaky-wave loss from the ray analysis would be expressed by } (p)^{\frac{z}{2d_z}}, \text{ actually this also expresses the energy that remains in the cavity the same as } e^{-az}. \text{ The above expressions of the loss should be equal since they only indicate the wave loss in different ways:}
\]

\[
e^{-az} = (p)^{\frac{z}{2h*\tan \theta}} \quad (19)
\]

\[
\Rightarrow -az = \frac{z}{2h*\tan \theta} \cdot \ln p \quad (20)
\]

\[
\Rightarrow \alpha = -\frac{\ln p}{2h*\tan \theta} \quad (21)
\]

It is also found that there is no difference in essence to use either kind of expressions (wave field \(e^{-az}\) or wave power \(e^{-2az}\)) in deriving the loss. Both of them would be able to obtain the same result as shown in equation (21).

In the case of the above HPRC antenna, the upper layer PRS plane is same. Thus the reflection coefficient \(p\) is constant. When compared with the change of the distance \(h\) between the PRS and the ground plane in the resonant cavity, \(\theta\) has not much change and is always very small since all of those antennas are based on the resonant cavity with a same dipole as the excitation. Combined with the results in section 3.3.1 according to the equation (8) in section 2.2.1, the leaky mode loss can be calculated respectively and shown in Table 1. The angle \(\theta\) (this is the angle \(\alpha\) in section 2.2.1) can be calculated from the ray propagation distance \(L\) (this is not the distance between the planes in section 2.2.1) of ray analysis and the resonant distance \(h\) (this is the distance \(L\) between the planes in section 2.2.1) in the above simulations of HPRC antennas. Both of the distances should be a little different or the leaky mode loss would be infinite if the angle equals to zero.
<table>
<thead>
<tr>
<th>MGP</th>
<th>$h$</th>
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<th>$I$</th>
<th>$\theta$</th>
<th>$a$</th>
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<td>180</td>
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<td>0.114665</td>
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<tr>
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<td>5.749453</td>
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<td>0.024166213</td>
</tr>
<tr>
<td>$a=4.3 \text{mm}$</td>
<td>4.5</td>
<td>-162.229</td>
<td>-45.74017</td>
<td>4.5247348</td>
<td>0.105</td>
<td>0.038488066</td>
</tr>
</tbody>
</table>

Table 1  Leakymode loss for the resonant cavity at 14GHz

It can be observed that the leakymode loss of the resonant cavity increases by virtue of the application of the metamaterial ground plane with negative reflection phases. This agrees well with the field distribution in Figure 3.35. The figure demonstrates that the field rapidly becomes weak. According to the values of leaky mode loss in Table 1, the field strength changing with the propagation distance is shown in Figure 3.38 and the calculated 2-D field distribution of PRS is plotted out in Figure 3.39.

![Field Strength](image)

Figure 3.38  Field strength of HPRC antennas with the leakymode loses at 14GHz
Therefore it is expected that the maximum directivity of HPRC antennas based on the metamaterial ground planes would increase if the leaky-mode loss is reduced by increasing the reflection coefficient of PRS planes [4]. Two new PRS planes with square patch size \( L_{prs} = 10.3 \text{mm} \) and \( L_{prs} = 10.7 \text{mm} \) are simulated. The reflection responses are shown in Figure 3.33. Then they are applied on the low-profile resonant cavity with the \( L_{mgp} = 4.3 \text{mm} \) metamaterial ground plane. Figure 3.40 shows the change of directivity. The distances \( h \) between the PRS and the ground planes becomes 4.5mm \( (L_{prs} = 10 \text{mm}) \), 4.25mm \( (L_{prs} = 10.3 \text{mm}) \) and 4.15mm \( (L_{prs} = 10.7 \text{mm}) \) respectively because of the minor change of PRS reflection phases. Simulation results demonstrate that the maximum directivity of HPRC antennas can be enhanced by introducing new PRS planes with a higher reflection coefficient.

The radiation patterns at 14 GHz with the maximum directivity are shown in Figure 3.41. The HPRC antenna with the PRS plane \( L_{prs} = 10.3 \text{mm} \) shows the lowest sidelobes. This is due to the fact that the antenna with the PRS plane \( L_{prs} = 10 \text{mm} \) shows lower maximum directivity and the antenna with the PRS plane \( L_{prs} = 10.7 \text{mm} \) shows stronger edge diffraction because of the higher reflection coefficient.
Figure 3.40  Directivity change of MGP-based HPRC antennas for different PRS planes

Figure 3.41  (a) E-plane
3.3.4 Measurements

The MGP with $L_{mgp} = 4.3\,\text{mm}$ has been fabricated using standard photolithographic techniques and substrate parameters as mentioned in section 3.3.1. The reflection phase of the MGP has been experimentally assessed as shown in Figure 3.42. Two directional horn antennas have been used to illuminate and receive the reflected field at a small angle (less than 10 degrees) from the normal to the MGP. An absorber has also been placed between the horn antennas in order to avoid proximity coupling interference. The reflection phase has been measured and normalised with respect to that obtained from a metallic plate of the same dimension in the same distance. Very good agreement between the measured and simulated results is observed. At the HPRC antenna operating frequency of 14GHz, the MGP reflects incident waves with a $-45^\circ$ phase shift.
The PRS with $L_{prs}=10\text{mm}$ has been fabricated with substrate parameters as described in section 3.3.2. The subwavelength resonant cavity is assembled by positioning the PRS over the MGP using plastic spacers. Thus four holes are drilled at the four corners of the planes. The distance between the two surfaces is exactly equal to the simulated $h=3.35\text{mm}$.

To exempt the effect of removing the central elements, a dipole has been designed and fabricated for the excitation of the resonant cavity as shown in Figure 3.43. There are two ways to put a dipole antenna into the resonant cavity. One is to put the dipole into the resonant cavity from the open side by using a long feeding coax-line. The long feeding coax-line will generate much loss and ripples with the return loss, especially when the dipole antenna is not baluned. However the use of a balun will make the dipole more complicated and further enhance the loss due to the introduction of new dielectrics in the resonant cavity. Thus it further reduces the maximum directivity. The other way is to put the dipole into the resonant cavity by going through the ground plane as shown in Figure 3.43.

