Low-profile and end-fire antennas for airborne applications

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LOW-PROFILE AND END-FIRE ANTENNAS FOR AIRBORNE APPLICATIONS

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ABSTRACT

Recent years have witnessed a growing demand for low-profile and end-fire antennas, which can be flush-mounted for aircraft, missile, and unmanned aerial vehicle (UAV) applications. However, there are a lot of limitations associated with traditional end-fire antennas. For example, the performance of log-periodic, Yagi-Uda and Vivaldi antennas is very sensitive to the conducting ground-plane on which these antennas are mounted. H-plane horn antennas can hardly achieve wide bandwidth and maintain low profile simultaneously. In addition, although helical antennas are excellent candidates to achieve end-fire radiation of circular polarization, they are non-planar structures, which are unsuitable to be mounted on moving platforms in consideration of aerodynamic characteristics. In this thesis, we aim to solve these limitations and propose novel solutions to realize low-profile end-fire antennas with various characteristics such as wide bandwidth, high gain, circular polarization, etc., which are potentially very useful to be conformal and flush-mounted for airborne applications.

First, the propagation characteristics of surface waves propagating on a grounded dielectric slab are theoretically analyzed. The slab is assumed to be anisotropic as a general discussion. Next, a wideband and low-profile surface wave antenna based on grounded ceramic slab is proposed. A surface wave launcher is designed by employing a probe-fed parallel plate waveguide with a modified parabolic reflecting wall and a metallic post, which are critical for effectively transforming the cylindrical wave to a unidirectional plane wave within a broad frequency range. The ceramic slab is smoothly tapered to transform the guided surface wave into radiated space wave over a wide bandwidth. The proposed antenna operates over a wide frequency range of 6.1-18 GHz, and has the advantages of wideband, low profile, and high radiation efficiency.

In addition, an H-plane horn antenna exhibiting an ultra-wide bandwidth from 3.4-18 GHz with an antenna thickness of only 5.508 mm is presented. A wideband coaxial-to-waveguide transition is proposed by employing a tapered

ridge and metallic posts in order to enhance the bandwidth and to effectively suppress higher-order modes in the waveguide. An ellipse-shaped copper taper is extended along the horn aperture in order to obtain a smooth transition from horn aperture to free space. A pair of metallic posts is also employed in the H-plane horn to improve the impedance matching. Both planar and conformal versions of the proposed antenna are investigated. The conformal structure is fully embedded in a large cylindrical platform and it is extended to an antenna array configuration.

Furthermore, we propose planar and conformal cavity-backed supergain slot antennas for end-fire radiation. In the proposed structures, slots with different lengths are etched out at proper positions of the cavity wall to generate multiple resonances as well as to decrease the Q value of the cavity, which broadens the bandwidth effectively. The required magnitude and phase of the excited electric field for achieving supergain are obtained by adjusting the inter-element spacing and the position of each slot. Furthermore, off-centered microstrip feedlines are utilized to improve the impedance matching and to excite the required cavity mode. Prototypes of back-to-back structure, conformal structure mounted on a cylindrical platform, as well as conformal array configuration with six separate antenna elements embedded in a conical platform are fabricated and tested. Measured results are in good agreement with simulated ones. They have the advantages such as compact cavity size, enhanced bandwidth, high gain, and high radiation efficiency.

Besides designing linearly polarized end-fire antennas, a planar helical antenna with rectangular cross-section is also presented for achieving end-fire radiation of circular polarization. The helix is formed using printed strips with straight-edge connections implemented by plated via-holes. The currents flowing on the strips and along via-holes of the helix contribute to the horizontal and vertical polarizations, respectively. Besides, the current on the ground plane is utilized to weaken the strong amplitude of the horizontal electric field generated by the one on the strips. Thus, a good circular polarization can be achieved. Furthermore, a tapered helix and conducting side-walls are employed to broaden the axial ratio

bandwidth as well as to improve the end-fire radiation pattern. The designed antenna operates at the center frequency of 10 GHz, which can achieve wide impedance and axial ratio bandwidths of 54% and 34%, respectively. The antenna thickness is only $0.11\lambda_0$ at the center frequency.

This thesis is finally concluded by making a few recommendations for future investigation. With further research on the suggested topics, it is expected that the present limitations may be overcome, possible solutions may be provided, and more novel structures utilizing the new concepts may be proposed.

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CHAPTER 1

INTRODUCTION

1.1 Background

End-fire antennas can direct the maximum radiation toward the axis of the antenna structure. They have been popularly investigated due to the excellent characteristics such as highly directional pattern and high gain. The recent growth of modern wireless communication has increased the demand for end-fire antennas. For example, in some airborne applications such as aircraft, missile, and unmanned aerial vehicle (UAV), low-profile end-fire antennas are usually required to be conformal with curved surfaces of the platforms in order to decrease the aerodynamic drag and meet some specific electromagnetic requirements [1].

1.1.1 End-Fire Array

An end-fire array consists of a number of radiating elements; their phase difference between adjacent elements is adjusted to obtain unidirectional pattern along the axis of the array and destruct interference in any other direction for the desired frequency.

Figure 1.1(a) shows the geometry of an N -element array. All elements are positioned along the z -axis with a uniform spacing d . Assume that the excitation amplitudes of all elements are identical, and each succeeding element has a β progressive excitation phase relative to the preceding one. For identical radiators, the total field can be formed by multiplying the array factor of the isotropic source by the field of a single element [2]. The array factor can be expressed as

$$AF = \sum_{n=1}^N e^{j(n-1)\psi} \quad (1.1)$$

where

$$\psi = kdcos\theta + \beta \quad (1.2)$$

To direct the maximum radiation toward the end-fire direction, either $\theta_0 = 0^\circ$ or 180° ,

$$\psi = kdcos\theta + \beta |_{\theta=0^\circ} = kd + \beta = 0 \Rightarrow \beta = -kd \quad (1.3)$$

or

$$\psi = kdcos\theta + \beta |_{\theta=180^\circ} = -kd + \beta = 0 \Rightarrow \beta = kd \quad (1.4)$$

Thus, the end-fire radiation can be obtained by assigning the phase difference $\beta = kd$ or $\beta = -kd$. Figure 1. 1(b) shows the typical radiation pattern of an end-fire array, and the maximum radiation is directed to $\theta = 0^\circ$.

Figure 1.1: End-fire array, (a) geometry of an N -element array positioned along the z -axis, (b) end-fire array pattern.

Hansen and Woodyard [3] in 1938 proposed the conditions to enhance the directivity of an end-fire array without destroying any of the other characteristics. In this case, the phase shift β should be

$$\beta = -(kd + \frac{2.92}{N}) \Rightarrow \text{for maximum in } \theta = 0^\circ \quad (1.5)$$

$$\beta = +(kd + \frac{2.92}{N}) \Rightarrow \text{for maximum in } \theta = 180^\circ \quad (1.6)$$

These requirements are known as the Hansen-Woodyard conditions for end-fire radiation. The directivity of a Hansen-Woodyard end-fire array is larger than that of the ordinary end-fire array using (1.3) or (1.4).

1.1.2 Traditional End-Fire Antennas

Many types of antennas can radiate end-fire radiation patterns such as Yagi-Uda antenna, log-periodic antenna, Vivaldi antenna, horn antenna, and helical antenna, which are shown in Figure 1.2.

Figure 1.2: Typical end-fire antennas: (a) Yagi-Uda antenna, (b) log-periodic antenna, (c) Vivaldi antenna, (d) horn antenna, and (e) helical antenna.

Yagi-Uda antenna generally consists of a driven element and parasitic elements. This type of antenna was first described by S. Uda [4] in a Japanese article. Later, the operating principle of this antenna was published by H. Yagi [5] in English. Yagi-Uda antenna uses mutual coupling between elements to produce a unidirectionally end-fire radiation pattern [6].

The classic log-periodic antenna structure was proposed by Isbell [7], which is comprised of several dipoles forming a coplanar array. In the array configuration, the adjacent dipole length and the inter-element spacing vary logarithmically. This kind of antenna is frequency independent structure, which can achieve end-fire radiation over a wide bandwidth.

Vivaldi antenna is another kind of wideband end-fire antenna, which was invented by P. J. Gibson [8] in the UK. It is also known as a tapered slot antenna [9]. The antenna structure is very simple and easy to fabricate using a dielectric slab metalized on two symmetric sides. It can be divided into two different types, coplanar [10] and antipodal [11] geometries. The Vivaldi antenna is a kind of continuously scaled antenna structure so that it can achieve unlimited bandwidth for end-fire radiation theoretically [8].

Horn antenna can be regarded as flared waveguide; it can effectively transfer guided waves propagating in the waveguide into radiated waves in free space. Its existence can be dated back to the late 1800s, and it became very popular since the period of World War II owing to its advantages of simple structure, wide bandwidth, and stable end-fire radiation beam [12], [13]. There are many different types of horn antennas, such as E-plane [14], H-plane [15], pyramidal [16], and conical [17] horn antennas, etc. Among them, H-plane horn antenna has been a favorite candidate for airborne applications due to its low-profile structure.

The abovementioned end-fire antennas are linearly polarized. The helical antenna offers a possible solution to achieve circular polarization in the end-fire direction. The helical antenna was invented by J. D. Kraus and its basic concepts were published in 1947 [18]. Basically, helical antenna consists of a ground plane and a conducting wire wound into a helical shape. It can radiate in both normal and axial modes. The axial mode can yield radiation with circular polarization along the axis of the helix, in which case the circumference of the helix is around one wavelength at the center frequency [19], [20].

1.1.3 Surface Wave Antennas

Surface wave antennas have received extensive attention since the 1940s. Generally speaking, a surface wave is bound to the interface of two different dielectrics and can propagate along the interface [1]. In order to induce radiation, the surface impedance must be varied or a surface wave gap should be produced [21], [22]. There are many types of surface wave antennas based on corrugated-metal surface [23]-[25], dielectric rod [26]-[29], and grounded dielectric slab [30], [31], etc.

Figure 1.3(a) shows a corrugated metal surface wave antenna consisting of a row of monopoles. The corrugated structure can be regarded as an effective surface, which can support the propagation of surface waves. At the feeding part, a monopole and a reflector couple a portion of the input power into a surface wave, traveling along the surface and radiating at the end of the structure.

