Improvement of antenna performances by using ferrites and periodic structures

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**IMPROVEMENT OF ANTENNA PERFORMANCES BY
USING FERRITES AND PERIODIC STRUCTURES**

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Improvement of Antenna Performances by Using Ferrites and Periodic Structures

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Abstract

This thesis investigates various methods to improve the performances of antennas. For different applications, the antennas are expected to have corresponding performances and characteristics. The efficiency is one of the most important parameters that should be considered, and the losses determine the efficiency directly. For different types of antennas operating in different frequency bands, there exist different factors which influence the performances. For example, in the HF band (3 MHz-30 MHz), the performances depend on the mismatch loss and the size of the antenna. In the microwave frequency band, besides the mismatch loss, the dielectric loss of the substrate material is also one of the most important factors. Whereas in the millimeter wave frequency band, the loss in the feeding network is the factor that limits antenna gain and efficiency when the sub-antennas are all matched well.

Considering various influence factors, different techniques and methods are investigated to improve the performances. First, the compact and effective ferrites are selected to improve the performances of the HF antennas. Effective methods of using ferrites are proposed to improve the performances of the HF antenna. By analyzing the current in the ground plane of the antenna, the

ferrites can be arranged effectively and efficiently at areas with high current density, so the effect of image current is reduced. Moreover, with the ferrite loadings, the bandwidth of antenna is increased significantly, and the size of antenna is decreased.

For the microstrip antennas which work in the microwave band, the widely used high-loss material is the primary factor to influence the efficiency of the antenna. It is a commercial method to apply planar periodic structures in the ground plane or on the patch. When the periodic structures are arranged in the ground plane, the substrate does not perform the properties of the base material. The effective permittivity and permeability of material can be obtained by simulating the unit cell. With the periodic structures in the ground plane, the consumed power in the antenna is reduced, and the bandwidth of the antenna is increased. When the planar structures are arranged periodically on the patch, the periodic structures can work the same as the patch. More energy can be radiated from the slots in the patch. By analyzing the dispersion relation, the resonant frequencies can be derived. With the periodic structures in the patch, the bandwidth of the antenna is increased greatly.

For the millimeter wave antenna, the loss in the feeding network leads to the small number of antenna elements in an array, and the gain of antenna array is limited. A low loss transmission method, substrate integrated waveguide, is proposed to construct the feed network instead of the traditional microstrip line. Moreover, the geometry of the antenna array is simplified. The losses within the antenna array are decreased further. As a result, the gain of the antenna array is increased.

In this thesis, the techniques to improve the antenna performances are proposed over three different frequency ranges. Due to the different frequency ranges, different techniques for antenna optimization are required. The high-performance antennas are simulated, and the simulation results are compared with the measurement results. The antennas demonstrate high performances with good efficiency and small size.

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List of Abbreviations

HF	High Frequency
SIW	Substrate Integrated Waveguide
VSWR	Voltage Stand Wave Ratio
EBG	Electromagnetic Bandgap
MEMS	Micro-Electro-Mechanical System
RF	Radio Frequency
CM	characteristic mode
FSS	Frequency Selective Surface
VHF	Very High Frequency
UHF	Ultra High Frequency
EM	Electromagnetic
DPS	Double Positive
DNG	Double Negative
ENG	Epsilon-Negative
MNG	Mu-Negative

TE	Transverse Electric
TM	Transverse Magnetic
TEM	Transverse Electric and Magnetic
PEC	Perfect Electrical Conductor
CSRR	Complementary Split-Ring Resonators
EML	Embedded Meander-Line
PCB	Printed Circuit Board
UC-EBG	Uniplanar Compact Electromagnetic Bandgap

Chapter 1

Introduction

1.1 Motivation

The radiation efficiency of an antenna is the ratio of the radiated power to the total input power, which is shown in (1.1). The total efficiency is considered the radiation efficiency and the impedance mismatch loss (M_L) as shown in (1.2). The efficiency discussed in this thesis is the total efficiency. The efficiency of an antenna is one of the most important parameters of an antenna. High directivity is not always an advantage, but high efficiency is advantageous. The antenna gain (G) can be related to the directivity (D) by (1.3). For any antenna that has a constant directivity, the high efficiency means high gain. As a result, the achievable propagation range of an antenna will be large with the same input power.

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$$E_r = \frac{P_{radiated}}{P_{input}} \quad (1.1)$$

$$E_t = M_L \cdot E_r \quad (1.2)$$

$$G = E_r D \quad (1.3)$$

For antennas operating at different frequency bands, the challenges to improve the antenna efficiency are different. The losses determine the efficiency directly. The antenna losses are typically due to the conduction losses, the dielectric losses, and the impedance mismatch losses. The conduction losses are caused by the finite conductivity of the metal which forms the antenna. The dielectric losses are due to the conductivity of a dielectric material near an antenna. The impedance mismatch losses usually exist in the poor impedance matched antennas.

In the high frequency (HF) range, the physical size of the passive antenna is large since the wavelength at HF band is large. The HF antenna is usually desired to be small in size because of the scarcity of land and space resources in some cases. However, when the size of the antenna is reduced, the efficiency of the antenna is also reduced. For example, the efficiency of bifolded monopole is less than -20 dB. Thus, it is really a challenge to improve the efficiency of HF antenna while reducing its size.

The microwave band is the most widely used wireless communication band

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in our daily life. The microwave band is applied widely in personal communication devices, e.g., WLAN, 2G, 3G, 4G telecommunication. The microstrip antenna is popular in microwave band, because of its small size, low cost, ease of fabrication, and it is easy to be integrated on printed circuit board. All these advantages make microstrip antenna suitable for industrial production. The low-cost materials are usually chosen to design and fabricate microstrip antennas to reduce the cost further, but the low-cost materials tend to be high-loss, which leads to the low efficiency of the antenna. Some energy is consumed within the antenna, and less energy will be radiated out into free space. Another shortcoming of the microstrip antenna is the narrow operating bandwidth. Thus, it is challenging to reduce the losses within the antenna and increase the bandwidth.

In the millimeter wave frequency range, there has been an increasing interest recently because of its high communication capacity. In this frequency range, especially around 60 GHz, the signal is reduced rapidly in air, so a high gain antenna is necessary. The antenna array is one of the methods to increase the gain and directivity communication distance. The more the antenna elements in an array are, the higher gain it can provide. However, the loss in the feed network limits the size of an array. Additionally, the complex geometry of the antenna array also lengthens the transmission path and increases the loss in the transmit network. Thus, it is useful to improve the gain of the millimeter wave antenna by using a high efficient transmission method and simplifying the geometry of antenna.

1.2 Objectives

Based on the motivation highlighted in the previous section and the influence factors of the performances of an antenna, the objectives of the thesis are listed as following.

- Design and implementation of broadband, small-size, and near-ground HF antennas.
- Design and implementation of broadband microwave antennas.
- Design and implementation of high efficiency and high gain millimeter wave antenna arrays.

1.3 Major Contributions of the Thesis

In this thesis, the proposed objectives have been achieved. The performances of different types of antennas operating in different frequency bands are improved. Several important contributions have been made.

The performances of HF antenna are improved by using the commercial ferrite tiles effectively. The image current theory is used to analyze the mutual resistance of the antenna. By analyzing the strength and distribution of the image current, the ferrite is applied effectively to reduce the effect of the image current. The commercial ferrite is also used to reduce the height of HF antenna. By loading the ferrites, the lower bound of bandwidth is extended, and the miniaturization of HF antenna is implemented.

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The performances of the microwave antenna are improved by etching periodic structures. By etching the structures in the ground plane, the efficiency and bandwidth of the patch antenna are increased, while the size of antenna is decreased. By using the periodic structures as the patch, more energy can be radiated into the free space, and the bandwidth is also increased.

The efficiency of the millimeter wave antenna array is improved by using the low loss feed network. The low loss transmission method, substrate integrated waveguide(SIW), is used to form the feed network for the large antenna array. The grid array antenna with subarrays is proposed to avoid the radiation distortion caused by the special structure of grid array. With the high gain antenna elements and low loss feed network, the high gain millimeter wave antenna array is implemented.

1.4 Organizations of the Thesis

The rest of the thesis is organized as follows.

Chapter 2 introduces the background and literature review. Some related works and methods are included.

In chapter 3, a method to implement small-size near-ground broadband HF antenna is proposed. The ferrite tiles are used to improve the performances of the HF antenna. Since the current density on the ground is found to indicate the magnetic field and image current density near the ground, ferrite tiles are used to coat the ground plane in regions where there is a high current density concentration. The ferrites can be used to reduce the effect of image current

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at any desired frequencies, so the antenna can radiate efficiently. In addition, the end loading is proposed to broaden the bandwidth and reduce the size of HF antenna. By loading with ferrites, the lower bound of bandwidth is extended. In order to verify the proposed methods, an HF low-profile antenna is implemented and fabricated. The height of this antenna can be reduced to 0.023λ by using the ferrite tiles, and the lower bound of the bandwidth can be reduced from 7 MHz to 2 MHz. Additionally, by modifying the current on the antenna, the total efficiency of the HF antenna is improved further. The small-size near-ground antenna is improved to work well in the entire HF band.

In chapter 4, a compact microstrip antenna with periodic patterns etched in the ground plane is proposed. By loading the periodic patterns, the total efficiency of the patch antenna is improved to as high as 88.5%, even it uses the high-loss material as the substrate. The etched patterns are Jerusalem crosses. In this chapter, the effective permittivity and permeability of the substrate are derived from the simulated results of a unit cell. Then a patch is stacked, and the patch antenna is designed to operate at 2.4 GHz. Compared to the conventional patch antenna, the proposed antenna achieves a 68.38% miniaturization of the patch, and a 2.84 times impedance bandwidth broadening. The proposed antenna has a small dimension in comparison to several published compact antennas. Furthermore, the operating frequency of the proposed antenna is proposed to tune over a large range of frequencies by physically adjusting the length of the surrounding slots or using tunable inductive elements. The proposed antenna is fabricated and measured. The measurement results are

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found to agree well with the simulation results.

In chapter [5](#), a broadband microstrip antenna is proposed. In the design, the periodic structures work as the patch which is excited by a slot in the ground. The transmission-line model integrated with the dispersion relation of a composite right/left-handed structure is applied to analyze the modes and estimate the resonant frequencies. In order to ensure the broadband width, 8 squares are etched in the ground plane to achieve good impedance match. The -10 dB bandwidth is from 4.69 GHz to 7.47 GHz. The realized gain of the proposed antenna is between 6.4 dBi and 10.4 dBi. The proposed antenna is fabricated and measured, and the measured results agree well with the simulated ones.

In chapter [6](#), a high gain antenna array operating in E-band is presented. The grid array antenna is used as the antenna element in the array due to its high gain and compact dimension. The subarrays of antenna are fed by a transverse slot, and they are connected by the low loss feed network, substrate integrated waveguide (SIW), to ensure the high efficiency. The whole antenna array is excited by the standard waveguide WR-12. In order to steer the radiation beam, a phased array is proposed. In the phased antenna array, a simple phase shifter is proposed. The phase shifter can be realized without complicating the geometry of the antenna array. The whole antenna array, including the antenna elements, feed network, and phase shifter, can be realized by using only two layers of substrate, and three layers of metal. The proposed antenna can be implemented on a low-cost soft RF substrate. The beam can be steered in the range of -65° to 65° .

1.4. ORGANIZATIONS OF THE THESIS

The thesis is concluded in Chapter [7](#) and the promising research directions are recommended.

Chapter 2

Literature Review

As mentioned in chapter [1](#), the efficiency, bandwidth and size are important for antennas, and there are many challenges involved in the improvement of the performances in different frequency bands. Many researchers have investigated effective and efficient methods to improve the performances of antennas.

Due to the long wavelength of the HF band, the size of the HF antenna is huge. Thus, it is desirable to reduce the size of the HF antenna. However, by reducing the size of the antenna, the efficiency of the antenna is also reduced. Shih *et al.* proposed to use platform mounted antenna to excite characteristic modes (CMs) of platform to increase the total efficiency and gain at lower frequencies [\[1\]](#). The platform is relative to the wavelength of the lower bound of the HF band. By exciting partial or full size of the platform, the effective size of antenna is increased so that the efficiency at lower frequencies is improved. However, the geometry and size of the platform affect the performance of this design greatly. This method cannot be applied in all types of antennas and

platforms. Choi *et al.* designed a compact HF antenna which has a high efficiency itself [2]. The antenna is designed based on the electrically small folded dipole. By using 180° phase shifter to connect two dipoles, the efficiency of antenna is enhanced. However, similar to all the horizontal polarized antennas, the balanced dipole antenna has to be placed high enough to operate. Lin *et al.* implemented a low profile HF antenna [3]. The commercial ferrites are used to cover the entire ground plane to reduce the mismatch loss and increase the efficiency at lower frequencies. However, the large amount of commercial ferrites makes the total weight and price quite high. Moon *et al.* used commercial ferrites to improve the efficiency of the antenna [4]. The main energy stored in the antenna is magnetic. By using the ferrites to reduce the energy stored in the antenna, more energy can be radiated out, and the total efficiency of the antenna is increased. Considering the realization, dimension and total cost, we choose commercial ferrites to improve the efficiency of HF antenna. In order to use the ferrites more efficiently, the methods used in [3] and [4] will be described in detail in section 2.1.