The length of the right arm is about $5\text{mm}$ and the length of the left arm is also about $5\text{mm}$. Thus the total length of the dipole is about $10.7\text{mm}$ with a gap of about $0.7\text{mm}$. 

Figure 3.42 Measurement result of the MGP with negative reflection phase
The dipole is fed by a coaxial cable that penetrates through the MGP. The distance between the dipole and the conducting surface of MGP is about 1mm to 2mm. The radius of the dipole is about 0.25mm and the diameter of the coax-line is about 2.17mm. The impedance of the coax-line is 50 Ohm. A photograph of the MGP with the dipole feeding is shown in Figure 3.44(a). The PRS plane is shown in Figure 3.44(b).

![Model of a dipole antenna and the integrated HPRC antenna](image)

The gain of the prototype subwavelength HPRC antenna is measured in an anechoic chamber with a well-defined horn antenna as a receiver. Figure 3.45 shows the measured gain vs. frequency for this prototype. A maximum gain of 19.2 dBi is achieved at about 14 GHz. The bandwidth is also about 2%, as predicted in the simulations. The measured S11 is shown in Figure 3.46. The S11 is just below -10 dB at about 14GHz. The small discrepancy between the maximum simulated directivity value of 20 dB and the measured gain is attributed to the S11 value which corresponds to a loss of 0.5 dB, to the tolerances in aligning the planar surfaces, and to the small opening in the MGP that allows for the coaxial line to pass through to the dipole.

The measured radiation patterns at the frequency of maximum gain are shown in Figure 3.47. The simulated results from section 3.3.2 have also been included for comparison. Good agreement between simulations and measurements is obtained, which verifies the results of section 3.3.2. A high-directive main beam is obtained from the prototype with the SLL remaining below -15 dB in both planes. The cross polarization level is less than -25dB in both planes.
Figure 3.44 Photos of the MGP-based HPRC antenna

(a) MGP with dipole

(b) PRS
Figure 3.45 Measured gain of the MGP-based HPRC antenna

Figure 3.46 Measured return of the MGP-based HPRC antenna
Figure 3.47 Radiation patterns of the MGP-based HPRC antenna at 14GHz
3.4 Conclusions

This chapter demonstrated the possibility of low-profile HPRC antennas by utilizing metamaterial ground planes. The resonant cavity can be reduced to quarter wavelength from half wavelength by virtue of the application of AMC ground planes. A rectangular patch has been utilized as the excitation of the resonant cavity by integrating with the conducting layer of AMC. The removal of the central elements will change the phase conditions of resonance in the central area and some directivity will be lost.

Subsequently a metamaterial ground plane with negative reflection phase is utilized as the ground plane of the HPRC antenna. With the application of the MGP, the previous quarter wavelength resonant cavity can be further reduced into a subwavelength resonant cavity. A dipole antenna has been used as the excitation of the HPRC antennas. A \( \lambda/6 \) subwavelength resonant cavity fed a dipole antenna has been validated in the measurements. The use of the dipole avoids the problem of the rectangular patch as demonstrated in section 3.2. By changing the location and the length of dipole, the return loss of the subwavelength HPRC antenna can be matched. There is good agreement between the simulation and measured results.
References


[18] Flomerics Electromagnetic Division, Microstripes™ ver 6.5, 2004
Chapter 4  

HPRC Antennas based on Novel EBG Planes

4.1 Introduction

In this chapter HPRC antennas based on novel metallodielectric EBG planes will be studied. Low sidelobes/backlobes and higher efficiency are achieved by virtue of the application of novel EBG surfaces. Utilizing EBG or PBG to improve antennas' efficiency by suppressing surface waves has previously found applications, for example in patch antennas [1].

In section 4.2 low-profile HPRC antennas based on mushroom-type EBG ground planes will be discussed. Initially a parametric study on the via diameter of mushroom-type EBG structures is presented for the further application of the mushroom-type EBG structure. After employing a mushroom-type EBG ground plane with a bandgap at the antenna operating frequency, the backlobes of the low-profile HPRC is reduced by approximately 10dB when compared with the backlobes of the low-profile HPRC antenna based on an AMC ground plane.
In section 4.3 a novel EBG PRS with a TM bandgap is designed to suppress TEM modes in the resonant cavity. The suppression will help to obtain low-sidelobe in E-plane and H-plane. A low-sidelobe HPRC antenna fed by a rectangular patch is fabricated and measured. A conducting-type PRS with a TE bandgap is used to compare the reduction of SLL.

In section 4.4 a conducting-type EBG PRS with reduced periodicity clearly shows higher efficiency than a conducting-type EBG PRS with larger periodicity.

4.2 HPRC Antennas with Mushroom-type EBG Ground Planes

4.2.1 Effect of the via diameter on mushroom-type EBG

A parametric study about the effect of the via diameter on mushroom-type EBG is presented before the further application of mushroom-type EBG structures. The simulations are carried out in Ansoft HFSS\textsuperscript{TM} ver9.2 and the results are shown in Figure 4.1. The substrate is Taconic Teflon with permittivity 2.2 and thickness 1.15mm. The periodicity is 4.5mm and the metallic square patch is 4.1mm\times 4.1mm. The dimension of the central square cylinder via is 0.1mm\times 0.1mm\times 1.15mm, 0.5mm\times 0.5mm\times 1.15mm and 1mm\times 1mm\times 1.15mm respectively.