Figure 1.3(b) is a surface wave antenna based on a dielectric rod. In order to excite surface waves, a circular waveguide is usually employed. The fundamental TE_{11} mode is often used in this structure to achieve end-fire radiation, while the first two higher-order modes, TE_{01} and TM_{01} modes, produce a null in the end-fire direction. The body taper of this structure is essential to suppress the sidelobe level of radiation pattern and to increase the antenna bandwidth, while the taper at the end of the structure is critical to reduce the reflection of the surface wave.

The surface wave antenna shown in Figure 1.3(c) employs a tapered grounded slab. A low-profile waveguide is preferred to excite surface waves propagating along the slabs. The slab is tapered in order to radiate surface wave into free space effectively. The taper also plays an essential role in suppressing the sidelobe level of the radiation pattern. In addition, the dielectric constant and thickness of the slab have significant effects on the propagation characteristics of surface waves. In order to achieve a very low profile structure, a slab with high dielectric constant is usually required.

Figure 1.3: Surface wave antennas, (a) corrugated-metal surface, (b) dielectric rod, (c) grounded dielectric slab.

1.1.4 Superdirective Antennas

Superdirective antennas can achieve a much higher directivity in the end-fire direction than that of a reference antenna of the same size [2]. It is generally accepted that antenna directivity can be enhanced by enlarging the aperture size of the antenna. However, in an antenna array, superdirectivity can be obtained by decreasing the element spacing, which contradicts with the common experience and has attracted a lot of attention over recent decades.

The first report on superdirective antenna was probably studied by Oseen in 1922 [32]. Taylor [33] was one of the first to use the term “superdirectivity”. In 1943, Schelkunoff [34] developed the theory of linear end-fire arrays for achieving very high directivities. In 1946, Uzkov [35] presented that the possible maximum directivity value can be N^2 for an N -element array of isotropic radiators. However, he only gave a brief outline of the theory. The validation of the theory can be found in [36]. It presented that for an N -element array of isotropic radiators positioned as shown in Figure 1.1(a) when kd is close to zero, the directivity can be expressed in terms of Legendre polynomials $P_n(\cos\theta)$ as

$$D(\theta) = \sum_{n=0}^{N-1} (2n+1) [P_n(\cos\theta)]^2 \quad (1.7)$$

Thus, the maximum directivity is in the end-fire direction $\theta = 0$, where $P_n(1) = 1$

$$D_{max}(\theta = 0) = \sum_{n=0}^{N-1} (2n+1) = N^2 \quad (1.8)$$

It was also revealed in [36] that the maximum end-fire directivity of a two-element dipole array is around 7.2 dBi, which is 1.2 dB higher than N^2 (6 dBi) maximum directivity of the array of isotropic radiators. For $N = 3$, the maximum directivity of three-element dipole array can reach 10.3 dBi, which is 0.8 dB higher than N^2 (9.5 dBi). As N increases, the difference of the maximum directivity between the dipole array and isotropic radiator array decreases monotonically [36].

1.1.5 Conformal Antennas

A conformal antenna is an antenna that conforms to a specific shape, which can be a part of vehicle, airplane, missile or other structures. Usually, it is built to integrate into a curved surface whose shape is determined by some considerations such as aerodynamic or hydrodynamic [37].

Conformal arrays were first designed and analyzed at least back to the 1930s when a number of dipole elements were arranged to a circular array, which was analyzed by Chireix [38]. They received extensive attention especially during the World War II when the field of phased array antennas was active. Later, numerical methods, electromagnetic analysis, and the understanding of antennas on curved surfaces were improved.

Nowadays, conformal antennas are playing very important roles in our daily life. For aircraft applications, different antennas should be employed for various functions, such as navigation, radar detecting, instrument landing systems, and so on. They are required to be smoothly integrated into the skin of the structures to decrease the aerodynamic drag and fuel consumption. The need for conformal antennas is even more pronounced for the large-sized apertures, which are necessary for military airborne surveillance radars and satellite communications [37].

1.2 Motivation

There is a growing interest to employ end-fire antennas to be surface-mounted on airborne applications such as missile, aircraft, and unmanned aerial vehicle (UAV) for navigation, target detecting and tracking, and data communication between aircraft and ground stations, etc. Special considerations must be taken into account for the design of surface-mounted end-fire antennas. For example, it is required that the antenna performance should be insensitive to the conducting ground plane. It is also desirable that the end-fire antennas can maintain very low-profile structures or even can be fully embedded into the surface of the platform in order to decrease aerodynamic drags. In addition, end-fire antennas with various characteristics such as wide bandwidth, high gain, and circular polarization are required for different applications. However, not all existing end-fire antennas can meet these requirements due to their corresponding limitations.

1.2.1 Limitations of Traditional End-Fire Antennas

Traditional Yagi-Uda antennas consist of a number of resonant elements such as dipole and monopole. High gain can be achieved by employing more directors, and bandwidth can also be enhanced by utilizing bow-tie shape element. However, they are not considered as low-profile structures because the dipole or monopole elements protrude from the surface if the antennas are mounted on airborne platforms, which causes extra aerodynamic drag. Quasi- Yagi-Uda can be an array of dipoles printed on the dielectric slab and fed by microstrip line [39]-[42]. Different from the traditional ones, this kind of Yagi-Uda antenna is low-profile. However, the printed Yagi-Uda antenna is not conduct-backed, which is not a suitable candidate to be surface-mounted.

Log-periodic and Vivaldi antennas can achieve end-fire radiation over an ultra-wide bandwidth, and they possess very simple structures and can achieve very low profiles. Unfortunately, similar to the printed Yagi-Uda antenna, there is no ground plane backing the structures. If they are very close to a conducting plane, the antenna performance will severely deteriorate, which are impracticable for surface-mounted applications.

Horn antennas can alleviate the ground effects to the antenna performance. In order to maintain a low-profile structure, H-plane horn antenna is usually preferred in airborne applications. However, the main constraint is that it is very challenging to simultaneously maintain a very low-profile structure and achieve a wide bandwidth. When the thickness of the antenna is much smaller than the free space wavelength λ_0 , the antenna performance deteriorates significantly. More specifically, when the antenna thickness is smaller than $\lambda_0/6$, a poor matching will be introduced due to the mismatch between the antenna aperture and free space [43]-[45].

In addition, helical antennas are excellent candidates for achieving high gain, broad bandwidth, and end-fire radiation with circular polarization. Unfortunately, they are non-planar, which limits the structures for airborne applications considering the aerodynamic requirement.

1.2.2 Limitations of Surface Wave Antennas

Generally speaking, surface wave antennas based on corrugated metal surface and dielectric rod are not planar structures, which will protrude from platform surfaces for the installation of airborne applications. Surface wave antennas based on grounded dielectric slabs are very promising candidates for surface-mounted and conformal applications owing to the low-profile structures. However, the bandwidth is generally narrow. One of the main constraints of the surface wave antenna is due to the bandwidth limitation of the surface wave launcher, which plays an essential role in determining the performance of the antenna.

The slot Yagi-Uda antenna launcher is one of the favorite candidates for TM_0 excitation for grounded slab surface wave antennas [46], [47]. This is due to the fact that the E-field of a half-wavelength slot in a ground plane can match that of the dominant mode of the slab [48]. However, the bandwidth of the slot Yagi launcher is narrow due to its resonant nature.

Other typical feeding devices widely used in surface wave antennas are rectangular waveguides and horns. They can effectively excite surface waves propagating along the grounded dielectric slab because their aperture fields can be chosen as nearly like the fields of the surface wave modes as possible [49]. However, it is very difficult to realize a good impedance matching for coaxial-to-waveguide transition within a very wide bandwidth since the characteristic impedance of traditional waveguide is a function of frequency and the corresponding value is relatively high compared to 50 ohms of the coaxial feeding line. In addition, as discussed earlier, the very low profile of waveguide and horn structures may limit the enhancement of their operating bandwidth. Therefore, it still remains a challenge to obtain a very wide bandwidth while simultaneously retaining a very low profile for surface wave antennas.

1.2.3 Limitations of Superdirective Antennas

Superdirective antennas are considered to be promising candidates to achieve high gain in the end-fire direction. However, it is a challenge to realize supergain antenna in practice. Superdirective antennas suffer from the following limitations [50]-[52].

1.2.3.1 Narrow Bandwidth

The narrow bandwidth is the major problem for superdirective arrays. In an array, superdirectivity is achieved by inserting more elements within a fixed length [2], which eventually leads to very large current magnitudes and rapid changes of phase in the excitation coefficients of the array elements. The reactive power and Q value increase rapidly. As we know, for a narrow-band antenna, the operating bandwidth is inversely proportional to $1/Q$. Therefore, the bandwidth of superdirective array is generally limited.

1.2.3.2 Sensitivity to the Excitation Amplitude and Phase

The excitation magnitude and phase of each radiator should be properly selected for superdirective arrays. The required tolerances of magnitude and phase of the element excitations for achieving superdirectivity are very sensitive, which makes the superdirective antennas hard to be realized, especially for antenna arrays employing large numbers of elements. Calculations of the required magnitude and phase for achieving the maximum end-fire directivity have been conducted for three-element array of isotropic radiators in [52]. It was clearly shown that the phase and magnitude of each radiator vary with different element spacings in order to obtain the maximum directivity. The maximum end-fire directivity also decreases rapidly with either 5% magnitude error or 5° phase error for the element spacing less than $0.15\lambda_0$. It should be mentioned that if the number of element increases, the values should be more sensitive.

1.2.3.3 Low Radiation Efficiency

As the element spacing of an antenna array gets closer, the radiation resistance decreases significantly. Due to the small radiation resistance, the radiators severely mismatch with free space. If the radiators are matched and forced to radiate power, a strong reactive field is produced and the Ohmic losses increase rapidly.

From the definition of radiation efficiency

$$\eta = \frac{R_{rad}}{R_{rad} + R_{loss}} \quad (1.9)$$

where η denotes the radiation efficiency, R_{rad} and R_{loss} represent the radiation resistance and loss resistance, respectively, we know that the radiation efficiency becomes lower with the decrease of R_{rad} and increase of R_{loss} .