Comparing with HF antennas, the microwave antennas are smaller and need much less space. The size of the antenna is not the dominant factor any more. Other factors become more dominant, such as the losses in dielectric material. The microstrip antennas usually use low cost and high loss material as the substrate, so the efficiency of the antenna is low. In [5], the coupling structures are proposed to increase the efficiency of microstrip antenna. Chen *et al.* designed a coupling structure antenna without resistors. Without using the resistors,

the loss in the divider network is decreased, and as a result, the efficiency is increased. However, the geometry of this kind of antenna is complex, and it is hard to fabricate. Yang *et al.* presented a high-efficiency antenna with planar structure in [6]. The antenna uses metasurface as the ground plane. The periodic structure possesses the characteristic of low loss, so the efficiency of the antenna gets high. Pan *et al.* designed an antenna using the metasurface as the patch. In [7], it is shown that the frequency selective surface (FSS) has the performance of filtering, and the metasurface inherits this property. Thus, the metasurface patch can enhance the efficiency of antenna because of the eliminated insertion loss of metasurface. The periodic structure can make the antenna more efficient [8–11]. It is helpful to know the electromagnetic property and dispersion relation of the material when we design the antenna, so section 2.2 will present how to get the effective permittivity, effective permeability, and dispersion relation of material.

The factors influencing the performances of millimeter wave antenna are different from HF antennas and microwave antennas. In the millimeter wave communication system, the loss in propagation is significant, e.g., 15-30 dB/km at 60 GHz [12], so the high directivity and high efficiency antenna arrays are required. The loss in the feed network of the antenna array becomes the main factor influencing the total efficiency of the antenna array, so it is necessary to keep the loss in feed network as small as possible. Zarifi *et al.* presented a high-efficiency slot antenna array in [13]. The antenna is designed based on waveguide gap technique, so it does not need electrical contact between

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each layer of the antenna. However, this antenna is complex, and it consists of three layers. It is hard to fabricate. In [14], Wu *et al.* designed a high-efficiency millimeter wave antenna which can be fabricated using soft substrate. The proposed antenna is an array of patch antennas. The patches are fed by slots, which avoid the use of high loss millimeter wave connectors. In the feed network, the microstrip lines are used. There is a large amount of loss in the feed network, because the width of microstrip line and the thickness of the substrate is relatively large. In [15], Park *et al.* used substrate integrated waveguide to design the feed network. In order to design high efficiency and easy-fabrication millimeter wave antenna, the design of SIW and gap coupling technique are the necessary techniques. We will introduce the advantage of SIW, how to design SIW, and how to couple energy from (or to) the SIW in section 2.3.

2.1 Using Ferrite Material to Improve Efficiency in HF Band

2.1.1 Ferrite Tiles over Entire Ground Plane to Improve Return Loss

It is necessary for a horizontally polarized dipole antenna to be located at a height of at least $\lambda/2$ to ensure the impedance and radiation performances of the antenna [16]. When the current flowing through the antenna is placed near the ground plane, the effect of the image current cannot be ignored [17]. The

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image current on the ground plane is in the opposite direction to the current in the antenna as shown in Fig. 2.1. The electromagnetic wave radiated from the antenna is reflected by the ground plane. The reflected wave undergoes a 180 degree phase shift. This reflected wave is therefore eliminated by the wave from the image current, which results in the cancellation of radiated power by the image current.

In [3], the authors designed a broadband (3 MHz - 6 MHz) bow-tie shaped HF antenna which is near the ground plane with a very low height by using ferrite materials to eliminate the destructive impact of the image current. When the antenna is near the conductive ground plane, the effect of image current cannot be ignored. The radiation resistance (R_r) of the antenna will be as (2.1) [17].

$$R_r = R_{11} - R_m \quad (2.1)$$

where R_{11} is the real part of the self impedance of the antenna, and R_m is the mutual resistance of the antenna and its image.

The trend of the radiation resistance over various heights of antenna is shown in Fig. 2.2 [17]. It can be observed that the radiation resistance (R_r) is decreased sharply when the antenna is very near to the ground plane, which is because of the increase of the mutual resistance (R_m).

In order to reduce the impact of image current and the mutual resistance, in [3], the authors used commercial ferrite tiles to cover the entire conductive

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Figure 2.1: Illustration of image theory.

Figure 2.2: Trend of radiation resistance with various heights of antenna [17].
ground. The properties of the used ferrite are shown in Fig. 2.3 [18]. The configuration of the antenna is shown in Fig. 2.4 [3].

The antenna is very near to the ground with the height of 1 meter ($0.02\lambda_0$). By laying the commercial ferrite tiles over the entire ground plane ($0.44\lambda_0 \times$

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Figure 2.3: Properties of the used commercial ferrite SN20 [18].

Figure 2.4: Configuration of near-ground HF antenna [3].

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$0.24\lambda_0$), the voltage standing wave ratio is improved. In Fig. 2.5 [3], the voltage standing wave ratio of the antenna under different situations are compared. It can be seen that the large amount of commercial ferrites can be used to improve the performance of the near ground antenna in terms of voltage standing wave ratio. However, the total efficiency is not shown. Even though the return loss is reduced, the placement of lossy ferrite results in the low efficiency of the antenna.

Figure 2.5: Simulated VSWR of the antennas without ground, with PEC ground, and with ferrites covered ground [3].

2.1.2 Ferrite Tiles under Antenna to Improve Radiation Efficiency

In section 2.1.1, Lin *et al.* used commercial ferrites to improve the voltage standing wave ratio and return loss of the antenna. The ferrite is used as the absorber

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to eliminate the image effect, so the total efficiency of the antenna may not be increased. Zhao *et al.* proposed broadband monopoles for VHF-UHF band applications in [19] and [20]. However, the total efficiency of the antenna is still too low. In [4], the authors implemented a broadband VHF antenna. The ferrite bars are placed between the body of the antenna and the ground plane to improve the realized gain of the antenna.

The geometry of the VHF antenna is shown in Fig. 2.6(a), and the antennas loaded with various ferrite patterns are shown in Fig. 2.6(b) - Fig. 2.6(d). The used ferrite is the same as that used in [3], and its properties are shown in Fig. 2.3. The simulated results are shown in Fig. 2.7. From Fig. 2.7, it is found that the ferrite loading can improve the realized gain of the antenna at lower frequencies. The ferrite loading does not improve the realized gain of the antenna at higher frequencies. The realized gain at higher frequencies is even decreased slightly. The use of ferrite loading does not increase the total volume of the antenna, but the weight of the antenna is increased greatly because of the high weight of the used ferrite. In order to decrease the weight of the antenna, the layout of the commercial ferrite bars is required to be optimized.

The authors, Moon *et al.*, used four loading ferrite bars with different heights and widths to decrease the weight of the antenna while maintaining the performances. The geometry of the ferrite loaded antenna is shown in Fig. 2.8(a) [4]. The magnetic field distribution within the ferrite bars is analyzed, and the magnetic field in the loading ferrite bars is shown in Fig. 2.9. In order to decrease the required ferrite further, the authors proposed to laying the ferrite

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Figure 2.6: VHF antenna with various ferrite loading configurations [4].

Figure 2.7: Realized gain of antenna with various ferrite loading configurations [4].

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in the areas with high magnetic field. The loading ferrite bars are minimized.

The optimized ferrite loading configuration is shown in Fig. 2.8 (b).

Figure 2.8: VHF antenna with optimized ferrite loading configurations [4].

Figure 2.9: Magnetic field distribution in the loading ferrite bars [4].

The voltage standing wave ratio is shown in Fig. 2.10 [4]. Comparing to the antenna without ferrite loading, the voltage standing wave ratio is improved greatly. The main energy stored in the loop antenna is magnetic. The loading ferrite has the property of high permeability. As a result, the energy stored in

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Figure 2.10: VHF antenna with various ferrite loading configurations [4].

the antenna is reduced, and the Q factor gets lower [21]. The voltage standing wave ratio becomes lower, and the antenna gets more broadband.

The realized gain of the antenna with optimized ferrite bars is shown in Fig. 2.7. Comparing with the antenna without ferrite loading, the realized gain is improved at the lower frequencies, but it is decreased at the higher frequencies. Comparing with the antenna with one full piece of ferrite (as shown in Fig. 2.6 (a)), the gain is almost unchanged at the lower frequencies, and it is improved slightly at the higher frequencies. The realized gain is improved at higher frequencies because of the less use of ferrite. The ferrite can absorb the electromagnetic energy, so the less ferrite allows more energy to be radiated. At the lower frequencies, the return loss is reduced significantly. Less energy will be reflected. Therefore, the total efficiency and the realized gain are improved.

2.2 Using Artificial Materials to Improve Efficiency in Microwave Band

In HF band, the ferrite is effective to improve the performances of antenna. When the frequency moves to higher band, microwave band, the periodic structure is popular for the improvement of antenna efficiency due to its special EM properties. Permittivity and permeability are two of the most important parameters. When we design a microstrip antenna, the permittivity (ϵ) and permeability (μ) can be used to describe the EM properties of the substrate. Most occurring materials in nature is double positive medium, which is a medium with both permittivity and permeability more than 0 ($\epsilon > 0$, and $\mu > 0$). In some applications, the materials with non-naturally occurring or specific properties are required, so the artificial materials appear [22].

In 1914, by embedding small wire helices, Lindell and the other authors worked on the artificial material [23]. From then on, artificial materials have attracted more and more attention, and also known as the periodic structure. In 1948, Kock [24] used periodic structure to make lightweight microwave lenses.

The artificial material based on periodic structure can be constructed by embedding various elements with novel forms in some host media, as shown in Fig. 2.11. The periodic structure provides a large number of independent parameters. The high degree of freedom allows designers to get the material with certain permittivity and permeability.

The material can be classified based on the value of permittivity and per-

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Figure 2.11: Top view of a material based on periodic structure synthesized by embedding various inclusions in a host medium [22].

meability. In Fig. [2.12], the classification of material is illustrated graphically. As mentioned previously, the most natural material is double positive (DPS). When the permittivity and permeability are both less than zero ($\epsilon < 0$, and $\mu < 0$), the material is double negative (DNG). This kind of material has not been found in nature to date. The material with negative permittivity and positive permeability ($\epsilon < 0$, and $\mu > 0$) is epsilon-negative medium (ENG). If the permittivity is positive, and the permeability is negative ($\epsilon > 0$, and $\mu < 0$), the material is mu-negative (MNG). It should be noted that the permittivity and the permeability of material are not always constant. They are changed with surroundings, and they are frequency dependent. Thus, the material may have different properties in different frequency bands. For special application, the periodic structure should be designed and fabricated to have the desired effective permittivity and permeability in the certain frequency band.

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Figure 2.12: Classification of material [22].

2.2.1 Effective Permittivity and Permeability of Artificial Material

When the periodic structure based material is used as the substrate of microstrip antenna, it is necessary to know its permittivity and permeability. The permittivity and permeability are two of the most important parameters of material. They describe the electronic and magnetic properties of the material. However, it is difficult to measure the permittivity and permeability of the periodic structure based material by using a simple equipment. Thus, it is necessary to get the values of ϵ and μ from the available parameters.

In general, the permittivity and permeability of a passive medium are

$$\epsilon = \epsilon_r \epsilon_0 = (\epsilon'_r - j\epsilon''_r) \epsilon_0 \quad (2.2)$$

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and

$$\mu = \mu_r \mu_0 = (\mu'_r - j\mu''_r) \mu_0 \quad (2.3)$$

where $\epsilon_0 = 8.85 \times 10^{-12} F/m$ and $\mu_0 = 4\pi \times 10^{-7} H/m$.

It is easy to calculate and measure the transmission and reflection wave in free space. In [25], Cotuk gave the general form of the wave matrix in a three layers mediums. In [26], Chen *et al.* used the reflection coefficient and transmission coefficient to define the S-parameters. The impedance and refractive index can be expressed in terms of S-parameters [27]. Thus, the effective permittivity and effective permeability can be extracted by simulating the unit cell easily by (2.4)-(2.8).

$$\epsilon_{eff} = \frac{n}{Z} \quad (2.4)$$

$$\mu_{eff} = nZ \quad (2.5)$$

$$Z = \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (2.6)$$

$$n = \frac{1}{k_0 t} \{Im[\ln(e^{ink_0 t})] + 2m\pi - jRe[\ln(e^{ink_0 t})]\} \quad (2.7)$$

$$e^{ink_0 t} = \frac{S_{21}}{1 - S_{11} \frac{Z-1}{Z+1}} \quad (2.8)$$

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2.2.2 Dispersion Relation of Unit Cell

The periodic structure can be used as the substrate to improve the efficiency of antenna, and it can also be used as the radiation elements. In [28] and [29], Liu *et al.* used periodic structure as the radiation elements of the antenna.

The radiator is formed by a 4×4 array in [28], which is shown in Fig. 2.13. Each unit is a mushroom-like patch. The dielectric constant of the substrate and the dimensions of the antenna is shown in Table 2.1. The fringing field cannot be ignored at the open edge of each unit cell, and the extended length (ΔL) can be calculated by (2.9) and (2.10) [30].

Figure 2.13: Geometry of the proposed slot-fed mushroom antenna and the coordinate system [28].

$$\frac{\Delta L}{h} = 0.412 \frac{(\epsilon_{eff} + 0.3)(w/h + 0.262)}{(\epsilon_{eff} - 0.258)(w/h + 0.813)} \quad (2.9)$$

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Table 2.1: The dielectric constant of substrate and diemensions of the broad-band antenna.