Dispersion diagrams of the periodic structures have been produced using HFSS\textsuperscript{TM} as introduced in Chapter 2. The results are shown in Figure 4.1 and they indicate that all modes (especially the first mode) shift to higher frequencies and the stop band becomes narrower when the diameter of via is increased. This can be attributed to the reduction of the equivalent inductance of a thicker via. It agrees with the equivalent $L$-$C$ circuit theory of mushroom-type EBG structure presented in [2]. The theory specifies the bandwidth $(B - \sqrt{\frac{L}{C}})$ and the working frequency $(f_o - \frac{1}{\sqrt{LC}})$ of mushroom-type EBG structures, where $L$ is the inductance of via and $C$ is the capacitance caused by the gap between the neighboured square patches.
Figure 4.1  (a)  0.1mm × 0.1mm × 1.15mm

Figure 4.1  (b)  0.5mm × 0.5mm × 1.15mm
The dispersion diagrams only show the pass/stop bands and can’t characterize which modes they are. The field plot functionality in HFSS™ can be utilized to do that. For example, Figure 4.2 shows the vector of E field for the wave propagating along the x axis. From that it can be deduced that the mode is a TM mode.
Following the procedure, the modes of above dispersion diagrams at the propagation phase $\beta d = 180$ degrees (the angle difference between the Master/Slave periodic boundaries) can be decided separately and shown in Table 2. It is observed that the first mode of the mushroom-type EBG structure is a TE mode when the diameter of the via is 0.1mm. Thus the existence of the via can affect TE modes of the mushroom-type EBG structure. The result also demonstrates that the starting frequency point of TE leaky modes does not equal to the frequency point with zero reflection phases of an AMC plane since the thin via hardly changes the reflection phase of the AMC from the simulation results in section 3.2.

<table>
<thead>
<tr>
<th>Mode</th>
<th>0.1mm</th>
<th>0.5mm</th>
<th>1mm</th>
</tr>
</thead>
<tbody>
<tr>
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<td>TM</td>
<td>TM</td>
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<tr>
<td>Mode2</td>
<td>TM</td>
<td>TE</td>
<td>TE</td>
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<tr>
<td>Mode3</td>
<td>TE</td>
<td>TE</td>
<td>TM</td>
</tr>
<tr>
<td>Mode4</td>
<td>TM</td>
<td>TM</td>
<td>TE</td>
</tr>
</tbody>
</table>

Table 2 Modes of mushroom-type EBG with the diameter of via

4.2.2 Antenna performance

Following the parametric study, the mushroom-type EBG structure is used as the ground plane of the HPRC antenna presented in section 3.2 in order to suppress the surface waves and reduce the backlobes. The via, whose dimension is 0.1mm×0.1mm×1.15mm, is selected due to its clear wide stopband around 14GHz as shown in Figure 4.1 and the minor change of its reflection phase as shown in Figure 3.7. The simulations are carried out in Microstripes™ ver6.5. In the antenna configuration, apart from the ground plane, the only change is the distance between the PRS and the ground plane from 7.1mm to 7.2mm. This is due to the minor change of the reflection phase of the mushroom-type EBG ground plane when compared to the AMC ground plane. The rest parameters are kept the same and the PRS contains 9×9 unit cells. The offset of feeding is 0.5mm far away from the centre of the rectangular patch antenna.
Figure 4.3 shows the return loss of the HPRC antenna based on the EBG ground plane. There is not significant bandwidth improvement in the return loss by virtue of the application of the mushroom-type EBG ground plane. Figure 4.4 shows that the maximum directivity has 0.5dB enhancement by virtue of the application of the mushroom-type EBG ground plane. Figure 4.5 shows that the sidelobe and backlobe level has clearly decreased about 5dB - 10dB when compared with the results in Figure 3.18. After the application of mushroom-type EBG ground planes, it is observed that the backlobes is about -35dB. The simulation results demonstrate that the mushroom-type EBG ground plane has successfully suppressed the surface waves of the metamaterial ground plane.

Figure 4.3 Return loss of the patch-fed HPRC antenna with mushroom-type EBG ground plane
Figure 4.4 Directivity of the patch-fed HPRC antenna with a mushroom-type EBG ground plane

Figure 4.5 Radiation patterns of the patch-fed HPRC antenna with a mushroom-type EBG ground plane at 14GHz
When the above HPRC antenna with an AMC ground plane is fed by a dipole antenna, the HPRC antenna shows low backlobes compared with the HPRC antenna fed by a rectangular patch antenna. The radiation pattern at 14GHz with maximum directivity is shown in Figure 4.7. All configuration parameters are the same as the previous ones. Simulation is carried out in Microstripes™ ver6.5. However section 2.3 demonstrates that there is the strongest field at the conducting layer of AMC surfaces. The field at the bottom of the AMC ground plane would diffract out and affect the radiation patterns of the HPRC antenna. Thus it is expected that a mushroom-type EBG ground plane will reduce the effect and enhance the performance of the HPRC antenna. Then another simulation is run to verify the expectation. The simulation results are shown in Figure 4.6 and Figure 4.7. The maximum directivity 22.7dB at 14GHz is close to the value of the HPRC antenna based on a PEC ground plane as shown in Figure 3.31. The radiation patterns at 14GHz are shown in Figure 4.7. The figure demonstrates that the HPRC antenna fed by a dipole achieves low sidelobes (including backlobes) by virtue of the application of the mushroom-type EBG ground plane.

![Figure 4.6](chart.png)

Figure 4.6 Directivity of the dipole-fed HPRC antenna with a mushroom-type EBG ground plane
Figure 4.7 Radiation patterns of dipole-fed HPRC antennas with different ground planes (14GHz)
4.3 Low-sidelobe HPRC Antennas based on Novel EBG Planes

4.3.1 Higher Sidelobes in H-plane for finite structures

Figure 3.13 and Figure 3.30 show that the sidelobes in H-plane are higher than E-plane. This phenomenon is also observed in HPRC antennas fed by waveguide slot antennas [3-5]. It is known that the sidelobes of HPRC antennas are related to edge diffraction from both planes since they are finite structures. Bigger PRS and ground planes will result in lower edge diffraction effects and produce lower sidelobes [6]. Thus more simulations are run to check whether the problem can be fixed when the antenna configuration has bigger PRS and ground planes.