1.3 Objectives

It has been shown in the previous discussions that existing end-fire antennas suffer from a number of limitations, which somehow limit their uses for airborne applications. In this thesis, we aim to present a comprehensive study of the state-of-the-art of end-fire antennas, provide possible solutions to alleviate the associated limitations and propose excellent antenna candidates for airborne installations with various characteristics such as low profile, wide bandwidth, high gain, high radiation efficiency, and circular polarization, etc.

The objectives of the thesis are summarized below.

- 1) Conduct a comprehensive literature review on existing end-fire antennas. The primary target of this is to understand the theories and recent trends of end-fire antennas, explore effective techniques to alleviate the limitations and look for novel methods to improve the antenna performance, which can help to spark the inspiration for designing excellent low-profile and end-fire antenna candidates to be surface-mounted on airborne platforms.

- 2) Design wideband end-fire antennas suitable to be flush-mounted on conducting platforms. It is expected that the proposed antenna can achieve a fractional bandwidth over 100% while retaining the antenna thickness smaller than $1/10 \lambda_0$ at the lowest operating frequency. It is required that the antenna performance should be insensitive to the ground plane on which the antenna is mounted. It is also desirable that the proposed conformal antenna can be fully embedded into the surface of moving platform in order to minimize the aerodynamic drag.
- 3) Explore suitable antenna candidates to achieve supergain with the characteristics of enhanced bandwidth, high radiation efficiency, and low profile. The proposed antenna is expected to be mounted on cylindrical and conical platforms to achieve end-fire radiation.
- 4) Circularly polarized end-fire antennas with a low profile are in great demand for airborne and space-tracking applications. It is our target to design a planar helical antenna to achieve a wideband end-fire radiation of circular polarization while maintaining a very low profile.
- 5) It is also targeted to fabricate and test all prototypes of proposed antennas; measured results shall be compared with simulated ones to validate the design concept. It is also expected to analyze the differences between the simulated and measured results to gain a better understanding of the influences of the measurement environment on the antenna performance.

1.4 Major Contributions of the Thesis

This thesis introduces several antennas with novel structures and excellent performance on the basis of solving the limitations of existing end-fire antennas. These proposed structures can achieve end-fire radiation with unique features such as wide bandwidth, high gain, and circular polarization. They possess low-profile structures, and their antenna performance is insensitive to the ground plane of metallic platforms on which they are mounted, which are potentially

very useful for airborne applications. Major contributions of the thesis are identified as follows.

- 1) Closed-form analytical expressions have been derived to theoretically study the characteristics of surface waves propagating along the grounded dielectric slab. The slab is assumed to be anisotropic as a general discussion. It is revealed that no pure TM or TE mode exists in the grounded slab if the field is non-uniformly distributed along the direction orthogonal to the propagation direction. It is also found that the propagation direction of surface wave deviates from the longitudinal direction of the finite-width slab with some certain angle. Unlike the existing analyses of surface waves whose fields are assumed to be uniformly distributed along the transverse direction of the infinitely large grounded slab, our method takes into account the variations of the fields distributed in the transverse direction of the slab, which has a more practical relevance and can provide a more accurate solution for engineering applications.
- 2) A wideband and low-profile surface wave antenna has been introduced, which can achieve an operating bandwidth from 6.1 to 18 GHz while the antenna thickness is only 3 mm. In the proposed antenna, a unidirectional wideband surface wave launcher is employed to effectively transform the cylindrical wave to a unidirectional plane wave within a broad frequency range. A grounded ceramic slab with high dielectric constant is gradually tapered to radiate surface wave within a wide frequency range. The guidelines for designing the proposed wideband and low-profile surface wave antenna are also summarized. Stable and quasi end-fire radiation beams with moderate gain values are obtained over the entire operating band.
- 3) A wideband and low-profile H-plane horn antenna has been proposed, which can achieve an ultra-wide bandwidth within 3.4-18 GHz and retain the antenna thickness of only 5.508 mm. Several strategies are employed in our proposed antenna by considering the matching improvement,

bandwidth enhancement, higher-order modes suppression, as well as the aerodynamic characteristic, which effectively eliminate the limitations and improve the performance of existing H-plane horn antennas. The guideline of designing the proposed wideband coaxial-to-ridged waveguide transition is summarized and the effects of the platform on the antenna performance are analyzed, which are very useful for facilitating the design of the proposed conformal antenna. In addition, the proposed conformal antenna is fully embedded into a cylindrical conductor by considering the aerodynamic requirement for moving platforms and has been further extended to an array configuration.

- 4) Both planar and conformal versions of a novel supergain slot antenna have been proposed in the thesis, which employ closely spaced three-element slots and a shallow metallic cavity. In the proposed configurations, several techniques are employed to overcome the associated limitations of supergain antennas so that the proposed slot antennas can achieve high gain with end-fire radiation, enhanced bandwidth, and high radiation efficiency. In addition, the proposed conformal antenna can be fully embedded into metallic platforms. Detailed analysis has also been carried out to study the influences of the cylindrical and conical platforms on the antenna performance. Furthermore, six conformal antenna elements are wrapped around the conical platform to form an array configuration. The array structure is immune to electromagnetic waves with arbitrary polarization, which is polarization-independent and potentially very useful for airborne applications.
- 5) A novel planar helical antenna with circular polarization has also been proposed for end-fire radiation. The helix is formed using printed strips with straight-edge connections implemented by plated via-holes. Different from previous planar circularly polarized helical antennas with a square cross-section in the open literature, our proposed antenna employs a rectangular helix. The proposed antenna can achieve wide

impedance and axial ratio fractional bandwidths of 54% and 34%, respectively. The antenna has a very low profile, and its thickness is only $0.11\lambda_0$ at the center frequency. The structure is very simple, lightweight, and low-cost, which can be easily fabricated using the printed circuit board technology. It is very suitable to be mounted on the wings of the unmanned aerial vehicle (UAV) and aircraft.

1.5 Organization of the Thesis

The thesis has been organized into seven chapters. The first chapter primarily introduces the background and motivation behind this work by identifying the basics and limitations of typical end-fire antennas. Based on the limitations, the objectives of the thesis are established. In addition, major contributions of the thesis are summarized.

Chapter 2 presents a comprehensive literature review of end-fire antennas for airborne applications. Traditional, as well as recent antenna designs, are introduced. Several effective techniques and methods are discussed to alleviate the limitations of existing end-fire antennas. Some examples of good antenna candidates suitable for surface-mounted on metallic platforms with a low profile and end-fire radiation patterns are also presented.

Chapter 3 first investigates the propagation characteristics of surface waves propagating along grounded anisotropic dielectric slabs. In the analysis, the field variations along the direction transverse to the propagation direction are carefully taken into account. Next, a wideband and low-profile surface wave antenna is proposed based on tapered grounded ceramic slab. The operating principle and detailed design guidelines of the proposed antenna are presented. Finally, a prototype of the proposed surface wave antenna is fabricated and tested. Measured results are also compared with simulated ones.

Chapter 4 proposes a wideband and low-profile H-plane horn antenna, which is extended into a conformal array configuration. The proposed H-plane horn antenna employs several strategies for improving the performance of the coaxial-

to-waveguide transition by considering the wide bandwidth, low profile, suppression of higher-order modes in the waveguide, as well as the phase distribution of the electric field along the horn aperture. In addition, the effects of the platform on the performance of conformal antenna are carefully analyzed. As a proof of concept, both planar and conformal versions of the antenna are fabricated and tested, the wide bandwidth and nearly end-fire radiation pattern are validated by measured results.

Chapter 5 demonstrates three examples of slot supergain antennas of end-fire radiation based on cavity-backed structure. One is a planar back-to-back structure, while the other two are conformal, which are mounted on cylindrical and conical platforms, respectively. The limitations of traditional supergain antennas are successfully alleviated by employing several techniques. The effects of the metallic platforms on antenna performance are carefully investigated. The conformal array configuration has also been fabricated and tested. The high gain and enhanced bandwidth of the proposed supergain antennas are verified by experiments.

Chapter 6 introduces a novel planar helical antenna of circular polarization based on printed strips with straight-edge connections implemented by plated via-holes. This chapter first investigates the operating principle of the proposed antenna using a one-turn helix structure. Parametric studies are then conducted to analyze the influences of the key parameters on the antenna performance. After careful investigations, the antenna structure is finalized with a tapered four-turn helix. Simulated and measured results are also compared to validate the proof of the concept.

Chapter 7 concludes the thesis by summarizing the achievements of this research. Suggestions for future work are also provided in this chapter.

CHAPTER 2

LITERATURE REVIEW

2.1 Introduction

End-fire antennas have received extensive attention over the past few decades. For airborne applications such as aircraft, missile, and unmanned aerial vehicle (UAV), it is desirable to employ surface-mounted and conformal end-fire antennas with various characteristics such as low profile, wide bandwidth, high gain, high radiation efficiency, as well as circular polarization, etc. In this chapter, a comprehensive literature review on existing end-fire antennas is conducted. The review is mainly divided into three parts: wideband end-fire antennas, supergain antennas, and low-profile circularly polarized antennas.

This chapter first examines the possible wideband end-fire antenna candidates, which are potentially useful for airborne applications. For some traditional end-fire antennas such as Yagi, log-periodic, and Vivaldi antennas, when they are close to the ground plane, the antenna performance will severely deteriorate. It is also well known that it remains a challenge to enhance the bandwidth of H-plane horn and surface wave antennas when they are made very low-profile. In this review, possible solutions are introduced in order to solve these limitations.

Superdirective antennas are considered as promising candidates for conformal end-fire applications. However, they suffer from narrow bandwidth, high mismatch loss, and low radiation efficiency. Many techniques and designs have been investigated to overcome the problems; these methods and examples are briefly reviewed.

In addition, end-fire antennas of circular polarization are reviewed. It is well known that traditional helical antennas can achieve circularly polarized end-fire radiation. However, they are non-planar structures, which are unsuitable for surface-mounted applications. Based on this consideration, planar helical antennas are also carefully investigated in this chapter.