ϵ_r	3.38	h	3.25 mm
G_L	60 mm	G_W	60 mm
w	9 mm	g	1 mm
p	10 mm	W_p	39 mm
L_p	39 mm		

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + 12 \frac{h}{w}\right)^{-1/2} \quad (2.10)$$

where h is the thickness of the substrate. ϵ_r is the dielectric constant of the substrate. w is the width of the unit cell.

The propagation constant in this effective extended region is shown by (2.11).

$$\beta_{eff} = k_0 \sqrt{\epsilon_{eff}} = \frac{2\pi f}{c} \sqrt{\epsilon_{eff}} \quad (2.11)$$

where k_0 is the wave number in free space. c is the speed of light. f is the frequency of operation.

When the gap between every two neighbouring unit cells is smaller or approximates to the extended edge (ΔL), the array of the unit cells can be considered as an entire piece, and it can work as the patch of the antenna. The resonant frequency of TM_{10} can be obtained by (2.12) and (2.13).

$$\beta_m W + 2\beta_{eff} \Delta L = \pi \quad (2.12)$$

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$$W = pN \quad (2.13)$$

where β_m is the propagation constant in the periodic structure. W is the width of the radiator. p is the period of the unit cell. N is the number of the unit cells in one direction. In [28], N is 4.

The resonant frequency of TM_{20} is shown in (2.14).

$$\beta_m p/2 + 2\beta_{eff} \Delta L = \pi \quad (2.14)$$

In order to estimate the resonant frequencies, it is necessary to know the propagation constant or the dispersion relation of the unit cell. The dispersion relation of the unit cell can be derived from the full wave simulation of one unit cell. The simulated model of the unit cell is shown in Fig. 2.14. It is easy to get the propagation constant from the simulated S-parameters. $e^{\gamma z}$ is the eigenvalue of ABCD matrix by using (2.15), and ABCD matrix can be obtained from S-parameters.

Figure 2.14: Simulated model of the unit cell [28].

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$$e^{\gamma z} = eig([ABCD]) \quad (2.15)$$

From (2.12) and (2.14), we can get

$$\frac{\beta_{mp}}{\pi} = 1 - \frac{2\beta_{eff}\Delta L}{\pi} \quad (2.16)$$

and

$$\frac{\beta_{mp}}{\pi} = 2(1 - \frac{2\beta_{eff}\Delta L}{\pi}) \quad (2.17)$$

Bring (2.11) into (2.16) and (2.17), the dispersion of TM_{10} and TM_{20} mode can be presented by (2.18) and (2.19).

$$\frac{\beta_{mp}}{\pi} = 1 - \frac{4f\Delta L\sqrt{\epsilon_{reff}}}{c} \quad (2.18)$$

and

$$\frac{\beta_{mp}}{\pi} = 2(1 - \frac{4f\Delta L\sqrt{\epsilon_{reff}}}{c}) \quad (2.19)$$

By using the simulated results of one unit cell and the equations (2.15), (2.18), and (2.19), the resonant frequencies can be obtained. The resonance frequencies of the antenna from [28] for TM_{10} mode and TM_{20} mode are 5.23 GHz and 6.05 GHz, respectively. The simulated resonance frequencies are 4.97 GHz and 5.98 GHz. The simulated results and the analyzed results agree well with each other. Thus, this method can be used to analyze the resonance

frequencies of the antenna with periodic structure based radiator.

2.3 Using Substrate Integrated Waveguide to Reduce Loss in Millimeter Wave Band

With the development of 5G communication techniques, the millimeter wave antennas are developed rapidly. High realized gain is one of the most important requirements because of the high communication loss in the air. In order to achieve high gain, the antenna array is proposed in the millimeter wave communication system. It is well known that the main loss in a large antenna array is mainly at the feed network. Thus, it is necessary to analyze the performances of the widely used transmission methods to decrease the losses in the feed network of the antenna array.

The rectangular waveguide, coaxial line and microstrip line are the most widely used EM transmission methods. Their geometries are shown in Fig. 2.15. These transmission methods are fully analyzed in [31]. The total attenuation constant of a transmission line or waveguide is

$$\alpha = \alpha_c + \alpha_d \quad (2.20)$$

where α_c is the attenuation constant due to the conductor loss. α_d is the attenuation constant due to the dielectric loss.

For TE waves or TM waves, the attenuation constant due to dielectric loss is given by (2.21).

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Figure 2.15: Geometries of transmission lines and waveguide. (a) rectangular waveguide. (b) coaxial line. (c) microstrip line.

$$\alpha_{d_{TE,TM}} = \frac{k^2 \tan \delta}{2\beta} N p/m \quad (2.21)$$

where $\tan \delta$ is the loss tangent of the dielectric material. β is the phase constant. k is the wave number, and it can be presented by (2.23). For TEM waves, $\beta = k$. The attenuation constant of TEM waves due to dielectric loss is given by (2.22) [31].

$$\alpha_{d_{TEM}} = \frac{k \tan \delta}{2} N p/m \quad (2.22)$$

$$k = \omega \sqrt{\mu_0 \epsilon_0 \epsilon_r} \quad (2.23)$$

The rectangular waveguide, coaxial line, and microstrip line are compared in Table. 2.2 [31]. The loss in the rectangular waveguide is the smallest, so the rectangular waveguide will be chosen to design the feed network of the millimeter wave antennas. However, the rectangular waveguide is hard to fabricate and

2.3. USING SUBSTRATE INTEGRATED WAVEGUIDE TO REDUCE LOSS IN MILLIMETER WAVE BAND

integrate with the antenna. The similar substrate integrated waveguide (SIW) is used instead of the normal rectangular waveguide because of the ease of fabrication. In the antenna array, SIW will be used in the feed network. It is helpful to understand how to design the SIW and how the energy can be coupled from (or to) the SIW when design the antenna array with SIW feed network. The necessary techniques will be introduced in Section 2.3.1 and Section 2.3.2.

Table 2.2: Comparison of rectangular waveguide, coaxial line, and microstrip line.

Characteristic		Rectangular Waveguide	Coaxial Line	Microstrip Line
Modes	Preferred	TE_{10}	TEM	$Quasi - TEM$
	Other	TM, TE	TM, TE	Hybrid TM, TE
Dispersion		Medium	None	Low
Bandwidth		Low	High	High
Loss		Low	Medium	High
Power Capacity		High	Medium	Low
Physical Size		Large	Large	Small
Ease of Fabrication		Medium	Medium	Easy
Integration with		Hard	Hard	Easy

2.3.1 Dimension of SIW

The dimension of traditional rectangular waveguide (as shown in Fig. 2.15 (a)) is easy to design. The cutoff wave number for TE_{mn} mode is as shown in (2.24).

$$k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2} \quad (2.24)$$

where a is the width of the waveguide. b is the height of the waveguide.

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The dominant mode of rectangular waveguide is TE_{10} mode. By setting $m = 1$ and $n = 0$ in (2.24), the dominant frequency is given as in (2.25). Thus, the width of the normal rectangular waveguide is presented by (2.26).

$$(f_c)_{10} = \frac{c}{2a\sqrt{\mu\epsilon}} \quad (2.25)$$

$$a = \frac{c}{2f_c\sqrt{\mu\epsilon}} \quad (2.26)$$

where μ and ϵ are the permeability and permittivity of the filling material in the waveguide.

The substrate integrated waveguide (SIW) is a synthetic rectangular waveguide. It is known to have similar guided wave and mode characteristics to the conventional rectangular waveguide. The SIW uses post walls instead of two sides of the rectangular waveguide, which makes it easy to fabricate. The geometry of SIW is shown in Fig. 2.16. The separated distance of the two post walls can be calculated by (2.27).

Figure 2.16: Geometry of substrate integrated waveguide.

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$$a_{SIW} = a + \frac{d^2}{0.95p} \quad (2.27)$$

where a is the width of the normal rectangular waveguide, and it can be derived from (2.26). d is the diameter of the vias, and p is the period of the vias. The diameter (d) and period (p) can be adjusted depending on the existing moulds in the fabrication process. They should satisfy the following conditions shown in (2.28) and (2.29) [32,33].

$$d < \frac{\lambda_g}{5} \quad (2.28)$$

$$p < 2d \quad (2.29)$$

where λ_g is the guided wavelength, and it can be calculated by (2.30).

$$\lambda_g = \frac{2\pi}{\sqrt{\frac{\epsilon_r(2\pi f)^2}{c^2} - \left(\frac{\pi}{a}\right)^2}} \quad (2.30)$$

2.3.2 Transition between Regular Waveguide and SIW

The substrate integrated waveguide has many advantages, such as low loss, low radiation, easy fabrication, and easy integration. However, the traditional waveguide is still widely used in the standard flanges. In order to connect the SIW fed antennas to the devices with standard flanges, it is necessary to transfer SIW to waveguide.

There are many kinds of transitions. One transition to realize the impedance

2.3. USING SUBSTRATE INTEGRATED WAVEGUIDE TO REDUCE LOSS IN MILLIMETER WAVE BAND

match and waveform transformation is implemented by using tapered structures [34, 35]. This kind of transition is broad band, but the structure is complex and big in size. It is not suitable to connect the antenna. Another kind of transition uses resonant structures to match the impedance and transfer the waveform [36, 37]. This kind of transition is compact, and easy to integrate with microstrip antennas.

The slotted transition is one of the simplest transitions which are used to transfer the rectangular waveguide and SIW. In [36], Li *et al.* proposed the slotted waveguide transition. The structure of the transition is shown in Fig. 2.17 [36]. As shown in Fig. 2.17, the standard flange can be easily mounted at the surface of the SIW, and there is no additional layers between SIW and the flange.

Figure 2.17: Geometries of one slot waveguide transition.

The EM wave is coupled between the traditional rectangular waveguide and

2.3. USING SUBSTRATE INTEGRATED WAVEGUIDE TO REDUCE LOSS IN MILLIMETER WAVE BAND

the substrate integrated waveguide through the slot in the transition. The length of the slot is half wavelength($\lambda_g/2$), and it is a little away from the center line of the SIW. As the EM wave in the slave waveguide is excited by the slot, the design of the slot is similar to the design of a slot antenna [38]. In Fig. 2.17, the SIW is shorted at the end. The center of the slot should be $3\lambda_g/4$ or odd times of quarter wavelength from the end of the SIW, so that it works as an open circuit at the center of the slot. The slot is required to be placed at the center of the rectangular waveguide to get the maximum power from it. Above all, the slot of the transition should satisfy the following three rules.

- The length of slot is half wavelength.
- The center of the slot is $3\lambda_g/4$ (or odd multiples of quarter wavelength) away from the end of SIW.
- The slot is at the center of the rectangular waveguide.

The theory of the slotted transition is analyzed in [36]. The coupled E field in the rectangular waveguide is shown in Fig. 2.18. From Fig. 2.18, it can be seen that the transition is excited from the SIW. The electromagnetic wave in the rectangular waveguide is coupled from the SIW through the slot. The transition is reciprocal, so the electromagnetic wave can also be coupled from the rectangular waveguide to the SIW.

Figure 2.18: Coupled E field in the rectangular waveguide.

2.4 Summary

In this chapter, the limitations of antenna performances were analyzed in various frequency bands. In order to improve the performances, some useful techniques and methods were reviewed. Section [2.1](#) introduced the method to use commercial ferrites to improve the performances of HF antenna. The bandwidth of the low-profile antenna can be increased by laying the ferrite tiles over the entire ground. The radiation performance can be improved by laying ferrite bars near the antenna in the regions with the high magnetic field. In section [2.2](#) the artificial material was presented. It is effective and efficient to improve the antenna performances in the microwave band by using the artificial materials. In order to understand the artificial material better, section [2.2.1](#) showed how to evaluate the effective EM properties of periodic structure, and section [2.2.2](#)

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described how EM wave travels in the artificial material. In section [2.3](#) the SIW was chosen as the transmission method in feed network of millimeter wave antenna array because of its low loss, high efficiency and ease of fabrication. In this section, it also presented how to design the SIW, and how to transit the electromagnetic waves between the regular waveguide and SIW.

Chapter 3

Ferrite Coating Design for Near-Ground HF Bowtie Antenna

3.1 Introduction

HF antennas are widely used in medium-to-long-range communications, e.g, aviation, maritime, military, commercial communication systems, etc. [39]. In these applications, HF antennas are usually horizontally polarized, passive and broadband. The horizontally polarized dipole antenna is designed to be placed at least a quarter of a wavelength ($\lambda/4$) over the ground [40]. This is because when the antenna is placed horizontally near the ground, an image current is formed over the ground plane. This image current destructs the radiation from the current along the antenna. Moreover, the input impedance of the antenna is decreased significantly because of the mutual coupling between the antenna

3.1. INTRODUCTION

and the ground. At the low HF frequency of 2 MHz, the antenna has to be placed as high as 37.5 m above the ground [41].