Simulation results show that the problem still exists when a PRS contains $17 \times 17$ unit cells and the problem will disappear when a PRS contains $21 \times 21$ unit cells. The PRS periodicity is 11mm. The square patch is $10\text{mm} \times 10\text{mm}$. The permittivity of the dielectric substrates of the PRS is 2.55 and the thickness is 1.5mm. The permittivity of the dielectric substrate on the ground plane is 2.2 and the thickness is 1.15mm. When the PRS contains $17 \times 17$ unit cells, the ground plane and the PRS plane is $200\text{mm} \times 200\text{mm}$. The simulated directivity is shown in Figure 4.8. The radiation patterns at the frequency with maximum directivity are shown in Figure 4.9. When the PRS contains $21 \times 21$ unit cells, the dimension of the ground plane and the PRS plane is $250\text{mm} \times 250\text{mm}$. The simulated directivity is shown in Figure 4.8 as well. The radiation patterns at the frequency with maximum directivity are shown in Figure 4.10. Thus a much bigger PRS surface is required to solve the problem and that will significantly reduce the aperture efficiency of HPRC antennas. Therefore other solutions are needed without the expense of aperture efficiency.
Figure 4.8 Directivity of the HPRC antennas with bigger PRS and ground planes

Figure 4.9 Radiation patterns of the HPRC antenna with a PRS containing 17×17 unit cells at 13.9GHz
Before finding the solution, we need to understand the root cause of the problem. Due to the high reflection of the PRS in the configuration of HPRC antennas, the planar resonant cavity behaves like a parallel plate waveguide but with the wave propagation constant being a complex number \[7\]. The complex wave can propagate up to a length and the length may decide the maximum directivity of antennas.

According to the geometry mode of parallel plate waveguide \[8\], there are three kinds of propagation modes in the cavity: TE, TM, TEM. From the mathematical expressions of TE/TM, it can be found that TE/TM is symmetrical in a symmetrical structure. Thus it is expected that the far-field of E-plane/H-plane yielded by TE and TM modes would be the same for HPRC antennas. The only difference of E-plane/H-field is the existence of the TEM mode. Because its phase velocity and group velocity are the same, it will not become leaky waves during the propagation in the cavity. At the edges of the planes, it will diffract and contribute to sidelobes. When the HPRC antenna is excited by a single polarization source (e.g. a dipole antenna), its E-field maybe directed towards the broadside and H-field is perpendicular to the broadside,
thus it is expected that the sidelobes in H-plane would be higher than the sidelobes in E-plane.

Therefore we can suppress the TEM mode in the resonant cavity to obtain low sidelobes in both planes in a relatively finite structure. Hence a TM bandgap along the wave propagation of the TEM mode need be introduced. There are two possible options to do that. One is to design a new ground plane that has a TM bandgap at the operating frequency of antennas (e.g. mushroom-type EBG structures). This has been demonstrated in Figure 4.5 that shows low sidelobes in both planes. The other is a novel EBG PRS. A detailed analysis of the EBG PRS has been discussed in section 2.3.2. The application of a novel aperture-type EBG surface to obtain low sidelobes in both planes will be presented as follows.

4.3.2 Novel aperture-type EBG PRS

In order to achieve a high directivity, the upper layer PRS plane in the configuration of HPRC antennas needs a high reflection coefficient [5]. Here the shape of aperture square-loop is employed to design an EBG PRS plane with a high reflection coefficient and a TM bandgap at the same working frequency. After several stages of optimisation, the desired element structure is obtained and shown in the Figure 4.11(a). The black area is copper and the grey area is dielectric. The dielectric substrate is Taconic Teflon and its thickness is 1.15mm with permittivity 2.2. The periodicity is 11mm. The inner edge of the aperture is 9.5mm×9.5mm and the outer edge of the aperture is 10.5mm×10.5mm (the width of the gap is 0.5mm).

During the optimisation, a microstrip line is placed over an EBG plane to obtain the bandgap of the EBG plane. This approximate way saves much time when compared with the computation of dispersion diagrams. The width of the microstrip line is 4mm and the length of the microstrip line is 80mm. To further save computation time, a small EBG plane consisting of 4×7 unit cells is used and shown in Figure 4.12. The simulation is carried out in Ansoft HFSSTM ver9.2. The measured results are shown in the figure as well. After the fabrication of a PRS surface, horn antennas are also used to measure the TM bandgap of the structure. It is observed that the measurement in a
microstrip line shows a broader stopband when compared with the measurement in horn antennas as shown in Figure 4.13(a). This is due to the introduction of the microstrip line that works like a ground plane. The quasi TEM mode wave only propagates between them [8]. However the space up or below the EBG surface allows wave propagation when using horn antennas.

In order to demonstrate that the bandgap of the EBG surface still works after the surface is integrated into the configuration of HPRC antennas, another configuration containing the EBG surface and a ground plane is simulated. The geometry model is shown in Figure 4.12(a). The distance between the EBG plane and the ground plane is 12mm, about half wavelength of 13GHz. The distance between the microstrip line and the ground plane is 10.85mm (=12mm – 1.15mm). The simulated bandgap is shown in Figure 4.13(a). It indicates that the bandgap of the EBG surface still works after being applied on HPRC antennas.

Figure 4.11 (a) Unit cell model of aperture-type PRS plane (b) Reflection coefficient (normal incidence) (c) Unit cell model of conducting-type PRS plane
Figure 4.12 Geometry model and photos of the novel EBG PRS
Figure 4.13 Bandgaps of the EBG surfaces

(a) Aperture-type EBG surface (TM bandgap)

(b) Conducting-type EBG surface (TE bandgap)
4.3.3 Antenna performance

The configuration of the HPRC antenna is similar to the previous one as illustrated in Figure 2.1. The new EBG PRS surface is fabricated and used in the HPRC antenna. The photo of the EBG PRS surface is shown in Figure 4.14(b). The EBG array is 9×9 unit cells printed in the centre of a 150mm×150mm dielectric slab. The permittivity of the dielectric slab is 2.2 and the thickness is 1.15mm. The dimension of the ground plane is 150×150mm. The excitation is a rectangular patch fed by a 50Ω coax-line. The size of the rectangular patch is 6.4mm×4.5mm. The offset of the feeding point is 0.8mm away from the centre of the patch. The distance between the PRS plane and ground plane is 12mm. The simulation is carried out in Microstripes™ version 6.5.