2.2 Wideband End-Fire Antennas

A lot of techniques have been employed to broaden the bandwidth of end-fire antennas in the open literature. Wideband transitions including microstrip-to-slot line transition and coaxial-to-ridged waveguide transition are widely used in many end-fire antennas, which play essential roles in the design of wideband antennas. Frequency independent structures are usually utilized to realize wide bandwidth. In addition, surface wave antennas are very promising to achieve end-fire radiation over a wide bandwidth while maintaining a very low profile.

Figure 2.1: Antennas with wideband microstrip-to-slot line transitions, (a) quasi-Yagi antenna, (b) tapered slot antenna.

2.2.1 Wideband Transition

2.2.1.1 Microstrip-to-Slot Line Transition

One of the most commonly used wideband transitions is the microstrip-to-slot line transition, which is preferred in many end-fire antennas such as quasi-Yagi and tapered slot antennas. In [41], a planar printed quasi-Yagi antenna with a stepped microstrip to slot line transition, as shown in Figure 2.1(a), was proposed,

achieving an excellent end-fire radiation over the operation frequency range of 3.6-11.6 GHz. Figure 2.1 (b) shows a tapered slot antenna employing another wideband microstrip-to-slot line transition [53]-[55]. In [55], a Vivaldi tapered slot antenna can obtain an operating band from 1.23 to 9.91 GHz for VSWR < 2 by utilizing this transition. Other structures such as ultra-wideband balun and coplanar slot feeding structures can also be employed to increase the antenna bandwidth [56], [57]. However, these structures are not conductor backed, which are not suitable for flush-mounted applications.

2.2.1.2 Coaxial-to-Ridged Waveguide Transition

Coaxial-to-rectangular waveguide transitions are usually utilized in designing horn antennas. They can also excite surface waves on grounded dielectric slabs. In the waveguide structures, it is desirable that only the fundamental TE₁₀ mode propagates inside a rectangular waveguide since the higher-order modes of the waveguide can degrade the radiation pattern of the antenna.

Figure 2.2: Structures of ridged waveguide, (a) single-ridged, (b) double-ridged.

Ridged waveguides can be employed to enhance single-mode operation bandwidth. Figure 2.2 shows the structures of single- and double ridged waveguides. The width and height of the waveguide are represented by a and b , respectively. s is the width of the metallic ridge. l is the distance from the ridge to the side wall. d denotes the gap between the ridge and top wall for single ridged waveguide or the gap between two ridges for the double-ridged waveguide. The cutoff conditions of the TE _{$n0$} modes in the waveguide are given by [58].

For odd TE_{n0} modes, $n = 1, 3, 5 \dots$

$$\cot\left(\frac{2\pi}{c}f_c l\right) - \frac{b}{d}\tan\left(\frac{\pi}{c}f_c s\right) - \frac{B}{Y_{01}} = 0 \quad (2.1)$$

For even TE_{n0} modes, $n = 2, 4, 6 \dots$

$$\cot\left(\frac{2\pi}{c}f_c l\right) + \frac{b}{d}\tan\left(\frac{\pi}{c}f_c s\right) - \frac{B}{Y_{01}} = 0 \quad (2.2)$$

where f_c is the cutoff frequency. B/Y_{01} is the normalized susceptance representing the effects of the step discontinuity, which can be obtained from [59]. By solving the above equations, it is known that the cutoff frequency ratio of TE_{30} mode and TE_{10} mode of a ridged waveguide can be much larger than that of the same waveguide without ridge [57]. Therefore, it is effective to widen the single-mode operation bandwidth of the rectangular waveguide by introducing metallic ridge.

The characteristic impedance of a ridged waveguide can be derived as [58], [60]-[63]

$$Z_0 = \frac{Z^\infty}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}} \quad (2.3)$$

where Z^∞ is the corresponding characteristic impedance at the infinitely high frequency, λ is the free-space wavelength, λ_c is the cutoff wavelength of TE_{10} mode in the ridged waveguide. The value of Z^∞ can be calculated using

$$Z^\infty = \frac{120\pi^2}{\frac{k}{\beta} \left\{ m \frac{2\beta}{k} \cos^2 \frac{\pi r}{k} \ln \csc \frac{\pi d}{2b} + \frac{\pi r}{2k} + \frac{1}{4} \sin \frac{2\pi r}{k} + \frac{d}{b} \frac{\cos^2 \frac{\pi r}{k}}{\sin^2 \frac{2\pi \delta}{k}} \left[\frac{\pi \delta}{k} - \frac{1}{4} \sin \frac{4\pi \delta}{k} \right] \right\}} \quad (2.4)$$

where $k = \lambda_c/a$, $\beta = d/a$, $r = s/a$, $\delta = (1-s/a)/2$, $m = 1$ for double-ridged guide, and $m = 2$ for single-ridged guide.

It is well known that for a rectangular waveguide, its characteristic impedance is a function of frequency. Meanwhile, the characteristic impedance of a traditional waveguide is relatively high compared to that of the standard coaxial line. For example, according to the data published in [63], for a rectangular waveguide with $a = 2b$, as d/b equals to one, the value of Z° becomes 120π , which is equal to the intrinsic impedance of free space. Therefore, it is very difficult for the traditional waveguide to match a coaxial line with characteristic impedance of 50 ohms within a wide frequency band. By introducing the ridge, the characteristic impedance of a rectangular waveguide can be significantly decreased, making its matching to a coaxial line much easier [64].

Based on the above analysis, it is concluded that a ridged waveguide section can be employed by the coaxial-to-waveguide transition to enhance the bandwidth effectively as well as to obtain a good impedance matching. This concept has been used in many reported papers [65]-[72]. In [66], a compact and broadband coaxial-to-double-ridged waveguide was proposed, which consists of a stepped impedance transformer and a mode converter, achieving a wide bandwidth of over 3:1. In [67], a double-ridged horn antenna was designed, which can cover a wide frequency range from 1 to 18 GHz. Although these antennas can achieve very wide bandwidth, they are not low-profile. Single-ridged waveguides can be employed in H-plane horn antennas to enhance the bandwidth and to maintain the low-profile structure. For example, a tapered ridge implemented by via-holes was introduced in a substrate integrated waveguide H-plane horn antenna, which can achieve a fractional bandwidth of 76% [70]. In [72], a conformal wideband and low-profile H-plane horn antenna mounted on a large conducting cylinder was presented by employing a step-ridged substrate integrated waveguide and a cone-shaped probe in direct contact with the ridge. This antenna can achieve stable radiation patterns over the entire frequency range of 6.1 to 19 GHz, which is very promising for airborne applications. However, further extending its bandwidth is still very challenging since the higher-order modes in the waveguides will appear as the bandwidth increases, which can degrade the radiation pattern at higher frequencies.

2.2.2 Frequency Independent Structure

In antenna modeling, characteristics such as impedance, pattern, and polarization are invariable if the electrical dimensions of the antenna remain unchanged [2]. Therefore, we can build frequency independent antennas that can be their own scale models to realize wide bandwidth. The log-periodic antenna is a kind of frequency independent structures, which can achieve end-fire radiation pattern over a wide bandwidth.

Figure 2.3: Cavity-backed log-periodic slot antenna.

The traditional log-periodic antenna is a dipole array. In order to achieve a low profile, slot structures based on the printed circuit board (PCB) are explored. A dumb-bell log-periodic slot array was presented in [73], which can achieve an ultra-wide bandwidth over the frequency range of 2-20 GHz. However, the performance of the antenna deteriorates severely when the structure is mounted on a metal surface of airborne platforms. Cavity-backed log-periodic slot structures were proposed to mitigate the effects of the ground on the antenna performance [74], [75]. As shown in Figure 2.3, the depth of the traditional cavity needs to be a quarter wavelength or its multiple, and the cavity dimensions also

need to vary in a log-periodic manner. Therefore, the cavity sizes limit the structures for low-profile applications.

2.2.3 Surface Wave Structure

Surface wave antennas may be another kind of candidates to achieve low profile and wide bandwidth. Surface waves can exist at the interface of two different media with the field exponentially decaying away from the interface, which have been investigated over a few decades. For example, Dyakonov surface waves were studied thoroughly in the literature [76]-[78]. The characteristics of the surface waves propagating at the interfaces between isotropic-to-isotropic, isotropic-to-anisotropic, and isotropic-to-indefinite media were rigorously researched in [79]-[83]. Moreover, surface waves have also been found in many structures such as dielectric rod [27], uniaxial and biaxial slab waveguides [84], [85], plasma slab [86], chiral material [87], and photonic crystal [88], to name only a few.

Figure 2.4: Surface wave bound onto a uniform grounded dielectric interface, decaying exponentially away from the surface.

Among all the geometries of supporting surface waves, one of the best known structures is the grounded dielectric slab, which is very promising for low-profile antenna applications. As shown in Figure 2.4, a uniformly grounded dielectric slab is infinite in the x and y directions, and the propagation is along the $+y$ direction and fields have no variation in the x direction. The lossless slab is of thickness h in the $+z$ direction and has a relative permittivity of ϵ_r . The following relation is obvious [1]

$$\beta_y^2 = k_0^2 + \alpha_z^2 \quad (2.5)$$

where β_y is the phase constant of a surface wave in the +y direction, k_0 represents the wave number in free space, α_z denotes the attenuation constant in the +z direction. Due to the fact that the field exponentially decays in air away from the dielectric surface, surface waves are intrinsically bound onto the dielectric interface.

The propagation constant β_y and cutoff frequency f_c of surface wave propagating in the grounded dielectric slab satisfies the following relationships [76].

For TM mode

$$k_1 \tan(k_1 h) = \varepsilon_r k_2 \quad (2.6)$$

$$k_1^2 + k_2^2 = (\varepsilon_r - 1)k_0^2 \quad (2.7)$$

$$f_c = \frac{nc}{2h\sqrt{\varepsilon_r - 1}}, \quad n = 0, 1, 2 \dots \quad (2.8)$$

For TE mode

$$-k_1 \cot(k_1 h) = \varepsilon_r k_2 \quad (2.9)$$

$$k_1^2 + k_2^2 = (\varepsilon_r - 1)k_0^2 \quad (2.10)$$

$$f_c = \frac{(2n - 1)c}{4h\sqrt{\varepsilon_r - 1}}, \quad n = 1, 2, 3 \dots \quad (2.11)$$

where $k_1^2 = \varepsilon_r k_0^2 - \beta_y^2$ and $k_2^2 = \beta_y^2 - k_0^2$.