In many countries, due to the huge size and large height of HF antennas, large land space is dedicated to such antennas. However, in a small country like Singapore, land is scarce, and it is critical to minimize the volume space occupied by such antennas. Many researchers have attempted to solve this challenging problem by designing low-profile antennas [42–44]. One of these methods is to use Frequency Selective Surface (FSS) or High-Impedance Surface (HIS) [45–47]. FSS and HIS are both used to provide constructive image currents, which can then improve the radiation efficiency. However, the antenna still has to be placed high enough above the ground, which is about 10-20 m for a 2 MHz antenna. Therefore, the large height and narrow-band effect of FSS and HIS structures limit their usage at high frequency broadband application. In our previous work, we used commercial ferrite tiles to implement a low profile HF antennas [3, 48]. It was shown in [3, 48] that the ferrite tiles are able to mitigate the radiation deterioration caused by image currents. By coating the entire ground plane with ferrite tiles, the height of the antenna can be reduced to 0.05λ above the ground. Due to the large size of the HF antenna, the amount of ferrite tiles coating the entire ground plane is huge indicating that the cost of using these ferrite tiles is high. Each tile ($10\text{ cm} \times 10\text{ cm} \times 0.67\text{ cm}$) costs about 5 USD. Furthermore, the weight loading on the ground plane is also large, since each tile weighs 0.34 kg. This makes it not practical for actual implementation.

3.2. PERFORMANCE ENHANCEMENT

In this chapter, we propose to use patterned ferrite tiles efficiently and effectively to implement the HF low-profile broadband antenna. Since the current in the ground plane indicates the strength of the image current, an economical and effective method to reduce this image current is to lay ferrite tiles only in areas where the current density is high. In order to reduce the current density, it is effective and necessary to place multi-layers of ferrite tiles in the certain area where the current density is relatively high. The realized gain has decreased due to the reduction of the reflected waves from the ground, but it is still larger than the gain reported in [3]. Moreover, the lossy ferrites can be used as loading instead of traditional lumped loading because of its broadband characteristics. By loading the ferrite at the end of antenna, the electrical length of the antenna is increased, and the reflection from the end of the antenna is also minimized. The operating frequency band of the HF antenna is extended to the lower frequency, and the input impedance of the antenna becomes flat.

3.2 Performance Enhancement

3.2.1 Bandwidth Improvement at Desired Frequencies

3.2.1.1 Image Theory

Fig. 3.1 (a) shows a horizontal dipole antenna operating at 10 MHz placed 3 meter(0.1λ) above the ground plane. The image current density due to the near ground dipole is also plotted. It can be seen that the maximum current density is 0.016A/m, which occurs right below the dipole antenna centre. In order to

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remove the image currents, the ferrite tiles are laid on the entire ground plane. The used ferrite material is SN-20 [18]. Each ferrite tile is $10\text{cm} \times 10\text{cm} \times 0.67\text{cm}$. Fig. 3.1 (b) plots the current density on the ground plane after the entire ferrite tiles are laid. The maximum image current without any ferrites was about 0.016 A/m as shown in Fig. 3.1 (a), while it is only a maximum of about 0.008 A/m after laying the ferrite tiles as shown in Fig. 3.1 (b). Compared to that without ferrites, both the density and the scale of image currents with ferrites are much smaller. Thus, laying ferrite tiles on the ground plane is effective to reduce the image current and therefore improve the performance of the antenna.

Figure 3.1: The dipole antenna and the image current. (a) is the case without ferrites, and (b) shows the current with ferrite tiles on the entire ground plane.

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3.2.1.2 Optimization of Area of Ferrites

A simple HF bowtie antenna is placed 1 meter above a conducting ground plane. The dimensions of the antenna are shown in Fig. 3.2. The length and width of the antenna are 14.27 meter and 4.5 meter, respectively. At the feeding point, a Balun is used to balance the signal feed and ensure broadband matching of the feed point. The antenna is folded slightly upwards such that the two sides of the bowtie is at 3.5 meters above the ground as seen in Fig. 3.2(b). This is done to improve the performance of the antenna and for ease of installation.

Figure 3.2: Geometry and parameters of the HF bowtie antenna. (a) top view. (b) side view.

The simulated VSWR is shown in Fig. 3.3, and the red solid line is the VSWR for the case over PEC plane. The value of the VSWR is as high as 18.4 at 10 MHz. In order to improve the performance of this antenna at 10 MHz,

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the current distribution on the ground plane at 10MHz is plotted in Fig. 3.4 (a). The maximum electric current occurs in the center of the structure. The area with high current density is approximately 4 meters in width. Therefore, a rectangular ferrite block with the size of 4 meter $\times$ 6 meter is placed in the center of the conducting ground to reduce the electric current. The current distribution on the ground plane after the ferrite block used is as shown in Fig. 3.4 (b). As can be seen, the electric current density is much reduced with the help of ferrites. With the decrease in current density, the VSWR performance of the antenna is improved. The antenna with ferrite tile coating over the entire ground plane is also simulated for comparison. The area covered with ferrite tiles is 16 meter in length and 7 meter in width. From Fig. 3.3, it can be seen that the VSWR with the rectangular shaped ferrite block is a little higher than the VSWR with ferrite tiles over the entire ground plane. This is as expected since the current density is reduced over the entire ground plane using the ferrite tiles. However, comparing the rectangular shaped ferrite tiles and the entire ground with ferrite tiles, the performance is comparable. And yet, the amount of ferrite tiles required is reduced by 78.6%. Thus, this method is found to be an effective and economical way of using ferrite tiles for antenna performance improvement.

3.2.1.3 Bandwidth Improvemnt at 15 MHz -25 MHz

In this section, an HF antenna is improved in the range of 15MHz-25MHz by laying ferrites to reduce the maximum current density induced on the ground

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Figure 3.3: Simulated VSWR results of the antenna with various loadings.

Figure 3.4: Current distribution on the ground of the antenna with and without ferrites at 10 MHz. (a) without ferrites. (b) with ferrites.

plane. In order to improve the performance, the antenna is simulated and the electric current distributions are plotted from 15MHz to 25MHz. In this frequency range, the intersection of the place, where the maximum currents occur, is under the center of the antenna. Thus, we pave the center with ferrites. After optimization, the size of the ferrite block is selected as 0.8 meter $\times$ 5 meter. Fig. [3.5](#) shows the optimized HF antenna.

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Figure 3.5: Photo of fabricated antenna.

The VSWR is measured using a vector network analyzer. The measured results are shown in Fig. 3.6. The blue dash line is the VSWR without any ferrites, and the solid red line shows the VSWR with ferrites. The VSWR of the original antenna is relatively higher in the range of 16-18MHz and 19.5-21.5MHz. With the ferrites laying under the center of the antenna, VSWR gets lower. From Fig. 3.6, it is easy to see that there is a visible improvement. The voltage standing wave ratio is almost under three from 15MHz to 25MHz.

3.2.2 Low Profile Implementation

The antenna height should be designed very carefully, because the ground plane plays an important role in the operation of antennas. The effect of ground cannot be ignored especially for horizontally polarized antennas. In terms of near field, the height of antenna affects the input impedance because of the mutual coupling between the antenna and the ground. When the horizontally polarized antenna is near the ground, the input impedance of the antenna

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Figure 3.6: Measured VSWR results of the antenna with and without ferrites.

decreases significantly. In terms of far field, when the antenna height is low, the radiation from the image current destructively interferes with the radiation from the antenna current.

Due to the large size and wavelength, it is a challenge to place the HF antenna near the ground. In [3], commercial ferrite tiles are used to build the low-profile HF antenna. By covering the ground with a large number of ferrite tiles, the effect of the image current is reduced. However, the cost and weight of ferrite load on the ground plane are very high. In order to use the ferrite more effectively, we propose to lay the ferrite tiles only in areas that are affected greatly by the image current. The current in the ground plane can represent the strength of the image current. Through analyzing the image current, we propose to lay ferrite tiles in the area with high current.

We use the half wavelength dipole as an example. The relative dielectric

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constant (ϵ_r) of the commercial ferrite tile is 12, and the electric loss tangent can be ignored. The relative magnetic permeability (μ_r) of the ferrite used is $500 - j220$ at 10 MHz [18]. The dipole antenna operates at 10 MHz. When the antenna is placed $\lambda/4$ space (7.5 m) above the ground, its impedance is 73Ω . When the height of the same antenna is reduced to 1.5 m as shown in Fig. 3.7 (a), the resistance of the antenna reduces significantly to 5.02Ω as shown in Fig. 3.8. This is due to the mutual coupling between the antenna and the ground plane.

Figure 3.7: The current distribution in the ground. (a) The current in the ground of the antenna without ferrite tiles. (b) The current in the ground of the antenna with 1 layer of ferrite tiles. (c) The current in the ground of the antenna with three layers of ferrite tiles.

The current density distribution on the ground is plotted and examined in Fig. 3.7. The maximum current density is 2.39 A/m formed under the center of the dipole. The current is strong over the ground mainly along the dipole, and becomes weaker when moving away from the center point of the dipole.

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(a)

(b)

Figure 3.8: Input impedance of the antennas with various antenna heights and patterned ferrites. (a) real part of input impedance. (b) imaginary part of input impedance

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In order to reduce this image current density, ferrite tiles are laid on areas under the dipole where current density is high. With one layer of ferrite laid, as shown in Fig. 3.7 (b), it can be seen that the current density under the dipole is significantly reduced. The maximum current density occurs along the edges of the ferrite block, and is around 0.15 A/m. The current density under the dipole is reduced to 0.027 A/m. Thus, the input impedance is increased to 46.61 Ω with the ferrite loading, as shown in Fig. 3.8. In Fig. 3.7 (b), although the image current density is reduced with one layer of ferrite, the current density over the ground along the dipole is still strong. In order to reduce the current density further and distribute the current uniformly, 3 layers of ferrite tiles are placed under the center of the dipole as shown in Fig. 3.7 (c). By using the multi-layer patterned ferrite, Fig. 3.7 (c) shows a relatively uniform distributed current over the ground plane. The maximum current density is further reduced to 0.10 A/m, and the current density under the center of the dipole is reduced to 0.018 A/m. As shown in Fig. 3.8, with the multi-layer ferrite, the input impedance of the antenna can be further improved to 50.47 Ω .

Due to the placement of lossy ferrite tiles, the radiation patterns are changed, and its realized gain is also decreased. In Fig. 3.9, the antenna with ferrite loading provides less radiated power than the antenna of a height of $\lambda/4$, but more than the same antenna at 1.5 m without ferrite loading.

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Figure 3.9: The radiation patterns of antennas with various antenna heights and patterned ferrites.

3.2.3 Bandwidth Enhancement

For the design of a passive broadband HF antenna, one of the challenges is to extend the lower bound of the bandwidth without increasing the physical size of the antenna. In this section, a method to decrease the antenna size by reducing the resonant frequency is explored. There are many methods to reduce the size of the antenna, such as the meandered line antenna [49], loop antenna [50], loaded antenna [51], etc. Considering the easy fabrication and implementation, the end loaded antenna is chosen. For end loaded antennas, lumped components can usually be used. However, due to the narrow band nature of the end loading with lumped component, the ferrite tiles are used for broadband loading. One ferrite block is located at the end of each dipole arm, as shown in Fig. 3.10. The same dipole and the same ferrite tiles are used

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here. The folded vertical wires increase the physical length of the antenna to broaden the bandwidth, while the ferrite loadings increase the bandwidth of the antenna further. Fig. 3.12 shows the input impedance of the antenna. It can be seen that the zero-reactance frequency is reduced from 10 MHz to 6.5 MHz. In order to get the same zero-reactance frequency, a series capacitor and inductor is connected to the end of each arm of the dipole. Comparing the impedance of the lumped loaded antenna in Fig. 3.12 the impedance of ferrite loaded antenna shows a flat response. It is easier to achieve broadband by using ferrite loadings, as seen from the simulated VSWR results in Fig. 3.11. Comparing with the antenna without loadings, the bandwidth of the ferrite loaded antenna is much broader.

Figure 3.10: Side view of the end loaded dipole antenna.

The realized gain of the ferrite loaded antenna is low because of the low efficiency, which is like any end loaded antenna [52]. The elevation radiation patterns are shown in Fig. 3.13. The reduction of realized gain is as expected of an end loaded antenna.

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

Figure 3.11: Simulated VSWR results of the antennas with and without ferrite loadings.

3.3 Implementation and Fabrication of Near-Ground HF Antenna

A near-ground broadband HF antenna is implemented and tested. The antenna is built and mounted on the zinc covered rooftop of one 7-story building. There is one piece of flat roof, and the antenna is designed to be mounted on the roof. The antenna comprises of several wires. The length, width, and height of the antenna are 35.0 m, 13.4 m, and 3.5 m, respectively. The antenna structure is suspended on both sides at a height of 3.5 m, and slopes down at the center to a height of 1 m. At the feeding point, a Balun is used to facilitate the measurement using coaxial cable. The layouts of the antenna and the building are shown in Fig. 3.14.

First, we examine the use of ferrite tiles to reduce the effect of image current.

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

(a)

(b)

Figure 3.12: Input impedance of the antenna with and without ferrite loadings. (a) real part of input impedance. (b) imaginary part of input impedance

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

Figure 3.13: The elevation radiation patterns of the antennas with and without ferrite loadings.