The simulated and measured return losses are shown in Figure 4.15. In Figure 4.16 the simulation shows that the maximum gain is located at 13.2GHz while the measurement result is 13GHz. Figure 4.16 also shows that the simulated maximum gain is slightly lower than the measured maximum gain. The reason is that the simulated return at 13.2GHz with maximum directivity is not completely matched and the return loss is only about –7dB. Actually the simulated maximum directivity at 13.2GHz is 21.639dB while the measured gain at 13GHz with maximum gain is only 20.8 dB.

It is observed that in both E-plane and H-plane there is a good agreement between simulations and measurements (Figure 4.17). The SLL in H-plane is about 22dB and the SLL in E-plane is close to 20dB. The asymmetrical pattern in the E-plane is caused by the offset excitation of the patch antenna. This will be further demonstrated by the following simulation when the HPRC antenna is fed by a dipole. Thus the SLL is expected more than 20dB assuming the HPRC antenna is excited by a symmetrical source. The cross polarization level is less than –25dB in both planes.
Figure 4.14
Photos of the novel aperture-type PRS and the ground plane with a rectangular patch.
Figure 4.15  Return loss of the HPRC antenna with the novel EBG PRS

Figure 4.16  Gain of the HPRC antenna with the novel EBG PRS
Radiation patterns of the HPRC antenna with the novel EBG PRS at the frequency with maximum directivity
4.3.4 Performance comparison based on different PRS

To further show the reduction of sidelobes after employing the novel EBG PRS plane, another EBG PRS plane with conducting elements is utilized. However, the structure has no TM bandgap around the operating frequency [9]. The unit cell is a square patch 10mm x 10mm as shown in Figure 4.11(c). It has a periodicity of 11mm that is equal to the periodicity of the aperture-type square loop. The thickness of the dielectric substrate is 1.15mm with permittivity 2.2. The reflection coefficient of the square patch EBG is shown in Figure 4.11(b), which is a bit higher than the reflection coefficient of the aperture-type EBG. The simulated bandgap of the conducting-type EBG using a microstripline is shown in Figure 4.13(b). There is a clear TE bandgap around 13GHz [9].

To exempt the effect of offset feeding of a patch antenna as shown in Figure 4.17, a 10mm centre-fed dipole is utilized in the simulations. Its radius is 0.25mm. The dipole is 3mm far away from the ground plane. The rest parameters are the same as the previous ones.

Similar to the HPRC antenna with the aperture-type EBG PRS, the HPRC antenna with the conducting-type EBG PRS has a maximum directivity at 13.2GHz. The HPRC antenna with the square patch EBG doesn’t show a higher directivity than the HPRC antenna with the aperture-type EBG as shown in Figure 4.18. They have similar directivities from 12.9GHz and 13.2GHz. So the radiation patterns at the frequencies 12.9GHz, 13GHz, 13.1GHz and 13.2GHz are compared in Figure 4.19, Figure 4.20, Figure 4.21 and Figure 4.22 respectively. It is observed that there is more than 10dB reduction with the sidelobes of H-planes.
Figure 4.18 Directivity comparison of HPRC antennas based on different PRS

Figure 4.19 (a) E-plane
Figure 4.19  Radiation patterns of HPRC antennas with different PRS at 12.9 GHz

Figure 4.20  (a) E-plane
Figure 4.20 Radiation patterns of HPRC antennas with different PRS at 13GHz

Figure 4.21 (a) E-plane
Figure 4.21 Radiation patterns of HPRC antennas with different PRS at 13.1GHz

Figure 4.22 (a) E-plane
4.4 Periodicity Analysis of PRS

In section 2.2.3 the relationship between the directivity efficiency and the periodicity of EBG PRS has been theoretically discussed. Related design will be presented in this section.

Let the working frequency of HPRC antennas be 14GHz and the dielectric substrate of PRS is Taconic Teflon with permittivity 2.2 and thickness 1.15mm, the calculated boundary value of grating lobes from the equation (18) is about 7mm. A novel EBG PRS, whose shape of unit cell structure is square loop, is redesigned and shown in Figure 4.23(b). To compare the result, another EBG PRS whose shape of unit cell structure is square patch with periodicity 11mm and another EBG PRS whose shape of unit cell structure is square loop with periodicity 4.5mm are utilized. All of them are shown in Figure 4.23. The black area is copper and the grey area is dielectric. The square patch is 10mm × 10mm. The outer boundary of the big square loop is 6.5mm×6.5mm and the inner boundary of the square loop is 3mm×3mm (the width of the loop is 1.75mm). The outer boundary of the small square loop is 4mm×4mm.
and the inner boundary of the small square loop is 3.45mm×3.45mm (the width of the smaller loop is 0.275mm). The simulated reflection coefficients in Microstripes™ ver6.5 are shown in Figure 4.23(d). The figure shows that the three PRS planes have the same reflection coefficient 0.967 at 14GHz.

![Image of PRS arrays and reflection coefficients](image)

Figure 4.23 Unit cell model of PRS and their reflection responses (normal incidence)

The square patch PRS array has 9×9 unit cells and the whole size is 99mm×99mm. The square loop PRS array has 14×14 unit cells and the whole size is 98mm×98mm. The smaller square loop PRS array has 22×22 unit cells and the whole size is 99mm×99mm. The other parameters are kept the same. The PRS plane and the ground plane are 150mm×150mm and the distance between the EBG PRS and the
ground copper plane is 11.2mm. Figure 4.24 shows that the maximum directivity of HPRC antenna with the square loop PRS is about 23.6dB at 14GHz, the HPRC antenna with the smaller square loop is about 23.7dB and the maximum directivity of HPRC antenna with the square patch PRS is only about 22.7dB. The radiation patterns at 14GHz are shown in Figure 4.25. Thus, about 2dB enhancement is obtained by reducing the periodicity of PRS.