It should be mentioned that the fundamental mode of a surface wave propagating along a grounded slab is TM₀ mode, which has no cutoff frequency. This mode may be potentially very useful for wideband surface wave antennas.

Surface wave antennas based on grounded dielectric slab have the advantages of low profile, moderate gain with end-fire radiation, and structurally conformal

to airborne installations with good aerodynamic characteristics [89]-[91]. Surface wave launcher plays a very important role in surface wave antenna, and its excitation efficiency and bandwidth have profound effects on the antenna performance. The slot Yagi-Uda antenna launcher is one of the favorite candidates for TM_0 mode excitation for grounded slab surface wave antennas. However, its bandwidth is usually narrow. Other typical feeding devices widely used in surface wave antennas are rectangular waveguides and flare horns with the aperture field chosen as nearly like the transverse field of the surface wave as possible. Nevertheless, it is very challenging for these waveguide-based feeding structures to achieve very wide bandwidth as well as to maintain a low profile.

Figure 2.5: Geometry of a tapered depth flared slot antenna, (a) top view, (b) side view.

Figure 2.5 shows the geometry of a tapered depth flared slot antenna. It is a surface wave structure, which consists of a rectangular waveguide and a tapered

slot filled with dielectric. The bandwidth of the structure is limited by the feeding waveguide and discontinuity at the beginning of the aperture [92], [93]. In [93], a tapered-ladder discontinuity minimizer and ridged waveguide were employed to improve the impedance matching as well as to increase the bandwidth of the antenna, achieving a wide range of operation bandwidth over a 4:1 frequency band. However, the thickness of the dielectric is $0.32\lambda_0$, which may not be competitive among the low-profile structures.

Another limitation associated with the rectangular waveguide launcher is that the distribution of the electric field of the TE mode at the waveguide aperture is not very uniform, which is proportional to the cosine function. Therefore, it leaves room for us to explore better surface wave launchers to obtain a more uniform distribution of the electric field at the aperture to improve the surface wave excitation efficiency.

2.3 Supergain Antennas

A gain of N^2 represents a remarkable “supergain” for N -element array of isotropic radiators. As we know, due to the limitations of superdirective antennas, it is a challenge to achieve supergain in practice. However, much work has been reported on superdirective antennas in the open literature. Effective techniques have been utilized to overcome the limitations.

2.3.1 Bow-Tie Structure

It is well known that the biconical antenna shown in Figure 2.6(a) can achieve a very wide bandwidth. It is formed by placing two cones of infinite extent together. The biconical antenna can evolve into a planar structure, which is called bow-tie. Bow-tie structures can be employed by dipole and monopole antennas as shown in Figures 2.6(b) and (c) to increase the bandwidth.

Figure 2.6: (a) Biconical antenna, (b) bow-tie dipole, (c) bow-tie monopole.

A supergain array based on bow-tie monopole structure was presented in [94]. In the configuration, three monopole elements were employed. Both director and reflector utilized the bow-tie structure to enhance the bandwidth. By adjusting the reflector and director's resonant frequencies, the gain of the antenna can also be maximized. Four compact three-element monopole Yagi antennas were designed in [94] to analyze the tradeoff between the bandwidth and gain, achieving bandwidth from 4.2% to 10.2%, and gain values from 9.5 dBi to 9.9 dBi.

2.3.2 Folded Arm Technique

In practical applications, coaxial lines with a characteristic impedance of 50 or 75 ohms are usually used to feed the antenna. The dipole has an input impedance of $Z_{in} = 73 + j42.5$ for the length of $\lambda/2$. For $\lambda/4$ monopole, the value is $Z_{in} = 36.5 + j21.25$. However, the input resistance of dipole or monopole array decreases when the element spacing becomes closer. It is shown in [95] that the input impedance of a two-element monopole array is only around 4.5 ohms with the element spacing of 0.1λ . The rapidly decreased input resistance causes a severe mismatch with coaxial lines.

Figure 2.7: (a) Folded dipole and the analysis modes, (b) three-element folded monopole supergain array.

One of the techniques to increase the impedance of a dipole is to fold the wire, which forms a rectangular loop. Figure 2.7(a) shows the folded dipole structure and the analysis modes. The structure can be analyzed using even and odd modes if the two larger sides of the loop are closely spaced [96]. When the length of the arm is half a wavelength and the antenna is resonant, the input impedance can be derived as

$$Z_{in} = 4Z_d \quad (2.12)$$

where Z_d denotes the input impedance of a dipole without being folded for the length of $l = \lambda/2$. It is seen that the input impedance is four times larger than that of a typical unfolded dipole.

If the dipole is multi-folded to N arms, the input impedance can be increased to:

$$Z_{in} = N^2 Z_d \quad (2.13)$$

Figure 2.7(b) shows the configuration of a three-element folded monopole supergain array analyzed in [97]. It is composed of a driven element, a reflector,

and a director. The element spacing is designed to be $0.02\lambda_0$. Each element is multi-folded, which aims to boost up the input impedance to improve the impedance matching. The structure was fabricated and measured. The antenna can achieve a gain value of 10.08 dBi, and measurement results are in good agreement with simulated ones.

The folded arm technique was also employed in many other examples of supergain antennas. In [94], the driven element of a three-element monopole Yagi array was folded. Measured results indicated that the gain can approach 10 dBi. In [98], a two-element dipole Yagi antenna with an element spacing of $0.02\lambda_0$ was designed, the director is folded twice. The maximum realized gain can reach 6.6 dBi. The monopole version was also fabricated and measured, showing that the antenna gain is 8.83 dBi, which is about 4 dB higher than that of a $\lambda/4$ monopole. In [99], a dipole element is closely located parallel to a perfect electric conductor ground plane, and the spacing is only 0.023λ . By employing multi-folded arms, the antenna can obtain a radiation efficiency of over 90% with a feed point resistance of 46 ohms. The gain of the antenna was measured to be 8.2 dBi.

2.3.3 Decoupling and Matching Networks

A more complicated but effective way to realize supergain antenna is to design the impedance matching and decoupling networks; its schematic of the supergain antenna is shown in Figure 2.8. The networks usually contain several lumped elements and transmission lines, which form some specific circuit topologies to decrease the element coupling and to improve the impedance matching.

A simple tuning technique to improve the matching was proposed in [95]. In this paper, the input impedances of two folded monopole antennas as a function of spacing were first investigated. When the two antennas are closely spaced, one of them exhibits a very large reactance. In order to adjust the reactance value to be zero, some capacitors were employed between the folded arms and ground plane. In [100], a multiport impedance-matching network was utilized to match the antenna as well as to optimize the radiation pattern for a maximum gain. Ideal

circuit topology was proposed and the circuit sensitivity to the tolerance of the components was also studied. A gain of 8.9 dBi and a fractional bandwidth of 5.8% can be achieved for two-element folded monopole antenna using this impedance-matching network.

Figure 2.8: Schematic of the supergain antenna with impedance matching and decoupling networks.

Decoupling and matching networks are usually narrowband. A broadband matching and decoupling network was proposed in [101]. Five different network versions for the same 0.1λ spaced two-port monopole antenna array were analyzed and compared. An 180° rat-race coupler with two series inductors (chip inductors) and two shunt capacitors (shunt open stubs) were used to provide an improvement in matching over the 100-MHz bandwidth from 2.4 to 2.5 GHz. The antenna can achieve the gain of 8 dBi and radiation efficiency of 86%.

2.4 Low-Profile Circularly Polarized Antennas

Besides linearly polarized antennas discussed above, circularly polarized antennas are also in great demand for airborne applications since they can suppress the multipath reflection, avoid the polarization mismatch, and support a flexible alignment for the receiving and transmitting antennas. Many CP antennas, such as helical antenna [20], [102], spiral antenna [103], patch antenna [104], [105], crossed dipole antenna [106], etc., have been investigated over the past decades. Among these reported CP antennas, helical antennas can exhibit CP end-fire radiation, high gain, and wide bandwidth [19]. However, traditional

helical antennas are not suitable for surface-mounted and space-constrained applications because their profiles are generally high.

2.4.1 Non-Planar Helical Antennas

Traditional helical antennas can achieve circular polarization in the axial-mode when the element circumference is approximately one wavelength, the pitch angle α of the helix is within the range of $12^\circ < \alpha < 14^\circ$, and the number of helical turns is usually larger than 3 [19]. However, helical antennas are not regarded as low-profile structures.

A lot of effects have been made to design compact and low-profile helical antennas. One of the methods is to combine a small number of turns and low pitch angle [107]-[109]. In [107], a two-turn helix with a very low pitch angle of 4° was employed to achieve a fractional bandwidth of 12% for circular polarization. Based on this configuration, the low-profile helical antenna array was considered in [108] and [109]. Meandering techniques were also employed in many structures to reduce the antenna size [110]-[113]. In [113], an axial length reduction of around 33% was achieved for a printed quadrifilar helical antenna when a meandering line was utilized.

In addition, a wire composite helical-spiral antenna was proposed in [114], which is shown in Figure 2.9. In this configuration, only one-turn helix is employed. The size of the antenna is reduced to 77% of a traditional spiral antenna. The current is gradually attenuating along the spiral arm and radiating into free space. Tilted beam away from the end-fire direction with circular polarization is obtained. Furthermore, a hemispherical helical antenna shown in Figure 2.10 can also achieve a compact size, low profile, and circular polarization in the end-fire direction [115], [116]. The proposed antenna can achieve a fractional axial ratio bandwidth of 14.6%, and circularly polarized radiation can cover a wide angular range of 90° .

Figure 2.9: Low-profile composite helical-spiral antenna, (a) perspective view, (b) side view.

Figure 2.10: Low-profile hemispherical helical antenna.

2.4.2 Planar Helical Antennas

2.4.2.1 Square Helix

A square helix can be formed using a conducting wire wrapped around a dielectric core with a square cross-section, as shown in Figure 2.11. The current

is flowing along the conducting wire at four sides of square dielectric core sequentially. Many theoretical analyses have been conducted to investigate the radiation characteristic of the square helix, while the effects of the dielectric constant of the core on the characteristics of one, two, and three-turn of helix have also been examined [19], [117], [118].