As seen in Fig. 3.15, the VSWR of the original antenna without ferrites is below 3 from 7 MHz to 30 MHz, but there is a peak around 12.5 MHz. In order to reduce the VSWR at 12.5 MHz, the current on the rooftop is plotted in Fig. 3.16 (a), and the image current is analyzed. The maximum current occurs under the center of the antenna. By using the proposed method, the ferrite tiles are placed under the center of the antenna as shown in Fig. 3.17. In Fig. 3.16 (b), with the ferrite tiles, current is redistributed and becomes more uniform. As a result, the VSWR (with central ferrite) is improved around 12.5 MHz as shown in Fig. 3.15. The realized gain is reduced slightly due to the absorption by the ferrite. It is seen that the effect of the image current can be reduced effectively, and the VSWR can be improved by laying the ferrite tiles only in the region with high current density.

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

Figure 3.14: Geometry of the HF antenna. (a) 3D-view of the antenna, (b) Side view, (c) Top view.

Second, the extension of the antenna bandwidth is examined. At the end of the two arms, two loading lines are added. Two ferrite blocks are located at the end of the loading lines as shown in Fig. [3.17](#). The simulated VSWR results

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

(a)

(b)

Figure 3.15: Simulated VSWR and maximum realized gain results of the antenna with and without patterned ferrite. (a) Comparison of VSWR. (b) Comparison of realized gain.

(with central and loaded ferrite) are shown in Fig. 3.15. By using the ferrite end loadings, the bandwidth of the antenna is extended from 7-30 MHz to 2-

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

Figure 3.16: The current distribution on the rooftop at 12.5 MHz. (a) without ferrite tiles, (b) with ferrite tiles.

Figure 3.17: The layout of the patterned ferrite.

30 MHz. Comparing the realized gain of the antenna without ferrite loadings, the gain of the antenna with ferrite loadings is almost unchanged, as shown in Fig. 3.15. In the lower frequency range, although the reflection coefficient is improved, the realized gain is not increased. The ferrite loadings may produce unexpected effects because of the complicated structure of the antenna. By

3.3. IMPLEMENTATION AND FABRICATION OF NEAR-GROUND HF ANTENNA

adding the ferrite loadings, the current along the antenna is redistributed, which may reduce the radiation efficiency.

Finally, the HF antenna with ferrite loading is fabricated. The patterned ferrite is shown in Fig. 3.17 and the photos of the central ferrite and the end ferrite loadings are shown in Fig. 3.18. The measured VSWR is shown in Fig. 3.19. With the patterned ferrite, the VSWR is decreased from 7.00 to 3.29 at 2 MHz, and the VSWR is also improved around 12.5 MHz. The bandwidth of the antenna is improved by using the ferrite tiles, which is expected from the simulated results. Comparing the VSWR shown in Fig. 3.15 and Fig. 3.19, the trends of the measured results are similar to the simulated ones. The differences between them are due to the differences in the material properties and the surrounding environment between the simulated model and the actual measurement site.

Figure 3.18: Photos of the ferrite layout.

The radiation pattern of the antenna is measured in two planes. They are in the $y - z$ plane indicated in Fig. 3.14 (a) and the azimuth plane at 7000 ft height. The measured results are normalized, and the shapes of the radiation patterns are compared with the simulated ones in Fig. 3.20. There are differences between the simulated and measured results because of the non-

3.4. CURRENT DIRECTION ADJUSTMENT

Figure 3.19: Measured VSWR results of the antenna with and without patterned ferrite.

ideal measurement conditions.

By using the patterned ferrite, the bandwidth of the low-profile antenna is broadened. The entire antenna is kept low profile, and can be implemented on a rooftop. Simulated and measured results agree well with each other, and this verifies the proposed technique of the effective use of patterned ferrite and ferrite loading for the implementation of a near-ground broadband HF antenna.

3.4 Current Direction Adjustment

By loading the ferrite, the voltage standing wave ratio at 2 MHz is reduced significantly. However, the realized gain is deteriorated under this circumstance. The use of ferrite can improve the VSWR, but it may reduce the antenna efficiency. For the antenna loaded with ferrites, the directions of the currents

3.4. CURRENT DIRECTION ADJUSTMENT

Figure 3.20: Measured and simulated radiation patterns of the antenna with ferrite tiles. (a) radiation patterns at 8.4 MHz in the y-z plane, (b) radiation patterns at 8.4 MHz at 7000 ft, (c) radiation patterns at 18.2 MHz in the y-z plane, (d) radiation patterns at 18.2 MHz at 7000 ft.

along some wires are in the opposite direction to others, which makes the realized gain reduced. The half of the antenna with the current vectors at 2 MHz are shown in Fig. [3.21](#), and the opposite currents are circled. The radiation efficiency is deteriorated by these contrasting currents.

In order to circumvent these unexpected currents with opposite directions,

Figure 3.21: The current along the antenna with loaded ferrites at 2 MHz.

two loading wires are added. These two wires are used to connect the antenna and the building. It can be observed that, with the additional loadings, the directions of some currents are fixed. The redistributed currents along the wires of the antenna are shown in Fig. 3.22. Comparing with Fig. 3.22, there are more currents pointing in the same direction. As a result, the realized gain is improved from -18.41 dB to -9.97 dB, which is shown in Fig. 3.23. Moreover, the VSWR remains good, as shown in Fig. 3.24.

3.5 Summary

A method to implement a near-ground broadband HF antenna has been proposed in this chapter. The impedance bandwidth can be improved at the desired

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Figure 3.22: The current along the antenna with two additional loading wires at 2 MHz.

Figure 3.23: Simulated realized gain of the antenna without ferrite loadings, with ferrite loadings, and with ferrite and 2 loading wires.

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Figure 3.24: Simulated VSWR results of the antenna without ferrite loadings, with ferrite loadings, and with ferrite and 2 loading wires.

frequencies by laying ferrite tiles in the region where there is a large amount of current in the ground, and the lower bound of the bandwidth can be extended by loading the antenna with ferrite. The bandwidth broadening does not affect the realized gain much. A near-ground broadband HF antenna has been implemented and fabricated using the proposed method. With the ferrite loading in a proper pattern, the voltage standing wave ratio of an HF antenna has been improved around 12.5 MHz, and the lower bound of its bandwidth is decreased from 7 MHz to 2 MHz. The agreement between simulated and measured results verifies the proposed method.

Chapter 4

Design of a Compact Patch

Antenna Loading Periodic

Jerusalem Crosses

4.1 Introduction

Microstrip antenna is widely used because of its ease of fabrication and low cost. However, the microstrip antenna has some shortcomings caused by the substrate material. The high dielectric constant and small thickness of the substrate results in an antenna that is low gain and narrow bandwidth.

In recent years, the artificial materials have received considerable attention. This is due to the special properties of the material, which have not yet been found in nature. The artificial materials are able to improve the antenna perfor-

4.1. INTRODUCTION

mances through the control of electromagnetic wave propagation. The artificial materials are able to address some of the shortcomings of the conventional microstrip antenna.

The electromagnetic bandgap (EBG) is one type of the artificial materials based on periodic structure. It can prevent the electromagnetic wave propagation over a certain frequency range, therefore, it is often arranged around the antenna to prevent the surface wave. In such a way, the back lobe of microstrip antenna can be reduced. EBG can also be used to decrease the coupling between antennas in an antenna array, which can be achieved by inserting EBG structures between the sub-antenna elements of the array [53]. Frequency selective surface (FSS) is another type of periodic structure based materials. Working as a PEC in the operating frequency and presenting the bandpass features out of the operating frequency, FSS is able to reduce the backward scattering [54–56].

Nowadays, the periodic structure based material is more and more frequently used in the microstrip antenna designs to replace the traditional substrate materials [57–60]. This material can be used to improve the operating performances of antenna, such as its radiation efficiency, gain, and impedance bandwidth.

The miniaturization of antennas, similarly, can be achieved by using the periodic structures. In [57], by placing a complementary split-ring resonators (CSRR) between the patch and the ground of the antenna, the size of the patch is significantly decreased. However, the thickness of the antenna is increased. this substrate is two layers of material, and it is difficult to fabricate. The

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radiation efficiency is reduced, and what's worse is this design makes the already narrow bandwidth even narrower. In [58], a microstrip antenna is miniaturized by loading a single-layer artificial material in which four U-shaped structures are etched under the patch. This design improves both the radiation efficiency and the bandwidth, but it is difficult to fabricate because of the vertical vias. In [61], a simple single layer patch antenna with embedded meander-line (EML) loading is proposed. This design is easy to fabricate, can reduce the patch size by about 35%, and broadens the bandwidth by 1.85 times compared to the conventional patch antenna. However, the etched EMLs allow energy leakage to the back, and thus, the main lobe and gain are reduced. In order to increase the gain, a metal plane reflector is placed beneath and in parallel to the antenna ground plane. However, in order for the reflector to be effective, the distance between the reflector and the antenna ground has to be as large as 5 mm ($0.058\lambda_0$). Therefore, even though the patch size is reduced, the overall dimension of the antenna still remains large, because the total thickness of the antenna is 4.2 times of the original. In [60], a layer of JC-FSS is inserted between the patch and the ground plane of the antenna. The bandwidth of the antenna is extended to 10.44%, but this design leads to the same problem as in [57]. The thickness of the antenna is as large as $0.092\lambda_0$.

In this chapter, we propose to etch the periodic Jerusalem crosses in the antenna ground plane in order to achieve antenna size reduction and bandwidth improvement. Besides performance improvement, this structure is also simple to implement and easy to fabricate. For fabrication purpose, the low-cost FR4 is

4.2. EFFECTVIE ϵ AND μ OF SUBSTRATE

chosen as the substrate. Although the loss tangent of FR4 is high, the radiation efficiency is improved significantly compared to the conventional patch antenna. The improvement of maximum realized gain is small, because of energy leakage through the defects in the ground plane as explained in [61].

4.2 Effectvie ϵ and μ of Substrate

When the periodic structures are etched in the ground, the EM properties of the substrate, e.g., permittivity and permeability, are not the same as FR4 any more. In general, the complex permittivity and permeability of a passive medium are

$$\epsilon = \epsilon_r \epsilon_0 = (\epsilon'_r - j\epsilon''_r) \epsilon_0 \quad (4.1)$$

and

$$\mu = \mu_r \mu_0 = (\mu'_r - j\mu''_r) \mu_0 \quad (4.2)$$

where $\epsilon_0 = 8.85 \times 10^{-12} F/m$ and $\mu_0 = 4\pi \times 10^{-7} H/m$. The wave number and the wave impedance are

$$k = \omega \sqrt{\epsilon_0 \epsilon_r} \sqrt{\mu_0 \mu_r} \quad (4.3)$$

and

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$$\eta = \eta_0 \frac{\sqrt{\mu_r}}{\sqrt{\epsilon_r}} \quad (4.4)$$

where η_0 is the wave impedance in free space. $\eta_0 = 120\pi \Omega$. The index of the refraction is

$$n = \sqrt{\epsilon_r} \sqrt{\mu_r} \quad (4.5)$$

It is easy to calculate and measure the wave amplitude of transmission and reflection wave in free space. In addition with (4.1)-(4.5), it is easy to get the effective permittivity and permeability. In Fig. 4.1, it shows uniform plane waves travelling through a slab. The thickness of the slab is a . The permittivity and the permeability are ϵ_r and μ_r , respectively. The wave impedance in the slab is η_1 , and in free space, it is η_0 .

Figure 4.1: EM wave travels trough a slab of material.

In [25], the author, Cotuk, gave the general form of the wave matrix in a

4.2. EFFECTIVE ϵ AND μ OF SUBSTRATE

three-layer of medium, as shown in (4.6).

$$\begin{bmatrix} c_1 \\ b_1 \end{bmatrix} = \prod_{n=1}^2 \frac{1}{T_n} \begin{pmatrix} e^{j\phi_n} & R_n e^{-j\phi_n} \\ R_n e^{j\phi_n} & e^{-j\phi_n} \end{pmatrix} \begin{bmatrix} c_3 \\ b_3 \end{bmatrix} \equiv \begin{pmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{pmatrix} \begin{bmatrix} c_3 \\ b_3 \end{bmatrix} \quad (4.6)$$

where ϕ_n is the phase shift in layer n . When the thickness of the third layer extends to infinite, b_3 can be treated as zero ($b_3 = 0$). The transmission coefficient and reflection coefficient are (4.7) and (4.8), respectively.

$$T = \frac{c_3}{c_1} = \frac{1}{a_{11}} \quad (4.7)$$

$$R = \frac{b_1}{c_1} = \frac{a_{21}}{a_{11}} \quad (4.8)$$

In [26], Chen *et al.* used the reflection coefficient and transmission coefficient to define the S parameters as

$$S_{11} = R \quad (4.9)$$

and

$$S_{21} = T e^{jk_0 a} \quad (4.10)$$

where k_0 is the wave number in free space.

The impedance and refractive index can be presented in terms of S-parameters

4.2. EFFECTIVE ϵ AND μ OF SUBSTRATE

by (4.11) and (4.12), respectively.

$$Z = \pm \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}} \quad (4.11)$$

$$n = \frac{1}{k_0 t} \{Im[\ln(e^{ink_0 a})] + 2m\pi - jRe[\ln(e^{ink_0 a})]\} \quad (4.12)$$

$$e^{ink_0 a} = \frac{S_{21}}{1 - S_{11} \frac{Z-1}{Z+1}} \quad (4.13)$$

It should be noted that the sign of Z should be determined carefully to ensure $Re(Z) \geq 0$ and $Im(n) \geq 0$, or equivalently, $|e^{ink_0 a}| \leq 1$. In (4.12), m is an integer related to the real part of the refractive index, which can be determined exactly by the method in [27].