The above results further validate the discussion on maximum directivity and efficiency in section 2.2.2. The square loops without grating lobes show a significant better performance (about 2dB enhancement for the maximum directivity).

![Figure 4.24 Directivities of HPRC antennas with different PRS](image-url)
Figure 4.25 Radiation patterns of HPRC antennas with different PRS at 14GHz
It must be pointed out that the periodicity and reflection coefficient of PRS planes cannot fully decide the maximum directivity of HPRC antennas. Here we present an example that a PRS with a small periodicity and a high reflection coefficient but the related HPRC antenna doesn't show a high directivity.

To obtain a small periodicity while keeping high reflection coefficient at 14GHz, aperture-type EBG is selected as a PRS plane. To get a PRS with broadband reflection coefficient, an aperture-type square patch array is chosen. The periodicity of the unit cell is 4mm and the square patch size is 2mm × 2mm. The dielectric substrate is Taconic Teflon with permittivity 2.2 and thickness 1.15mm. Its reflection coefficient is shown in Figure 4.26. At 14GHz, the reflection coefficient is about 0.98 which is much higher than the reflection coefficients shown in Figure 4.23. Thus the expected directivity would be more than 22dB.

There are 25 × 25 unit cells in the PRS plane and the substrate of PRS is 150mm × 150mm. The ground plane with dielectric substrate is also 150mm × 150mm. The distance between the EBG PRS and the ground plane is 10.15mm. The rest parameters including the excitation of a dipole are kept the same as the previous ones.

(a) unit cell
(b) reflection coefficient

Figure 4.26 Unit cell model of aperture square patch PRS and reflection response
Figure 4.27 Directivity of HPRC antenna with aperture square patch PRS

Figure 4.28 Radiation patterns of HPRC antenna with the aperture square patch PRS at 14GHz
The simulated directivities as shown in Figure 4.27 are much lower than the expected value (>22dB). The maximum directivity is only about 19dB. However, the 3-dB directivity bandwidth is about 1GHz around 14GHz. Figure 4.28 shows the radiation pattern at 14GHz with the maximum directivity. It is observed that there are not much edge diffractions from the resonant cavity. The related field distributions are shown in Figure 4.29. It is observed that the fields in the resonant cavity rapidly go weak although the reflection coefficient of the PRS is kept high.
4.5 Conclusions

This chapter initially discussed HPRC antennas based on mushroom-type EBG ground planes. The via diameter study for mushroom-type EBG structures show that the first mode might be TE or TM mode. By virtue of the application of an EBG ground plane, the HPRC antenna fed by a rectangular patch shows lower backlobes than the previous HPRC antennas with an AMC ground plane.

By introducing a novel aperture-type EBG PRS, a low-sidelobe HPRC antenna in both planes is obtained in a relatively finite structure. A conducting-type EBG PRS is utilized to compare the reduction. The simulation results show that there is a clear reduction with the sidelobes of H-plane.

By optimising the periodicity of EBG PRS, a higher efficiency HPRC antenna may be designed.
References


Chapter 5  HPRC Antennas with Tilted Beams

5.1 Introduction

Beam tilting may be an essential requirement for practical applications of low-profile HPRC antennas. In Section 2.2 the theoretical analysis of HPRC antennas with tilted beams has been discussed. The results predict that tilted beams may be obtained when the distance between the PRS and the ground plane is greater than the required resonant distance. This chapter will study HPRC antennas with tilted beams by numerical simulations. Dipoles and rectangular patches are utilized as the excitations. The proposed structure is much simpler when compared with a phased array antenna where a complex feeding network is normally required [1].
5.2 Dipole as excitation

A dipole has been demonstrated as the excitation of HPRC antenna in previous analysis and measurements. The radiation pattern of a dipole is symmetrical and the radiated wave can propagate toward two directions in the resonant cavity as discussed in section 2.2.1. If we directly employ the HPRC antenna configuration with a larger resonance distance, then a dual-beam would appear [2]. Usually a single tilted main beam is required. Here a novel antenna configuration is proposed that achieves a single directive tilted beam. The proposed model is shown in Figure 5.1. A metallic (copper) wall with dielectric substrates is utilized to block the wave propagation along one side of the resonant cavity. The beam angle can be obtained from an approximate equation given in section 1.2.

The thickness of all dielectric substrates is 1.15mm. The permittivity of the PRS substrate is 2.55. The permittivity of the ground substrate and the metallic wall substrate is 2.2. The dimension of the ground substrate and the PRS substrate is 80mm×150mm. The dimension of the metallic wall substrate is 80mm×14mm. The PRS array consists of 8×12 unit cells. The shape of the unit cell is a square patch. The periodicity of the unit cell is 11mm and the dimension of the square patch is 10mm×10mm. The distance between the PRS and the PEC boundary is 1mm.
length of the centre-fed dipole is 10mm and the dipole is 3mm far away from the
ground plane and the metallic wall. The radius is set to 0.1mm and the characteristic
impedance is 50 Ohm. Simulation is carried out in Microstripes™ ver6.5.

The simulated directivity is shown in Figure 5.2. It is observed that there is no much
difference in directivity from 13.2GHz to 15GHz. This is due to the change of the
beam angle that will be shown in Figure 5.5. To demonstrate the directivity
enhancement by the PRS, the antenna configuration without PRS is also simulated,
and the directivity is shown in Figure 5.2. It is evident that the PRS plane results in a
significant directivity enhancement.