Figure 2.11: Square helix with one-turn.

In [106], rigorous formulas were derived for the radiated field of a square helical antenna. For the simplest case of one-turn square helix, the normalized electric field strength $F(\theta, \varphi)$ in the end-fire direction was derived as

$$F_{\theta}(\theta = 0, \varphi) = -i \frac{8}{\pi} e^{i\varphi} \quad (2.14)$$

$$F_{\varphi}(\theta = 0, \varphi) = \frac{8}{\pi} e^{i\varphi} \quad (2.15)$$

which lead to

$$\frac{F_\theta(\theta = 0, \varphi)}{F_\varphi(\theta = 0, \varphi)} = -i \quad (2.16)$$

It is obvious that in the end-fire direction $\theta = 0$, the horizontal and vertical electric field components have equal amplitude and 90° phase difference, so that circular polarization can be obtained.

Based on the analysis, many helical antennas with square cross-section were proposed [119]-[122]. In [120], a square helical antenna was integrated into a semiconductor silicon substrate, achieving good CP performance while maintaining the substrate thickness of $0.22\lambda_0$ at the center frequency. In [122], a bifilar helix antenna was designed for unmanned aerial vehicle applications. In order to enhance the operation bandwidth, a tapered helix was employed in the bifilar helical antenna, achieving end-fire circular polarization over a fractional bandwidth of 1.68:1.

2.4.2.2 Rectangular Helix

Compared to the square helix, a rectangular helix has the advantage of low profile. As shown in Figure 2.12, the rectangular helix can be fabricated using a single layer of the substrate based on printed circuit board technology, which consists of printed strips and straight-edge connections implemented by plated via-holes. The proposed helix structure has found many applications in slow-wave structures [123]-[126] and antennas [127]-[136].

Although helical antennas with rectangular cross-section have been investigated over recent decades, little literature has been reported on circularly polarized end-fire antennas so far. In [128]-[130], planar helical antennas with very low profile were presented, but they only work in normal modes. A compact integrated two-wire helical antenna was proposed in [131], which can achieve end-fire radiation for the Bluetooth application. However, it is a linearly polarized antenna and the operation bandwidth is very narrow. In [132], a novel antenna based on confined planar helix was designed. Slow waves can propagate

along a conducting screen and radiate at the end of the structure for end-fire radiation, but the antenna is also linearly polarized. In addition, circularly polarized planar helical antennas were designed in [133], [134]. Unfortunately, they suffer from the limitation of narrow bandwidth, and the circular polarization is not positioning along the end-fire direction.

Figure 2.12: Rectangular helix consisting of plated strips with straight-edge connections implemented by plated via-holes.

A recent design based on planar helix was proposed to achieve circular polarization in the end-fire direction [135], [136]. The antenna consists of two layers of substrates and a large ground. Two different substrate layers are separated by a certain distance and connected by metallic posts. End-fire radiation with a relatively wide axial ratio bandwidth of over 30% was obtained. However, the height of the planar helix is a bit thick, which is $0.27\lambda_0$ at the center frequency. In addition, a large ground vertical to the axis of the helix is employed in this planar helical antenna, which limits the structure for low-profile applications.

2.5 Summary

In this chapter, end-fire antennas with various characteristics such as wide bandwidth, supergain, and circular polarization have been comprehensively reviewed. Several techniques have been introduced to solve the associated problems of these end-fire antennas effectively.

It has been shown in the literature review that cavity-backed structures can help to alleviate the ground effect on the antenna performance and antenna bandwidth can be enhanced by using wideband transitions. It has been noted that the limitations of superdirective antennas can be overcome by employing bow-tie structures, folded arm technology, and decoupling and matching networks. In addition, planar helical antennas have also been proposed to achieve end-fire radiation of circular polarization, which may be promising to achieve low-profile structures for conformal applications.

Although the reviewed techniques can alleviate the limitations of existing end-fire antennas to a certain extent, there still exist a lot of challenges for our further investigation. For example, it is very hard to achieve a wide bandwidth while maintaining the antenna thickness less than $\lambda_0/6$ for H-plane horn antenna since the impedance matching deteriorates significantly when maintaining a very low profile. In addition, most published supergain antennas are dipole and monopole arrays; little has been reported on slot structures, which leaves room for us to have a further investigation. Furthermore, it still remains a challenge to realize circularly polarized end-fire radiation for a planar helical antenna with a very low profile. The large ground plane vertical to the axis of the helix is also not suitable for low-profile applications. All these challenges offer a clear direction for our research in the following chapters.

CHAPTER 3

SURFACE WAVE ANTENNA BASED ON GROUNDED DIELECTRIC SLAB

3.1 Introduction

As discussed in the previous chapter, surface waves can propagate along the grounded dielectric slab with the field exponentially decaying away from the interface. They can also radiate into free space for end-fire radiation at discontinuities or when a surface wave gap is produced. These excellent characteristics have attracted extensive attention over recent decades.

A lot of studies have been carried out to characterize surface waves propagating along a grounded dielectric slab. For instance, it has been shown that the infinitely large grounded dielectric slab can support pure TM and TE surface wave modes, and their propagating constants on isotropic slab were analyzed in [79]. The corresponding fields and power of the surface waves were studied in [30]. In addition to the case of isotropic grounded slab, the guidance conditions of TM and TE surface waves in grounded anisotropic slab were considered in [137]. The solution of suppressing surface waves on grounded anisotropic slab was explored in [138]. However, these analyses are impractical because they only focus on structures with infinite length and width, and assume that there is no field variation along the direction orthogonal to the propagation direction of surface waves. To the best of our knowledge, no detailed analysis has been introduced for the case that the fields are non-uniformly distributed along the direction transverse to the propagation direction, which is a more practical concern for most engineers. In addition, most substrates used in real applications exhibit nonnegligible anisotropic properties due to either the intrinsic anisotropy of the material or the manufacturing process [139]. The effect of the anisotropy of the material on the propagation characteristics of surface waves is another concern for practical engineering.

On the other hand, surface wave antennas based on grounded dielectric slabs exhibit the advantages of low profile, moderate gain with end-fire radiation, and structurally conformal to airborne installations with good aerodynamic characteristics. It is known that surface wave launchers play an important role in the antenna performance. Rectangular and flare horns are favorite low-profile candidates to feed the surface wave antennas based on grounded dielectric slab, but one of the limitations associated with these launchers is that it is challenging to achieve a very wide bandwidth.

In this chapter, we first investigate the characteristics of surface waves propagating on a grounded dielectric slab, especially focus on the slab with finite width. The slab is considered as anisotropic as a general discussion. By assuming that the surface waves are propagating along the longitudinal direction of the slab, the field variations along the transverse direction are considered. By solving Maxwell's equations in closed form, it is revealed that no pure TM or TE mode exists if the fields are non-uniformly distributed along the transverse direction of the grounded slab. Instead, two hybrid modes, namely quasi-TM and quasi-TE modes are supported. It is also found that surface waves propagate away from the longitudinal direction of the finite-width slab with some certain angle, and the angle decreases as the slab width increases. It is worth mentioning that the proposed method is also suitable for analyzing the characteristics of surface waves propagating on isotropic grounded slab as a special case.

Next, a compact and wideband surface wave antenna that exhibits very low profile and can be flush-mounted on a conformal platform is presented. The proposed antenna comprises a tapered grounded ceramic slab, a unidirectional surface wave launcher, and a tapered impedance transition. The ceramic slab is smoothly tapered to transform the guided surface wave into radiated space wave over a wide frequency range. The surface wave launcher employs a probe-fed parallel plate waveguide with a modified parabolic reflecting wall and a metallic post, which are critical for effectively transforming the cylindrical wave to a unidirectional plane wave within a broad frequency range. The proposed surface wave antenna has the following advantages. 1) The antenna can achieve a wide

bandwidth of 100% with a thickness of only $0.12 \lambda_0$ at the center operating frequency. 2) Stable and quasi end-fire radiation beams with moderate gain values are obtained over the entire operating band. 3) The antenna is structurally simple and conductor-backed, which is potentially useful to be conformal and flush-mounted for missile and unmanned aerial vehicle applications.

Figure 3.1: Surface waves on the interface of air region and grounded anisotropic slab with infinite length in x and y directions.

3.2 Characteristics of Surface Waves Propagating on Grounded Anisotropic Slab

3.2.1 Infinitely Large Anisotropic Grounded Slab

As shown in Figure 3.1, we assume that the grounded slab is infinite in both x and y directions, the surface waves propagate along the longitudinal $+y$ direction with an $e^{-j\beta y}$ propagation factor. The thickness of the slab is t . The slab is considered as anisotropic as a general discussion, and the relative permittivity $\bar{\epsilon}$ and permeability $\bar{\mu}$ of the anisotropic slab are expressed as

$$\bar{\epsilon} = \begin{bmatrix} \epsilon_x & 0 & 0 \\ 0 & \epsilon_y & 0 \\ 0 & 0 & \epsilon_z \end{bmatrix}, \quad \bar{\mu} = \begin{bmatrix} \mu_x & 0 & 0 \\ 0 & \mu_y & 0 \\ 0 & 0 & \mu_z \end{bmatrix} \quad (3.1)$$

where the diagonal elements represent the eigenvalues of $\bar{\bar{\epsilon}}$ and $\bar{\bar{\mu}}$, respectively, and their directions constitute the principle axes of the anisotropic slab. In order to analyze the characteristics of surface waves propagating along the grounded anisotropic dielectric slab, we separately consider the fields in anisotropic and air regions, then solve Maxwell's equations and apply boundary conditions across the interface. For the considered structure, the E-field and H-field along the y direction for both regions are expressed as

In Region I

$$E_y^I = A_I f_e(x) e^{-j\beta_y y} \sin(k_z z) \quad (3.2a)$$

$$H_y^I = B_I f_h(x) e^{-j\beta_y y} \cos(k_z z) \quad (3.2b)$$

In Region II

$$E_y^{II} = A_{II} f_e(x) e^{-j\beta_y y} e^{-k_{0z} z} \quad (3.3a)$$

$$H_y^{II} = B_{II} f_h(x) e^{-j\beta_y y} e^{-k_{0z} z} \quad (3.3b)$$

where A_I, A_{II}, B_I , and B_{II} are the amplitude coefficients of the corresponding fields in each region. β_y is the phase constant of surface waves along the y direction. k_z is the wavenumber in the anisotropic medium in the z direction. The sine function of E_y^I and cosine function of H_y^I in the anisotropic medium are selected due to the vanishing electric field and maximum magnetic field on the PEC ground plane, respectively. Due to the fields of surface waves decaying exponentially away in air from the interface, the z components of fields in air region is expressed as $e^{-k_{0z} z}$. In addition, $f_e(x)$ and $f_h(x)$ represent the x components of E-field and H-field, respectively. For the infinitely large slab, we assume that the propagation constant in the x direction is R_x , $f_e(x)$ and $f_h(x)$ then satisfy the following form

$$f_e(x) = f_h(x) = f(x) = \begin{cases} e^{-R_x x} & (x \geq 0) \\ e^{R_x x} & (x \leq 0) \end{cases} \quad (3.4)$$

If $R_x = 0$, then $\partial f(x)/\partial x = 0$, which indicates that the field has no variation in the x direction. On the other hand, if $R_x \neq 0$, the field of surface wave is non-uniformly distributed along the x direction.