Inverting (4.4) and (4.5), the effective permittivity and effective permeability can be obtained as following

$$\epsilon_{eff} = \frac{n}{Z} \quad (4.14)$$

and

$$\mu_{eff} = nZ \quad (4.15)$$

4.3 Substrate Design

The substrate is constituted by a piece of dielectric backed with a defect ground plane. In the ground plane, a 3 by 3 Jerusalem cross array is etched. The patterned ground plane and the unit cell are shown in Fig. 4.2. Each unit cell contains four small patches in the center, and four arms around the patches, which is shown in Fig. 4.2 (b). The parameter values of the unit cell are as follows, $a = 7mm$, $l = 5.5mm$, $w = 0.4mm$, and $g = 0.25mm$. In this design, the dielectric parameters of the FR4 are $\epsilon_r=4.3$, and $\tan\delta=0.025$. The thickness of the substrate is 0.8 mm.

Figure 4.2: (a) Pattern ground plane. (b) Unit cell of the etched uni-planar compact structure. $a = 7mm$, $l = 5.5mm$, $w = 0.4mm$, and $g = 0.25mm$.

The unit cell with Jerusalem cross can be analyzed using the equivalent circuit. The gaps between small patches provide the capacitance, and the arms can be seen as the microstrip inductors. The capacitors and inductors form an LC circuit, and the oscillation frequency affects the properties of this piece of material. The equivalent circuit of the unit cell is indicated in Fig. 4.3. The

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dispersion diagram can be obtained by simulating the unit cell. The supported wave in the periodic structure can also be analyzed from the dispersion diagram. The dispersion diagram is shown in Fig. 4.4, which shows the relation between the wavenumber and the frequency. The EM wave is prevented in the unit cell from 4.62 GHz to 6.15 GHz.

Figure 4.3: Equivalent circuit of unit cell with Jerusalem cross.

Figure 4.4: Dispersion diagram of unit cell.

If the electrically small patterned structures are arranged in the material periodically, the material can be taken as a uniform dielectric with a certain

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permittivity and permeability. The effective permittivity will get smaller because of the absence of metal due to the etched pattern in the ground plane. The permittivity and permeability of the material can be derived from the simulated transmission and reflection coefficients of one unit cell [62, 63]. The simulated unit cell and ports are indicated in Fig. 4.2 (b). This structure is simulated using CST Microwave Studio (a 3D EM simulation software), and the transmission and reflection coefficients are acquired. The effective permittivity and permeability of the substrate can be then obtained by (4.11) - (4.15).

The effective permittivity is shown in Fig. 4.5 (a). From 1 GHz to 4 GHz, the real part of ϵ_{meta} is greater than 0, and it is almost below the original permittivity ϵ_r of FR4 ($\epsilon_r = 4.3$) except at around 1.45 GHz. The low permittivity of the substrate is due to the leakage of electric field from the defects in the ground plane. The effective ϵ of the substrate varies with frequency. The effective permeability of the substrate is plotted in Fig. 4.5 (b). From 1 GHz to 3.0 GHz, the real part of μ_{meta} is larger than 1 (the relative permeability of FR4 and air). It falls to 0 when the frequency is larger than about 3.0 GHz. Above 3.0 GHz, the effective permeability is below 0, and the material performs like a negative permeability material. The imaginary parts of ϵ_{meta} and μ_{meta} indicate that there is leakage loss of the gap in the ground.

4.4 Antenna Design

A patch antenna operating at 2.4 GHz will be designed. The resonant frequency of the antenna is on the right-hand dispersion region of the unit cell. The metal

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Figure 4.5: (a) The permittivity of the substrate. (b) The permeability of the substrate.

patch is designed as a simple square. The length of the patch is the half of the wavelength in the substrate. The length of the patch can be calculated using (4.16) [64]. In Fig. 4.5, the effective permittivity and the effective permeability of the substrate at 2.4 GHz can be obtained to be 1.9 and 14.4, respectively. The imaginary parts are ignored, since they are quite small, which indicates that the loss is negligible. Even though the permittivity of the substrate is less than 4.3 (the original permittivity of FR4), the length of the patch is reduced due to the dominant increase in the permeability of the substrate, μ_{eff} . The

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practical length of the patch should be longer because of the fringing effect. Based on the theory and simulation, the length of the patch is determined to be 16.7 mm for an operating frequency of 2.4 GHz.

$$L = \frac{c}{2\pi f \sqrt{\epsilon_{eff} \mu_{eff}}} \quad (4.16)$$

Figure 4.6: The illustration of microstrip antenna.

In this design, the periodic pattern is arranged only beneath the patch to allow for ease of the impedance matching. A 3×3 array of unit cell patterns is etched in the ground plane. Each unit is as shown in Fig. 4.2 (b). There is energy leakage from the defects in the plane ground, so the back lobe is increased regardless of the size of the ground plane. Therefore, the ground plane size is kept at $22mm \times 40mm$. On the top side, a 50Ω microstrip insert feedline is used to feed the patch antenna and also to achieve the impedance matching. The insert depth is calculated to be 9 mm. The top and bottom views of the proposed antenna are as shown in Fig. 4.6.

4.5 Results and Discussion

4.5.1 Simulated Results

The proposed antenna is simulated by CST studio. A conventional patch antenna, which also operates at 2.4 GHz and uses the same FR4 PCB as the substrate, is simulated for comparison purposes. The simulated reflection coefficients of both proposed antenna and the conventional patch antenna are shown in Fig. 4.7. The solid line and dash dot line are the simulated reflection coefficients of the antennas with proposed substrate and conventional dielectric, respectively. The -10 dB impedance bandwidth of the antenna using the proposed substrate is about 7.1%, while the bandwidth of the conventional patch antenna is only 2.5%. The quality factor is proportional to the permittivity of the substrate, and is inversely proportional to the permeability as shown in (4.17).

$$Q = \frac{\pi w_p \sqrt{\epsilon_{eff}}}{2G_r h \eta_0 \sqrt{\mu_{eff}}} \quad (4.17)$$

where G_r is the radiation conductance. η_0 is the electromagnetic wave impedance in free space. Thus, by decreasing ϵ_{eff} and increasing μ_{eff} , the Q factor of the antenna becomes smaller, and the bandwidth gets broader.

The radiation patterns of the proposed antenna are plotted in Fig. 4.8. From the radiation patterns, it can be seen that the back lobe of the antenna is increased. Some energy is leaked to the back of the antenna through the slots in the ground plane. The directivity of this antenna gets smaller, and it looks like

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Figure 4.7: The simulated and measured return loss.

an omni-directional antenna. The dimension and radiation performances of the proposed antenna and the conventional patch antenna are listed in Table [4.1](#).

Figure 4.8: The simulated and measured radiation pattern. (a) E-plane (y-z plane). (b) H-plane (x-z plane).

In Table. [4.1](#), the radiation efficiency is as high as 88.5%, while the radiation

4.5. RESULTS AND DISCUSSION

Table 4.1: Comparison of the proposed antenna and the conventional patch antenna.

	Patch Size (mm^2)	Bandwidth	Radiation Efficiency
Proposed Antenna	16.7×16.7	7.1%	88.5%
Conventional Antenna	29.7×29.7	2.5%	46.3%

efficiency of the conventional patch antenna is only 46.3% because of the high tangent loss of the FR4 dielectric substrate. The radiation efficiency takes into consideration the reflection, conduction, and dielectric losses. The efficiency is defined as the ratio of the power radiating out to the total delivered power [64], so the efficiency is in (4.18).

$$e = \frac{R_r}{R_r + R_L} \quad (4.18)$$

where R_r is the radiation resistance, and it represents the radiation by the antenna. R_L is the conduction and dielectric losses. With periodic patterns etched in the ground plane, the size of the patch gets smaller, and the radiation resistance is increased. Thus, based on (4.18), the radiation efficiency of the antenna is improved.

4.5.2 Fabrication and Measured Results

The antenna with substrate is fabricated. Without any vertical vias and connectors, this one-layer antenna is very easy to fabricate and can be done using the common PCB milling machine. The photos of the fabricated antenna are

4.5. RESULTS AND DISCUSSION

shown in Fig. 4.9 (a). In order to visualise the significant difference in the dimensions of the conventional patch antenna and the proposed substrate patch antenna, these two antennas are placed together and compared in Fig. 4.9 (b). It is easy to see that the proposed antenna is able to achieve a significant reduction in the size of the antenna. The proposed compact antenna is also compared with some of the published compact antennas introduced in section 4.1. Their dimensions are listed in Table 4.2. Compared with three published small size antenna, the proposed antenna has the smallest dimension, and other performances of the proposed antenna are quite good.

Table 4.2: Comparison of the antenna using various unit cells.

Antenna	Unite Cell	Antenan Size	BW	Rad. Eff.	Realized Gain
Proposed	J-cross in ground	$0.17\lambda_0 \times 0.32\lambda_0 \times 0.006\lambda_0$	7.10%	88.5%	2.14 dBi
[58]	Inverted U shape	$0.17\lambda_0 \times 0.32\lambda_0 \times 0.030\lambda_0$	29.3%	85.5%	-0.80 dBi
[60]	Jerusalem cross	$1.23\lambda_0 \times 1.23\lambda_0 \times 0.092\lambda_0$	8.41%	85.0%	6.57 dBi
[61]	Meander line	$0.70\lambda_0 \times 0.89\lambda_0 \times 0.017\lambda_0$	2.22%	90.3%	7.00 dBi

The compact antenna is measured in an anechoic chamber. The measured S-parameter and radiation patterns are plotted and compared with the simulated ones in Fig. 4.7 and Fig. 4.8, respectively. The measured results and the simulated results agree well with each other.

4.5.3 Frequency Tuning

The flexibility of operating frequency tuning is one of the most important issues to determine the performance of an antenna [65]. In section 4.3, the unit cell

4.5. RESULTS AND DISCUSSION

(a)

(b)

Figure 4.9: The photos of microstrip antenna. (a) The top view and back view of proposed antenna. (b) The proposed antenna vs. the conventional patch antenna.

and the properties of substrate have been analyzed. In the proposed design, the length of the arm slots determine the inductance and the properties of the substrate. Thus, the resonance frequency can be tuned conveniently by

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changing the length of the arm slots, (l), as shown in Fig. 4.2(b). By varying the length of the slot, l , the effective permittivity and permeability of the substrate can be tuned. It is a common practice to adjust the size or the frequency of operation of the antenna by changing the effective permittivity or permeability of the substrate.

In this design, when l is increased from 0.5 mm to 5.5 mm in steps of 1 mm, while keeping the other parameters constant as those listed in Fig. 4.2, the resonance frequency can be tuned from 3.74 GHz to 2.4 GHz. The tuning frequency range is large. In Fig. 4.10, it can be seen that the impedance match is good, and the return loss is better than -17 dB for varying values of l even with the unchanged depth of the insert feed. The radiation patterns at several tuned frequencies are plotted in Fig. 4.11. In Fig. 4.11, the back lobe gets a little smaller with the decreasing of l , because less energy is leaked when the slot is reduced. The main lobes are stable at all the tuned frequencies.

Figure 4.10: The dependence of resonant frequencies on parameter l .

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Figure 4.11: The radiation patterns at tuned frequencies. (a) E-plane (y-z plane). (b) H-plane (x-z plane).

However, the physical dimension of the slot can be adjusted hardly without the re-fabrication. Therefore, we explore the possibility of using lump components for the tuning of the operation frequency of the antenna. This technique is also employed in the implementation of the reconfigurable frequency selective surfaces [66]. One of the simplest methods is to use tunable inductive elements (e.g., PIN diodes) to change the electromagnetic wave propagation properties in the periodic structures. In Fig. 4.12, the patterned ground plane with tunable inductors is shown. The inductor is arranged at the position shown in Fig. 4.12 (b). The tunable inductor is connected in parallel with the equivalent inductor provided by the connecting arm. When the tunable inductor is adjusted from 100 nH to 1 nH, the operating frequency of the antenna can be tuned from 2.33 GHz to 2.62 GHz. The tunable frequency range is not so large as the variable conductance range of the lumped elements, because the lumped elements cannot affect the electromagnetic wave propagation so much. The simulated results

4.5. RESULTS AND DISCUSSION

are shown in Fig. 4.13 and the impedance match is kept well. The radiation patterns in the tunable frequency range are also kept well, which is shown in Fig. 4.14.

Figure 4.12: The equivalent circuit of the tunable antenna. (a): the pattern in the ground plane. (b): one unit cell with diodes.

Figure 4.13: The simulated result of the tunable antenna.

Figure 4.14: The radiation patterns at tuned frequencies. (a) E-plane (y-z plane). (b) H-plane (x-z plane).

4.6 Summary

In this chapter, a compact microstrip antenna is designed. The patch of the antenna is stacked on a substrate with the patterned ground plane. The periodic structures are etched in the ground plane. The effective permittivity and permeability of this substrate can be obtained from the simulated reflection and transmission coefficients. With the defects in the ground plane, the effective permittivity, ϵ_{meta} , of the substrate gets smaller, and the effective permeability, μ_{meta} becomes much greater. Due to the small permittivity and large permeability change of the substrate, the size of the patch is decreased to 31.62%, and the impedance bandwidth is increased to 2.84 times of those of the conventional patch antenna. Even though the energy leakage through the defected ground plane is significant, the maximum realized gain is increased by 0.37 dB, and the radiation efficiency is improved from 39.5% to 88.5%. Furthermore, The

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operating frequency of this design can be tuned over a large frequency range, from 2.4 GHz to 3.74 GHz by adjusting the length of the slot of the periodic structure. If this antenna is to be used as a tunable antenna, an inductor element can be added across (in parallel with) the arm slot. With the tunable inductors, the antenna can be tuned from 2.33 GHz to 2.62 GHz. In this chapter, a small-size high-performance antenna backed with periodic structure is proposed. The proposed antenna can be fabricated easily, and it is flexible to tune the resonant frequency.