![Figure 5.2 Directivity of HPRC antennas with tilted beams](image)

The radiation pattern at 14GHz is presented in Figure 5.3. The observed tilted beam
angle is about 19 degrees. It is worth noting that the tilted beam is directive in both
planes $\overline{E_\theta}$ and $\overline{E_\phi}$. To achieve a similar performance with conventional leaky-wave
antennas (e.g. waveguide leaky-wave antennas), an array of them would be required
(e.g. an array of leaky-waveguides) [3]. Such an array would require a complex
feeding network. The proposed structure in this section yields a much more simplified
design. The related field distribution (H-plane) and the surface current of the PRS are shown in Figure 5.4. The field distribution (H-plane) of the antenna without PRS and the surface current of the ground plane are also presented in Figure 5.4.

(a) 3-D radiation patterns

(b) H-plane

Figure 5.3 Radiation patterns of HP RC antennas with tilted beams at 14GHz
Figure 5.4 Field distribution of HPRC antennas with tilted beams at 14GHz

(a) Field distribution with PRS (H-plane)

(b) Field distribution without PRS (H-plane)

\[ E(V/m) \]

\[ -40dB \quad 0dB \]

(c) Surface current of PRS

(d) Surface current of the ground plane

\[ J(A/m^2) \]

\[ -40dB \quad 0dB \]
The theory in section 2.2.1 (the equation (13)) demonstrated that the beam could be steered by changing the distance between the PRS and the ground plane. Thus another simulation with a distance 13mm has been finished to verify that. Compared with the previous antenna configuration parameters with a distance 12mm, nothing else is changed apart from the distance.

The simulated directivities are presented in Figure 5.5(a) for the different distances (12mm and 13mm). The result further demonstrates that a wide band HPRC antenna can be achieved and the only difference among the different frequency points from 13GHz to 15GHz is the beam angle as shown in Figure 5.5(b). When the distance is 12mm, the antenna doesn't obtain a high directivity at 13GHz because the distance is shorter than the required cut-off distance of resonance. When the distance is 13mm, all directivities of HPRC antennas from 13GHz to 15GHz are about 21dB and the beam angle is 18 degrees to 33 degrees. When the distance changes from 12mm to 13mm, the beam angle at 14GHz is changed to 27 degrees from its original 19 degrees.

Figure 5.5 (a) Directivities of HPRC antennas with tilted beams
Thus the simulated results prove that the beam angle can be steered by changing the distance between the PRS and the ground plane.

From previous investigation results in Chapter 3 and Chapter 4, it is known that the dimension of the PRS will affect the radiation patterns of the antennas [4]. It is expected that the dimension of the PRS plane in the antenna configuration of Figure 5.1 will affect the radiation patterns of the antenna too. Apart from that, we need to study two more parameters: the height of the side metallic wall and the distance between the PRS and the side metallic wall. These sensitivity studies would be useful in the fabrication procedure of HPRC antennas.

The simulated tilted beam radiation patterns of HPRC antennas with the height of back metallic wall are shown in Figure 5.6. There is not any clear difference between them. Thus, when the height of the metallic wall is more than the distance (e.g. 12mm) between the PRS and the ground plane, the radiation patterns have not much difference. A lower metallic wall will definitely allow more backward leaky-waves. Therefore the cut-off height of the metallic wall equals to the distance between the PRS and the ground plane.
Figure 5.6  Radiation patterns (H-plane) of HPRC antennas with the height of metallic wall at 14GHz

Figure 5.7  Radiation patterns (H-plane) of HPRC antennas with the gap width between PRS and metallic wall at 14GHz
Subsequently the gap width between the PRS and the side metallic wall is also studied and the result is shown in Figure 5.7. It can not be observed any clear difference when the gap width changes from 1mm to 3mm. Thus small changes in the gap width are not very significant in generating the sidelobes at broadside.

The above results demonstrate that HPRC antennas with tilted beams can be obtained when fed by dipoles. However they are quite difficult to match the input impedance. To address the problem, a different excitation is employed. A patch antenna is utilized since it provides more flexibility in matching the input impedance [5].

5.3 Rectangular patch as excitation

After investigation, it has been found that HPRC antennas with tilted beams can be matched when fed by a rectangular patch. The positioning of the rectangular patch is shown in Figure 5.8.

(a) Bird's eye view  (b) Side view

Figure 5.8 Cavity model of HPRC antennas with tilted beam (fed by a rectangular patch)
Figure 5.9  Directivity of HPRC antennas with tilted beams (fed by a rectangular patch)

Figure 5.10  Return loss of HPRC antennas with tilted beams (fed by a rectangular patch) and the rectangular patch
Figure 5.11 Radiation patterns of HPRC antennas with tilted beam (fed by a rectangular patch) at 14GHz
The rectangular patch is printed on the dielectric substrate of the side wall facing the resonant cavity, not on the dielectric substrate of the antenna's ground plane. The dimension of the rectangular patch is 6.4mm x 4.7mm. The distance between the centre of the rectangular patch and the ground plane is 4mm. The offset of feeding is 1mm far away from the centre. The radius of the feeding coaxline is set to 0.1mm and the impedance is 50 Ohm. Other parameters are kept as above. The simulation is carried out in Microstripes™ ver6.5.

The simulated directivity is shown in Figure 5.9. The bandwidth of the directivity is a bit narrower than the antenna fed by a dipole. The return loss of the HPRC antenna is
shown in Figure 5.10. The $S_{11}$ at 14GHz is lower than $-10$ dB. The bandwidth of the return loss is only about 0.3GHz. The isolated rectangular patch with the metallic wall is also simulated and shown in Figure 5.10 as well. It is observed that the return loss between the rectangular patch antenna and the patch-fed HPRC antenna are very similar. Therefore it would be straightforward to match the return loss of the HPRC antennas by using a pre-designed patch at the desired operating frequency.