By solving Maxwell's equations based on (3.2) and (3.3), the E-field and H-field in both anisotropic slab and air regions in x and z directions are found as

In Region I

$$E_x^I = \frac{e^{-j\beta_y y} \sin(k_z z)}{j(\epsilon_x \mu_z k_0^2 - \beta_y^2)} [\beta_y A_I \frac{\partial f_e(x)}{\partial x} + \omega \mu_0 \mu_z k_z B_I f_h(x)] \quad (3.5a)$$

$$H_x^I = \frac{e^{-j\beta_y y} \cos(k_z z)}{j(\epsilon_z \mu_x k_0^2 - \beta_y^2)} [\beta_y B_I \frac{\partial f_h(x)}{\partial x} + \omega \epsilon_0 \epsilon_z k_z A_I f_e(x)] \quad (3.5b)$$

$$E_z^I = \frac{e^{-j\beta_y y} \cos(k_z z)}{j(\epsilon_z \mu_x k_0^2 - \beta_y^2)} [\beta_y k_z A_I f_e(x) + \omega \mu_0 \mu_x B_I \frac{\partial f_h(x)}{\partial x}] \quad (3.5c)$$

$$H_z^I = \frac{e^{-j\beta_y y} \sin(k_z z)}{j(\epsilon_x \mu_z k_0^2 - \beta_y^2)} [-\beta_y k_z B_I f_h(x) - \omega \epsilon_0 \epsilon_x A_I \frac{\partial f_e(x)}{\partial x}] \quad (3.5d)$$

In Region II

$$E_x^{II} = \frac{e^{-j\beta_y y} e^{-k_{0z} z}}{j(k_0^2 - \beta_y^2)} [\beta_y A_{II} \frac{\partial f_e(x)}{\partial x} + \omega \mu_0 k_{0z} B_{II} f_h(x)] \quad (3.6a)$$

$$H_y^{II} = \frac{e^{-j\beta_y y} e^{-k_{0z} z}}{j(k_0^2 - \beta_y^2)} [\beta_y B_{II} \frac{\partial f_h(x)}{\partial x} - \omega \epsilon_0 k_{0z} A_{II} f_e(x)] \quad (3.6b)$$

$$E_z^{II} = \frac{e^{-j\beta_y y} e^{-k_{0z} z}}{j(k_0^2 - \beta_y^2)} [-\beta_y k_{0z} A_{II} f_e(x) + \omega \mu_0 B_{II} \frac{\partial f_h(x)}{\partial x}] \quad (3.6c)$$

$$H_z^{II} = \frac{e^{-j\beta_y y} e^{-k_{0z} z}}{j(k_0^2 - \beta_y^2)} [-\beta_y k_{0z} B_{II} f_h(x) - \omega \epsilon_0 A_{II} \frac{\partial f_e(x)}{\partial x}] \quad (3.6d)$$

Based on the field distributions and boundary conditions, the characteristics of the surface waves can be analyzed as follows.

1) *TM Mode Surface Wave*: For the pure TM surface wave, it is known that $B_I = B_{II} = 0$, the continuity conditions of E_x and E_y at $z = t$ give

$$A_I \sin(k_z t) = A_{II} e^{-k_{0z} t} \quad (3.7a)$$

$$\frac{\partial f(x)}{\partial x} \frac{A_I \sin(k_z t)}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} = \frac{\partial f(x)}{\partial x} \frac{A_{II} e^{-k_{0z} t}}{k_0^2 - \beta_y^2} \quad (3.7b)$$

which lead to

$$\varepsilon_x \mu_z = 1 \quad (3.8a)$$

or

$$\frac{\partial f(x)}{\partial x} = 0 \quad (3.8b)$$

It is obvious that (3.8a) is not satisfied since the dielectric slab cannot be air for our discussion. Therefore, (3.8b) is the only solution, indicating that for pure TM surface wave propagating along the y direction of the grounded anisotropic slab, the field distribution along the x direction must be uniform.

2) *TE Mode Surface Wave*: For pure TE surface wave, it is obvious that $A_I = A_{II} = 0$. Similarly, according to the boundary conditions of H_x and H_y at the interface $z = t$, (3.8b) also should be satisfied.

3) *Hybrid Mode Surface Wave*: Based on the above discussions, it is concluded that if the pure TM or TE surface wave exists on the grounded dielectric slab, the field distribution along the x direction must be uniform. In other words, if $\partial f(x)/\partial x \neq 0$ or $R_x \neq 0$, hybrid mode surface waves that consist of quasi-TM mode with small H_y and quasi-TE mode with small E_y propagate along the slab. In this case, the continuity of tangential fields across the interface leads to

$$\begin{aligned} & \beta_y^2 R_x^2 k_0^2 (1 - \varepsilon_x \mu_z) (1 - \varepsilon_z \mu_x) \\ &= [\mu_z k_z (k_0^2 - \beta_y^2) - k_{0z} (\varepsilon_x \mu_z k_0^2 - \beta_y^2) \cot(k_z t)] \\ & \times [\varepsilon_z k_z (k_0^2 - \beta_y^2) + k_{0z} (\varepsilon_z \mu_x k_0^2 - \beta_y^2) \tan(k_z t)] \end{aligned} \quad (3.9)$$

By applying Maxwell's equation, the dispersion relations in both regions are

In Region I

$$\begin{aligned}
& \beta_y^2 R_x^2 k_z^2 \left(\frac{1}{\varepsilon_z \mu_x k_0^2 - \beta_y^2} - \frac{1}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} \right)^2 \\
& = -k_0^2 \left(\mu_y + \frac{\mu_x R_x^2}{\varepsilon_z \mu_x k_0^2 - \beta_y^2} - \frac{\mu_z k_z^2}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} \right) \left(\varepsilon_y + \frac{\varepsilon_x R_x^2}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} - \frac{\varepsilon_z k_z^2}{\varepsilon_z \mu_x k_0^2 - \beta_y^2} \right)
\end{aligned} \tag{3.10}$$

In Region II

$$\beta_y^2 - R_x^2 - k_{0z}^2 = k_0^2 \tag{3.11}$$

From (3.9)-(3.11), it is seen that three equations with four unknowns β_y , R_x , k_z , and k_{0z} are obtained, which implies that there is no unique solution for β_y unless R_x is determined. It is worth mentioning that for pure TM and TE surface waves, β_y , k_z , and k_{0z} can be solved for given k_0 , slab thickness t , and relative permittivity $\bar{\varepsilon}$ and permeability $\bar{\mu}$ by substituting $R_x = 0$ into (3.9)-(3.11). The above analyses can also be applied to the grounded isotropic slab, in which case we can use the relative permittivity $\varepsilon_r = \varepsilon_x = \varepsilon_y = \varepsilon_z$ and permeability $\mu_r = \mu_x = \mu_y = \mu_z$ instead of $\bar{\varepsilon}$ and $\bar{\mu}$.

3.2.2 Finite-Width Anisotropic Grounded Slab

Figure 3.2(a) shows the grounded anisotropic slab with a finite width of $2a$ and thickness t . By applying the boundary conditions at the interface $x = \pm a$, the propagation constant in the x direction R_x can then be determined. For analysis, the cross-section of the grounded slab is sub-divided into six regions, as shown in Figure 3.2(b). Region I is the anisotropic slab, while the others are air regions. Since the fields in two corner Regions V and VI are very small, which are assumed to be negligible [141], we only take into account Regions I, II, III, and IV for analysis.

Figure 3.2: Surface waves on the interface of air region and grounded anisotropic slab with finite width, (a) perspective view, (b) cross-section view.

The E-field and H-field along the y direction for Regions I and II have the same expressions as (3.2a)-(3.3b). However, due to the discontinuity at $x = \pm a$, reflected waves exist along the x direction. Therefore, the x components $f_e(x)$ and $f_h(x)$ of the E-Field and H-field are expressed as

$$f_e(x) = e^{-R_x x} + r_1 e^{R_x x} \quad (3.12a)$$

$$f_h(x) = e^{-R_x x} + r_2 e^{R_x x} \quad (3.12b)$$

where r_1 and r_2 represent the magnitude ratios of $e^{R_x x}$ and $e^{-R_x x}$ for $f_e(x)$ and $f_h(x)$, respectively. The fields in both Regions I and II in x and z directions can be obtained by substituting (3.12a) and (3.12b) into (3.5a)-(3.6d).

In Regions III and IV, the fields are decaying away from the interface $x = \pm a$ along the x direction. Therefore, the x components of fields can be expressed as $e^{-R_{0x} x}$ and $e^{R_{0x} x}$, respectively. Based on the continuity conditions of the tangent fields at the interface, the E-field and H-field along the y direction for Regions III and IV are given as follows.