Chapter 5

A Broadband Microstrip Antenna Based On Periodic Structures

5.1 Introduction

With the demand for high rate mobile telecommunications, there is a corresponding demand for antennas that are high gain and broadband, Microstrip antennas are often used in mobile devices due to their small size, low cost, and ease of fabrication [67]. However, one of the shortcomings of the microstrip antennas is the narrow bandwidth of typically 1-5 % [68]. The narrow bandwidth limits the microstrip antenna in the applications of high data rate.

Many methods are proposed to increase the bandwidth [69-71]. One method is to use a thick and low-permittivity material as the substrate, but the efficiency of the antenna is decreased due to the large amount of surface wave [72].

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The bandwidth of the microstrip antenna can also be improved by adding more layers of substrate and/or resonators above or under the common microstrip antenna [73]. The multilayer patches and stacked multiresonator microstrip antenna also have the property of broadband, but it requires highly accurate fabrication techniques. Moreover, the profile of the antenna is increased.

Another method to increase the bandwidth is to use coupled elements. The parasitic patch antenna, with additional gap-coupled narrow strips or patches, can also achieve a broadband of 25% [74]. But the increase of bandwidth is at the cost of the increase in size of the antenna. The size of parasitic patch antenna is at least three times of the conventional patch antenna.

Etching a U-slot is also an effective method to broaden the operation bandwidth of a microstrip antenna. With a U-slot in the patch, the bandwidth can be increased to as broad as 30% [75].

In [29], a low profile broadband aperture coupled grid slotted patch antenna is proposed. By operating the radiating patch, the bandwidth of the antenna is increased to 28% with the maximum gain of 9.8 dBi.

In this chapter, a broadband and high-gain antenna based on periodic structure is proposed. The proposed antenna is able to achieve a bandwidth of more than 45% with an average gain of about 9 dBi. The periodic structures act as the radiator. Two rows of squares are etched in the ground plane for impedance matching purposes, then the broad bandwidth is achieved without increasing the size of the antenna.

5.2 Antenna Design

The broadband microstrip antenna is designed on the substrate of Rogers RO4003. The dielectric constant(ϵ_r) is 3.38, and the loss tangent($\tan\delta$) is 0.0027. The configuration of the proposed antenna is shown in Fig. 5.1. The antenna comprises of two layers. The upper layer is the radiation element based on an array of 4×4 small patches, which is shown in Fig. 5.1 (b). The bottom layer is the feed network of the antenna. There is one slot in the center of the ground plane to excite the upper layer radiator, and 8 squares are etched to achieve a good impedance match. The geometry of ground plane is shown in Fig. 5.1 (c). Fig. 5.1 (d) shows a 50Ω microstrip which is used to feed the antenna.

5.2.1 Radiator based on periodic structures

The radiator of the proposed antenna is shown in Fig. 5.1 (b). The thickness of the upper substrate is 3.25 mm. There are 16 small patches arranged in a 4×4 array on the top layer of the antenna. The width of the small patch (W_p) is 9mm. The period of the patches (P_p) is 10 mm, and the gap is 1 mm. The small patches form the composite right/left-hand (CRLH) material. The slot between the neighboring small patches acts as the left-handed series capacitor. The right-handed contribution comes from the capacitance from the fringing field at the slot.

In order to get the dispersion relation for the mode analysis, one unit cell is simulated. The simulated unit cell is shown in Fig. 5.2. The dispersion relation

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Figure 5.1: The antenna configuration. (a) Side view. (b) Top view. (c) Mid layer. (d) Back view.

can be obtained using the method as in [29]. It is easy to acquire the ABCD matrix by using the full wave EM simulation software CST Microwave Studio. Since $e^{\gamma z}$ is the eigenvalue of ABCD matrix, the propagation constant (γ) can be obtained from the ABCD matrix by using [5.1].

$$e^{\gamma z} = eig([ABCD]) \quad (5.1)$$

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Figure 5.2: The simulated unit cell. (a) Top view of the metamaterial. (b) Unit cell of metamaterial

The dispersion diagram of the unit cell is shown in Fig. 5.3. In Fig. 5.3 it can be seen that the unit cell is the CRLH structure which is as the expected. The operating principle of the antenna can be investigated using the transmission-line model. The resonance mode frequencies can be derived from (5.2)-(5.5). By evaluating (5.1)-(5.5), the resonant frequencies are estimated. In this case, they are calculated to be 4.94 GHz, 5.72 GHz, and 7.25 GHz. The operating modes are located on the right-handed region of the CRLH periodic structure.

$$f = \frac{\beta_e c}{2\pi\sqrt{\epsilon_{re}}} \quad (5.2)$$

$$\beta_e = \frac{\pi - \beta_u p_p m}{2\Delta L} \quad (5.3)$$

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Figure 5.3: Dispersion diagram of unit cell.

$$\frac{\Delta L}{h} = 0.412 \frac{(\epsilon_{re} + 0.3)(W/h + 0.262)}{(\epsilon_{re} - 0.258)(W/h + 0.813)} \quad (5.4)$$

$$\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} (1 + 12h/W)^{-1/2} \quad (5.5)$$

where c is the speed of light in free space, f is the operating frequency.

5.2.2 Feeding method

In the proposed antenna, the periodic structures based radiator is excited by an aperture in the ground plane. The width and length of the slot are 2 mm and 30 mm, respectively. There are 8 square holes proposed in the ground plane to match the impedance. The holes are arranged in the region with less

5.3. SIMULATED AND MEASURED RESULTS

E-field to avoid large field leakage. The size and position of the square holes are optimized. The geometry of the ground plane is as shown in Fig. 5.1 (c). The width of the square hole (W_s) is 4 mm. The horizontal period of the holes (P_s) is 5.5 mm. The distance from the squares to the slot (g) is 3.75 mm.

5.3 Simulated and Measured Results

5.3.1 Simulated Results

The proposed antenna is simulated using CST Microwave studio. The dimensions of the proposed antenna are shown in Table. 5.3.1 and the parameters are indicated in Fig. 5.1. The reflection coefficient is shown in Fig. 5.4. The solid line shows the simulated result. The proposed microstrip antenna achieves a broad -10 dB impedance bandwidth from 4.69 GHz to 7.47 GHz. There are three valley values which are at 5.06 GHz, 6.15 GHz, and 7.16 GHz. The simulated results are close to the estimated values from (6.1)-(6.5).

Table 5.1: The dimensions of the proposed antenna.

W	60 mm	W_p	9 mm
L	60 mm	P_p	10 mm
h_1	3.25 mm	W_s	4 mm
h_2	0.813 mm	P_s	5.5 mm
L_f	39 mm	g	3.75 mm
W_f	1.85 mm		

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Figure 5.4: Return loss of proposed antenna.

The proposed antenna has a wide impedance bandwidth from 4.69 GHz to 7.47 GHz as shown in Fig. 5.4. The maximum realized gain of this antenna is also plotted in Fig. 5.5. From 4.7GHz to 7.4 GHz, the maximum realized gain of the proposed antenna is in the range from 6.4 dBi to 10.4 dBi. In Fig. 5.6 the radiation patterns of the proposed antenna at 5 GHz, 5.9 GHz, and 7 GHz are plotted. The solid lines are the simulated results. The direction of the main lobes is not changed, and the realized gain is 8.7 dBi, 10.3 dBi, and 6.7 dBi, respectively at 5 GHz, 5.9 GHz, and 7 GHz. Thus, the antenna can be applied in the HiperLAN applications.

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Figure 5.5: Maximum realized gain vs. frequency.

5.3.2 Fabrication and Measured Results

The proposed antenna is fabricated. The photos of the antenna are shown in Fig. 5.7. The antenna is measured in an anechoic chamber. The measured S-parameter and radiation patterns are shown in Fig. 5.4 and Fig. 5.6 respectively. The dashed lines are the measured results. As shown in Fig. 5.4, the measured -10 dB bandwidth is from 4.74 GHz to 7.80 GHz. From 7.5 GHz to 8 GHz, the measured result differs from the simulated one, which is because of the fabrication errors. Fig. 5.5 shows the measured gain of the antenna, and in Fig. 5.6, the radiation patterns at three frequencies are plotted. The measured maximum realized gain of the antenna is 8.30 dB, 9.98 dB, and 6.22 dB, respectively at 5 GHz, 5.9 GHz, and 7 GHz. The measured results agree well with the simulated ones.

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Figure 5.6: The radiation pattern at 5 GHz, 5.9 GHz, and 7 GHz. (a) E plane at 5 GHz. (b) H plane at 5 GHz. (c) E plane at 5.9 GHz. (d) H plane at 5.9 GHz. (e) E plane at 7 GHz. (f) H plane at 7 GHz.

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Figure 5.7: Photos of fabricated antenna. (a) Top layer view. (b) Bottom layer view. (c) Middle layer view

5.4 Summary

In this chapter, a broadband microstrip antenna is proposed. The periodic structures are used to replace the traditional radiator. The radiator comprises of a 4×4 array of square patches, and it is excited through a slot in the ground plane. In the ground plane, 8 squares are etched to achieve the good impedance matching. The proposed antenna can work well from 4.69 GHz to 7.47 GHz. The bandwidth of the proposed antenna is larger than 45%, and the realized gain is from 6.4 dBi to 10.4 dBi (an average of around 9 dBi). The antenna is fabricated, and the measured. The measured results agree well with the simulated ones. According to the broadband and other good performances, the proposed antenna can be used in the HiperLAN applications.

Chapter 6

A Phased Grid Array Antenna with Sub-array for E-Band Application

6.1 Introduction

The demands of millimeter wave frequency have risen in commercial applications in recent years. The unlicensed band around 60 GHz attracts more and more attention [76–79]. Some types of antenna in E-band have been investigated [80,81]. All the designs of the antennas focus on satisfying the requirements of millimeter wave applications. One of the most important demands is the high gain, because the propagation loss around 60 GHz is very large [82–86]. Depending on the atmospheric conditions, the atmospheric absorption can be

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as high as 15-30 dB/km at 60 GHz [12].

The grid array antenna, intended for use in highly integrated 60 GHz radios, is investigated in [87], which can provide the high gain. The gain of the antenna is as high as 13.5 dBi, and the gain can be increased further with the increase of the number of the loops or grids of grid array antenna. However, the gain will be limited by the special structure of grid array. When the grid array antenna is too large (40 loops in [88]), the current along antenna will not be distributed as the description in [89-91], and not all the current will be in phase, which leads the great distortion of the radiation pattern. Zhang, *et al.* proposed utilizing subarrays to avoid the destruction of the radiation pattern caused by incorrect current distribution [88]. The subarrays proposed in [88] are fed by a conventional T-junction network, which is hard to fabricate and integrate with the traditional substrate. Instead, Razavi, *et al.* applied the microstrip ridge gap waveguide to connect the subarrays of antenna array [92], which is also difficult to integrate with PCB.

In this chapter, we propose a high gain grid array antenna with subarrays. Each grid subarray antenna is fed by a slot in substrate integrated waveguide (SIW) to ensure the easy fabrication and integration of the whole antenna system [93-95]. The application of subarrays will avoid the destruction of radiation pattern and the reduction of efficiency in a large grid array antenna. Compared with the widely used microstrip lines and coaxial, the SIW can reduce the transmission loss, which has been discussed in [31]. The SIW can be connected to and excited by a standard waveguide WR-12 using a slotted transition, which

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makes the proposed antenna work easily with the devices using standard interfaces. Thus, the loss of the grid array antenna is reduced by correcting the current along antenna, using low loss feed network, and applying highly efficient feed methods. Additionally, the antenna can achieve beam steering by etching slots in the SIW. Finally, the phased array antenna is easy to fabricate using normal soft substrate.

6.2 Antenna Design

The grid array antenna is designed on a low loss soft substrate. The substrate is double sided Rogers RO3006. The permittivity(ϵ_r) of the substrate is 6.15. The tangent loss($\tan\delta$) is 0.002. The thickness is 0.254 mm. The antenna consists of two layers. The grid array antenna with subarrays is on the top layer, and the feed network is on the bottom layer.

6.2.1 Slot Fed Grid Array Antenna

The grid array antenna with subarrays is constructed on the very top layer. The grid array antenna can be treated as an array of dipoles [96]. The traditional microstrip grid array antenna consists of two different lengths of microstrip lines. The short microstrip lines act as the radiation elements with the length(w_g) of half guided wavelength($\lambda_g/2$). The long microstrip lines are the transmission lines used to connect each radiation element together, and its length(l_g) is around one guided wavelength(λ_g). The width of the microstrip

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line is s_g .