The radiation pattern at 14GHz is shown in Figure 5.11. The sidelobes beside the mainlobe are higher when compared to the sidelobes of the HPRC antennas fed by dipoles. This is due to the surface wave excited by the rectangular patch antenna [5]. The related field distribution is shown in Figure 5.12.
5.4 Conclusions

This chapter demonstrated that HPRC antennas with tilted beams could be obtained when the related conditions are met. A metallic wall is utilized to block the wave propagation on one side of the resonant cavity. Thus the whole HPRC antenna can obtain a single tilted beam. The tilted beam is directive in the principal planes $E_\theta$ and $E_\phi$. The configuration is much simpler than the configuration of conventional microstrip array antennas or leaky-waveguide arrays that are used in order to obtain a similar pattern. The beam angle can be steered by changing the distance between the PRS and the ground plane. A dipole and a rectangular patch can be utilized as the excitations of HPRC antennas. The return loss of a HPRC antenna with tilted beams can be matched when the antenna is fed by a rectangular patch.
References


Chapter 6  Conclusions and Future Works

This thesis has researched and described a study into a new class of metamaterial-based high-directivity planar resonant cavity antennas. The HPRC antennas may find numerous applications in various modern wireless communication systems such as fixed wireless access, e.g. the LMDS systems and point-to-point links. Compared with recently reported HPRC antennas [1-3], the HPRC antennas using metamaterial surfaces as introduced in the thesis show lower profile, lower sidelobes, higher efficiency and tilted beams with simple excitations.

Ray analysis of the beam forming function has been utilized to predict the possibility of low-profile (including subwavelength) resonant cavities and tilted beams. The theoretical analysis of periodic aperture arrays and grating lobes helps to design higher aperture efficiency HPRC antennas using EBG PRS planes.
Two types of metamaterial surfaces (Artificial Magnetic Conductor and Electromagnetic Bandgap) have been investigated before their further applications in the configuration of HPRC antennas. Reflection responses and dispersion diagrams of the structures have been obtained (in Microstripes, Ansoft designer and HFSS™).

With the introduction of AMC, the distance between the resonant cavity planes can be reduced from its original half-wavelength into quarter-wavelength. Simulations show that the application of AMC also leads to higher backlobes (about 10dB enhancement) when the antenna is fed by a rectangular patch. To accommodate the rectangular patch, several central elements in the conducting layer of AMC are removed. The removing would yield much loss (about 2~3dB) in the directivity of HPRC antennas. The application of metamaterial ground planes with negative reflection phases further reduced the profile of HPRC antennas. In the thesis a λ/6 distance is validated by related measurements. A dipole is selected as the excitation of the sub-wavelength resonant cavity to avoid the just discussed shortcoming by a rectangular patch. It is also found that the non-full reflection of metamaterial ground planes hardly affects the maximum directivity of low-profile HPRC antennas since the loss is very small (about 0.2dB).

The higher backlobes of low-profile HPRC antennas can be reduced by virtue of the application of mushroom-type EBG ground planes. Simulations also demonstrate that the application of the mushroom-type EBG ground planes can reduce the sidelobes of low-profile HPRC antennas when they are fed by dipoles.

According to the simulation results, it is observed that HPRC antennas (in a relatively finite structure) generally show higher sidelobes in H-plane than in E-plane. Theoretical analysis indicates that the existence of TEM modes in the resonant cavity contributes the problem. Subsequently a novel aperture-type single-layer EBG PRS is utilized to get low sidelobes HPRC antennas in a relatively finite structure. The aperture-type EBG surface has a high reflection coefficient at the normal direction and a TM bandgap along its surface. The TM bandgap can be used to suppress the TEM mode in the resonant cavity and thus the HPRC antenna (in a relatively finite structure) obtains low sidelobes in both planes. A conducting-type PRS with a same periodicity but having a TE bandgap is utilized to compare the reduction. A
significant reduction (more than 10dB) in the sidelobes of H-planes is observed. To validate the above result, a low-sidelobe HPRC antenna fed a rectangular patch is fabricated and measured. There are very good agreements between simulations and measurements.

Following up the previous theoretical analysis of periodic aperture arrays and grating lobes, an optimisation of the EBG PRS periodicity is carried out in order to obtain higher efficiency HPRC antennas. Three different EBG PRS with the same reflection coefficient are employed in the thesis. All configuration parameters of those HPRC antennas are same. Simulation results show that the HPRC antenna shows the best performance when the EBG periodicity is small enough that it avoids the grating lobes. However the periodicity and reflection coefficient of PRS are not the only factors that decide the maximum directivity of HPRC antennas. A related example is presented to demonstrate that.

At some practical applications, low-profile HPRC antennas with tilted beams are essentially required. A phased array antenna has been used as an excitation of HPRC antenna to obtain a high-directivity tilted beam [5]. In the thesis, simulation results demonstrate that HPRC antennas with novel configurations can obtain single tilted beams only with a simple excitation (e.g. a dipole). The main beam is high-directive in principal planes ($\overline{E}_\phi$ and $\overline{E}_\theta$). The beam angle can be steered by changing the distance between the PRS and the ground plane. The return loss of HPRC antennas can be matched when they are fed by a rectangular patch.

Continuing with the research as discussed in the thesis, there are several interesting topics for future works:

**A. HPRC antennas with single tilted beams**

In the thesis only simulation results are presented. For the further application, related measurements are needed. A rectangular patch antenna can be selected as the feeding of the HPRC antenna. However the surface wave excited by a rectangular patch would yield higher sidelobes as shown in Figure 5.11. If a dipole is directly utilized as the excitation of the HPRC antenna, it is difficult to match the input impedance. A high impedance surface has been utilized to match...
the monopole antenna [7]. Thus a high impedance surface might be introduced to match the HPRC antenna fed by a dipole. Figure 5.4 and Figure 5.12 show that the field distributions of H-planes in the resonant cavity don’t smoothly go to weak. Some investigations might be required.

B. Planar High-gain Leaky-wave Antennas

A subwavelength HPRC antenna has been experimentally demonstrated in the thesis. According to the equation (8) and Figure 2.8, it is expected that the profile of HPRC antennas can be further reduced. The last configuration of HPRC antennas might consist of a dielectric slabs with embedded single or double layer metallic arrays. Some new feeding technology such as coplanar waveguides might be utilized. Thus, planar high-gain leaky-wave antennas can be achieved.
References