The distance between two neighbouring antennas in an antenna array is usually half wavelength in the free space($\lambda_0/2$) [64]. Considering the period of the subarrays and the permittivity of the used substrate, two columns of grids are arranged for one subarray. The geometry of the subarray is shown in Fig. 6.1. In order to minimize the coupling between two adjacent subarrays, the parallel microstrip lines are removed, which is indicated in Fig. 6.1 [97]. Each subarray looks like a comb as shown in Fig. 6.1 (a). The antenna is designed for operating at 60 GHz, and the dimension of the subarray antenna is shown in Table 6.1.

Table 6.1: Dimension of grid subarray antenna.

l_g	w_g	s_g
2.550 mm	1.275 mm	0.120 mm

The grid array antenna is fed by a slot, which is easy to fabricate compared with the coaxial feed. The slot is etched in the ground plane of the grid array antenna which is also the top metal of SIW feed network. The slot is under one transmission line of the antenna, and the energy is coupled from the SIW to the antenna through the slot. The positions of the slot and the antenna are indicated in Fig. 6.1 (a). The design of slot is similar to the aperture fed antenna. The length of the slot is around half guided wavelength($\lambda_g/2$) to achieve slot resonance [98]. In the proposed design, the antenna is fed by the end shorted SIW in the design as shown in Fig. 6.1 (b). The slot is about half

Figure 6.1: Geometry of a single sub-grid array antenna. (a) top view. (b) bottom view.

guided wavelength($\lambda_g/2$) away from the end of the shorted SIW. In order to get 60 GHz operating frequency, the length of the slot(l_{slot}) is 1.25 mm, and the distance between the slot and the center of end vias(x_s) is 0.75 mm.

6.2.2 SIW Feed Network

The SIW feed network is designed considering the operating frequency of the antenna, the space between subarrays, and phase differences between each antenna subarray. Fig. 6.2 shows the feed network of the grid array antenna with four subarrays. The dimension of SIW depends on the frequency directly, and it is relative to the width of conventional rectangular waveguide supporting the

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same EM wave.

Figure 6.2: Geometry of grid array antenna with four subarrays and SIW transmission network. (a) top view. (b) bottom view and SIW transmission network.

In order to get the only propagation mode to be TE_{10} mode, the width of rectangular waveguide(a_d) can be obtained from (6.1).

$$a_d = \frac{c}{2f_c\sqrt{\epsilon_r}} \quad (6.1)$$

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where c is the speed of light in free space, f_c is the center frequency, ϵ_r is the dielectric constant of the material.

The SIW is composed of two post walls. In the fabrication process, it will be economical if we can consider using the existing moulds and tools. The moulds and tools for posts can be chosen based on (6.2) and (6.3). d is the diameter of each post. p is the period of the posts. The distance between two post walls can be obtained from (6.4).

$$d < \frac{\lambda_g}{5} \quad (6.2)$$

$$p < 2d \quad (6.3)$$

$$a_s = a_d + \frac{d^2}{0.95p} \quad (6.4)$$

where λ_g is the guided wavelength.

The design of the SIW transmission network is similar to the design of dividers which ensures the same phase at each branch. At the T junction, there is a metallic post which is used to match the impedance and decrease the reflection wave. At the bend of branch, the slash row of posts plays the similar role as the matching post to guide the electromagnetic waves. The distance between the branches equals to the distance between two antenna subarrays, which is half wavelength($\lambda_0/2$).

6.2.3 Slotted Transition between SIW and WR-12

Although SIW is easy to fabricate and integrate with the circuit, the traditional rectangular waveguide flange is more widely used and flexible. Thus, it will be useful to allow the connection between the SIW and the standard waveguide flange. Considering the operating frequency of the proposed antenna, we choose WR-12 as the connector.

There are many types of transition. The slotted transition is used here because of its simple structure and easy fabrication. By using the slotted transition between SIW and rectangular waveguide flange, no additional layers are required to add to the antenna. The slot is arranged in the bottom metal layer of the antenna as shown in Fig. 6.2 (b) and Fig. 6.3

Figure 6.3: Geometry of slotted transition and rectangular waveguide.

The electromagnetic waves travel from the rectangular waveguide to sub-

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strate integrated waveguide through the slot. The length of the slot is half guided wavelength (λ_g) to ensure the frequency of coupling waves. The slot cannot be located in the middle of SIW, otherwise the impedance will not be matched well. The center of the slot is $3\lambda/4$ (or odd multiples of quarter wavelength) away from the end of the shorted SIW. In order to couple the maximum field, the slot should be at the center of the traditional rectangular waveguide. In this design, the length of the slot (l_{sw}) is 1.7 mm. The center of the slot is 0.685 mm away from the left post wall of SIW, and it is 1.55 mm away from the end of SIW.

The grid array antenna with four subarrays is simulated using CST studio. The dimension of the antenna is shown in Table 6.2. The simulated S-parameter is shown in Fig. 6.4. The center operating frequency is 60 GHz. The -10 dB bandwidth is from 59.8 GHz to 60.2 GHz, which is in the unlicensed band in almost all the world. The simulated radiation pattern of the antenna is shown in Fig. 6.5. The maximum realized gain occurs in the front direction, and it is 15.11 dB.

Table 6.2: Dimension of the antenna.

l_g	w_g	l_{slot}	x_s
2.550 mm	1.275 mm	1.250 mm	0.685 mm
a_s	l_{sw}	x_{sw}	y_{sw}
1.490 mm	1.700 mm	0.685 mm	1.550 mm

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Figure 6.4: Simulated S parameter of grid array antenna.

Figure 6.5: Radiation pattern of grid array antenna. (a) E plane (x-z plane). (b) H plane (y-z plane).

6.3 Phased Array

It is well known that the beam of an antenna array is narrow. With the increase of the number of antennas in an array, the gain gets higher, but the beam becomes narrower. In order to steer the beam of antenna array without moving

6.3. PHASED ARRAY

the antenna, a phased array is proposed [99][101].

In our design, we propose to use the slotted substrate integrated waveguide phase shifter because of its simple configuration and low insert loss. The geometry of the slotted substrate integrated phase shifter is shown in Fig. 6.6.

Figure 6.6: Geometry of slotted substrate integrated waveguide phase shifter.

Fig. 6.6 shows the equivalent circuit of the slotted substrate integrated waveguide phase shifter. C_s is the coupling slot capacitance, and C_f is the capacitance from the fringing field at the slot. C_s depends on the length and width of the slot. C_f is determined by the length of the slot and the height of the SIW. When the width of the slot and the height of the SIW are constant, C_s and C_f only depend on the length of the slot. Thus, when the length of slot is changed, the phase of the supported electromagnetic wave will be shifted.

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When the length of slot is varied from 0 to 6.4 mm, the phase can be shifted in 360° . Table 6.3 lists the phase differences referring to the configuration that no slot is in the SIW.

Figure 6.7: Equivalent circuit of the slotted substrate integrated waveguide phase shifter.

Table 6.3: Phase differences of various length of slot.

Length of Slots (mm)	Phase Difference (deg)	Length of Slots (mm)	Phase Difference (deg)
0	0	3	-140
0.8	-30	3.4	-180
1.7	-60	4.5	140
2	-80	5	120
2.13	-90	5.7	90
2.36	-110	6.4	30
2.53	-120		

Two simulated results are selected to show two extreme beam directions. The radiation patterns are shown in Fig. 6.8. It can be seen that the beam of the proposed antenna can be steered from 0° to 65° . The phase shifters are set

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symmetrically, so the beam of the proposed antenna can be steered from -65° to 65° .

Figure 6.8: Radiation patterns in x-z plane of phased array.

6.4 Summary

In this chapter, a phased grid antenna with subarrays is designed. Each subarray is fed by a slot, and the low loss SIW is used to connect each subarray. At the port, a simple slot transition is applied to transfer the standard waveguide flange WR-12 and the SIW feed network. The 4-subarray antenna can achieve the realized gain as high as 15.11 dBi, and the gain can be increased further by adding more subarrays easily. The beam of antenna can be steered by etching

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phase shifter at the bottom metallic layer. Since there are no additional layers required for the phase shifter, the antenna structure is simple, and easy to fabricate. The proposed antenna can be realized in two layers of PCB. The beam of the antenna can be steered in the range of $-65^\circ - +65^\circ$.

Chapter 7

Conclusion and Future Work

7.1 Conclusion

This thesis proposed the methods to improve the antenna performances by using ferrites and periodic structures in different frequency bands. Firstly, the dominant influencing factors are analyzed in different frequency bands. The mismatched loss influences the radiated power of the near-ground HF antenna in most time. In microwave band, the dielectric loss of substrate is the dominant factor that influences the efficiency of the microstrip antenna. For millimeter wave antenna array, the loss in the feed network is the main loss factor that reduces the efficiency. Secondly, the methods to improve the performances of the antennas are investigated in different frequency bands.

In HF band, we have proposed to use ferrites to reduce the influence of image current of near-ground antenna. The mismatch loss and efficiency of the HF antenna can be improved. In order to utilize the commercial ferrites more

7.1. CONCLUSION

effectively, the current on the ground plane is analyzed. The current density on the ground indicates the amount and strength of the image current, so the utilization of the ferrites are more effective when the ferrites are arranged in the area with high current density. The proposed method can be used to improve the performance of the antenna at any desired frequency. Moreover, the ferrites can also be used to increase the bandwidth and reduce the size of the near-ground HF antenna. By loading ferrites, the lower bound of the bandwidth can be extended. The fabrication and measured results verify the proposed method which uses the commercial ferrite to improve the performances of the near-ground HF antenna.

In microwave frequency band, we have used the periodic structures to improve the performances of the microstrip antennas. The properties of the original material will be changed when the periodic structures are arranged in the material. By etching the periodic structures in the ground plane, the permittivity is decreased, and the permeability is increased in a certain frequency range. Thus, the size of the antenna is reduced, and the bandwidth is increased. The loss within the material can be reduced even if the inexpensive high-loss material is used, so the energy consumed within the antenna gets less. The efficiency of the antenna is improved. The periodic structure can also be used as the radiator. From the analysis of the transmission line model and the dispersion relation, it can be get that, with the periodic structures in the patch of the antenna, more energy can be radiated out, and the bandwidth of the antenna can be increased greatly.

7.2. FUTURE WORK

In millimeter wave frequency band, we have used the low-loss transmission methods, substrate integrated waveguide, to improve the total efficiency of millimeter wave antenna array. Comparing with the widely used microstrip and coaxial transmission line, the SIW possesses the characteristics of high efficiency, easy fabrication, and easy integration. By using SIW to form the feed network, the efficiency of antenna array is improved, and more antenna elements can be intergrated in the antenna array. The high gain of the antenna array can be achieved. Additionally, the beam of the antenna array can be steered in a large range of 130° by adding slotted phase shifters.

7.2 Future Work

It is well known that more antenna elements in an array can provide higher gain, but the beam of the antenna array is narrow. If the beam can be steered, the signal coverage will be extended. In section [6.3](#), we proposed the phased array antenna with slotted phase shifter. In the design, the slot is used to shift the phase of EM wave in each branch of the array. However, the necessity of refabrication limits the applications of the antenna array.

The flexibility of beam steering determines the performances and applications of the phased antenna array. In the future, we are going to continue investigating the millimeter wave antenna array. We will implement the reconfigurable phase shifter with the simple structure, and apply it on the design of the phased array antenna.

Besides the reconfigurable phased array antenna, we are also going to in-

7.2. FUTURE WORK

investigate the beam-steerable antenna without the beamforming network. The reconfigurable millimeter wave antenna will be designed. In order to steer the beam, the antenna will integrate an inner mechanism(e.g., RF switches, varactors, mechanical actuators, tunable materials, etc). The inner mechanism can be used to redistribute the RF currents over the antenna surface intentionally, so that the reversible modification can be achieved for the beam steering.

Publication List

Journal Publications

1. S. Mi, Y. H. Lee, and Z. Shen, "Ferrite coating design for near-ground HF bowtie antenna," *IEEE Trans. on Antennas Propag.*, vol. 65, 2074–2078, April 2017.
2. S. Mi, and Y. H. Lee, "Design of a compact patch antenna loading periodic Jerusalem crosses," *Progress In Electromagnetics Research M*, vol. 47, 151-159, 2016.
3. S. Mi, and Y. H. Lee, "A broadband metamaterial based microstrip antenna," *IEEE Antennas Wireless Propag. Lett.* (under review)

Conference Publications

1. S. Mi, and Y. H. Lee, "A broadband metamaterial based microstrip antenna," *AP-S/USNC-URSI*, Jul. 2016.
2. S. Mi, and Y. H. Lee, "A waveguide-fed grid E-band array antenna," *APCAP*, Jun. 2015.

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3. X. Ma, S. Mi, and Y. H. Lee, "Design of a microstrip antenna using square Sierpinski fractal EBG structure," APCAP, Jun. 2015.
4. S. Mi, Y. H. Lee, and Z. Shen, "Optimization of ferrite coating for near-ground HF bowtie," AP-S/USNC-URSI, Jul. 2014.
5. S. Mi, and Y. H. Lee, "Design of a compact patch antenna using UC-EBG structure," AP-S/USNC-URSI, Jul. 2013.
6. S. Mi, and Y. H. Lee, "An economical surface coil design without lumped elements for MRI application," PIERS, Mar. 2013.
7. S. Mi, and Y. H. Lee, "Multi-band gap-coupled microstrip patch antenna," APCAP, Aug. 2012.
8. P. Ngor, S. Mi, and P. H. J. Chong, "Energy consumption reduction by multi-hop transmission in cellular network," PIERS, Mar. 2012.

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