In Region III

$$E_y^{\text{III}} = A_{\text{III}} e^{-R_{0x} x} e^{-j\beta_y y} \sin(k_z z) \quad (3.13a)$$

$$H_y^{\text{III}} = B_{\text{III}} e^{-R_{0x} x} e^{-j\beta_y y} \cos(k_z z) \quad (3.13b)$$

In Region IV

$$E_y^{\text{IV}} = A_{\text{IV}} e^{R_{0x}x} e^{-j\beta_y y} \sin(k_z z) \quad (3.14a)$$

$$H_y^{\text{IV}} = B_{\text{IV}} e^{R_{0x}x} e^{-j\beta_y y} \cos(k_z z) \quad (3.14b)$$

where A_{III} , B_{III} , A_{IV} , and B_{IV} are corresponding amplitude coefficients. The E-field and H-field in the x and z directions can then be found as

In Region III

$$E_x^{\text{III}} = \frac{e^{-j\beta_y y} e^{-R_{0x}x} \sin(k_z z)}{j(k_0^2 - \beta_y^2)} [-\beta_y R_{0x} A_{\text{III}} + \omega \mu_0 k_z B_{\text{III}}] \quad (3.15a)$$

$$H_x^{\text{III}} = \frac{e^{-j\beta_y y} e^{-R_{0x}x} \cos(k_z z)}{j(k_0^2 - \beta_y^2)} [-\beta_y R_{0x} B_{\text{III}} + \omega \varepsilon_0 k_z A_{\text{III}}] \quad (3.15b)$$

$$E_z^{\text{III}} = \frac{e^{-j\beta_y y} e^{-R_{0x}x} \cos(k_z z)}{j(k_0^2 - \beta_y^2)} [\beta_y k_z A_{\text{III}} - \omega \mu_0 R_{0x} B_{\text{III}}] \quad (3.15c)$$

$$H_z^{\text{III}} = \frac{e^{-j\beta_y y} e^{-R_{0x}x} \sin(k_z z)}{j(k_0^2 - \beta_y^2)} [-\beta_y k_z B_{\text{III}} + \omega \varepsilon_0 R_{0x} A_{\text{III}}] \quad (3.15d)$$

In Region IV

$$E_x^{\text{IV}} = \frac{e^{-j\beta_y y} e^{R_{0x}x} \sin(k_z z)}{j(k_0^2 - \beta_y^2)} [\beta_y R_{0x} A_{\text{IV}} + \omega \mu_0 k_z B_{\text{IV}}] \quad (3.16a)$$

$$H_x^{\text{IV}} = \frac{e^{-j\beta_y y} e^{R_{0x}x} \cos(k_z z)}{j(k_0^2 - \beta_y^2)} [\beta_y R_{0x} B_{\text{IV}} + \omega \varepsilon_0 k_z A_{\text{IV}}] \quad (3.16b)$$

$$E_z^{\text{IV}} = \frac{e^{-j\beta_y y} e^{R_{0x}x} \cos(k_z z)}{j(k_0^2 - \beta_y^2)} [\beta_y k_z A_{\text{IV}} + \omega \mu_0 R_{0x} B_{\text{IV}}] \quad (3.16c)$$

$$H_z^{\text{IV}} = \frac{e^{-j\beta_y y} e^{R_{0x}x} \sin(k_z z)}{j(k_0^2 - \beta_y^2)} [-\beta_y k_z B_{\text{IV}} - \omega \varepsilon_0 R_{0x} A_{\text{IV}}] \quad (3.16d)$$

The boundary condition across the interface $z = t$ provides the same result as (3.9), while the field continuities at the interface $x = \pm a$ give

$$\beta_y^2 k_z^2 \left(\frac{1}{\varepsilon_z \mu_x k_0^2 - \beta_y^2} - \frac{1}{k_0^2 - \beta_y^2} \right) \left(\frac{1}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} - \frac{1}{k_0^2 - \beta_y^2} \right)$$

$$= k_0^2 \left(-\frac{\mu_x R_x}{\varepsilon_z \mu_x k_0^2 - \beta_y^2} + \frac{R_{0x}}{k_0^2 - \beta_y^2} \frac{e^{-R_x a} - r_1 e^{R_x a}}{e^{-R_x a} + r_1 e^{R_x a}} \right) \left(-\frac{\varepsilon_x R_x}{\varepsilon_x \mu_z k_0^2 - \beta_y^2} + \frac{R_{0x}}{k_0^2 - \beta_y^2} \frac{e^{-R_x a} + r_1 e^{R_x a}}{e^{-R_x a} - r_1 e^{R_x a}} \right) \quad (3.17)$$

and

$$r_1 = -r_2 = \begin{cases} 1 & \text{quasi-TM mode} \\ -1 & \text{quasi-TE mode} \end{cases} \quad (3.18)$$

In addition, the dispersion relation of both Regions III and IV can be found as

$$\beta_y^2 - R_{0x}^2 + k_z^2 = k_0^2 \quad (3.19)$$

Therefore, from (3.9)-(3.11) and (3.17)-(3.19), it is seen that six equations with six unknowns β_y , R_x , R_{0x} , k_z , k_{0z} , and r_1 are derived, so that the propagation characteristics of the quasi-TM and quasi-TE surface waves can be determined.

3.2.3 Numerical Analysis

3.2.3.1 Comparison of Different Methods

In this section, three different methods are employed to compare the propagation characteristics of surface waves on an anisotropic grounded dielectric slab. In the first method, it is assumed that the slab is infinite in both x and y directions, and there is no field variation along the x direction. This method has been utilized in most open literature, but it can only provide approximate results for real engineering applications, in which finite-width slabs are usually employed, and the field distributions along the x direction are non-uniform. The propagation characteristics are summarized in (B.1)-(B.6) in Appendix B. The second method is based on Marcatili's method [140]. This method considers the field variations along the x direction of the finite-width slab. However, the analysis assumes that $H_z = 0$ and $H_x = 0$ for quasi-TM and quasi-TE modes, respectively. Based on this approximate method, we derived the closed-form expressions for propagation characteristics of quasi-TM and quasi-TE modes for the considered structure, which can be obtained using (C.2)-(C.6) and (C.8)-(C.12), respectively, as

summarized in Appendix. C. The third one is based on our derivations using (3.9)-(3.11) and (3.17)-(3.19). In our method, hybrid surface wave modes on grounded dielectric slab with finite width are analyzed, which addresses the main concern and provides a more accurate solution for practical applications.

As an example, we use an anisotropic PTFE substrate in [139] with $\varepsilon_x = 2.89$, $\varepsilon_y = 2.95$, $\varepsilon_z = 2.45$, and $\mu_x = \mu_y = \mu_z = 1$ to analyze the propagation constants of surface waves, and the results at $f = 16$ GHz are shown in Figure 3.3(a). In addition, the corresponding results of isotropic slab with $\varepsilon_x = \varepsilon_y = \varepsilon_z = 2.45$, $\mu_x = \mu_y = \mu_z = 1$ are also shown in Figure 3.3(b) for comparison. In our calculation, we are looking for solutions that waves propagate along the y -direction and decay in the z direction away from the interface. For the first method, the slab width is assumed to be infinite with $R_x = 0$, while for the other two methods, the slab width is set as $2a = 37.5$ mm with $R_x \neq 0$. Figures 3.3(a) and (b) both indicate that the results of finite-width slab obtained by Marcattili's method and our method are very close. However, it is observed that they are smaller than the corresponding values obtained using the first method. Furthermore, comparing the case of anisotropic with isotropic slabs, it is seen that the anisotropic property of the slab has significant effects on both quasi-TM and quasi-TE modes, which will be discussed in detail in the next section.

As shown in Figure 3.4(a), the values of β_y/k_0 as a function of t/λ_0 with different slab widths are compared. In this study, our method based on (3.9)-(3.11) and (3.17)-(3.19) is employed for the case of finite-width slab, while for the case of $a = \infty$, the first method is utilized. It is seen that when the slab width increases, the values of β_y/k_0 increase, and they are closer to the case of $a = \infty$ with $R_x = 0$. This phenomenon also validates our derivations in a way.

Figure 3.3: Comparison of β_y/k_0 as a function of slab thickness t/λ_0 for grounded dielectric slab, (a) anisotropic slab with $\epsilon_x = 2.89$, $\epsilon_y = 2.95$, $\epsilon_z = 2.45$ and $\mu_x = \mu_y = \mu_z = 1$, (b) isotropic slab with $\epsilon_x = \epsilon_y = \epsilon_z = 2.45$, $\mu_x = \mu_y = \mu_z = 1$.
($a = \infty$ for $R_x = 0$, $2a = 37.5$ mm for $R_x \neq 0$.)

Figure 3.4: Effects of slab width on (a) β_y / k_0 and (b) $|\beta_y / R_x|$ as a function of slab thickness for grounded anisotropic slab ($\epsilon_x = 2.95$, $\epsilon_y = 2.89$, $\epsilon_z = 2.45$ and $\mu_x = \mu_y = \mu_z = 1$).

Furthermore, it is worth pointing out that the propagation constant in the x direction R_x is revealed to be a pure imaginary number in our calculation, which indicates that the surface waves propagate along the x direction in addition to the y direction, so that the propagating direction deviates from the y direction with some certain angle. The values of $|\beta_y/R_x|$ as a function of t/λ_0 with different slab widths are plotted in Figure 3.4(b). The results indicate that $|\beta_y/R_x|$ is very sensitive to the slab width. When the slab width increase, the value of $|\beta_y/R_x|$ increases significantly, meaning that the propagation direction is closer to the y direction.

3.2.3.2 Effects of Permittivity Tensor on Propagation Characteristics of Surface Waves

The effects of the relative permittivity tensor on propagation characteristics of surface waves are investigated. Based on (3.9)-(3.11) and (3.17)-(3.19), the corresponding values of β_y/k_0 and $|\beta_y/R_x|$ as a function of slab thickness are calculated. As an instance, the calculated frequency is set as $f = 16$ GHz, and the slab width is $2a = 56.25$ mm.

Figures 3.5(a) and (b) show the effects of ε_x on β_y/k_0 and $|\beta_y/R_x|$ as a function of slab thickness. It is seen that when both ε_y and ε_z remain to be 2.45 and the value of ε_x varies from 2.45 to 3.15, β_y/k_0 and $|\beta_y/R_x|$ are almost stable for quasi-TM modes. However, the corresponding values of quasi-TE modes change significantly with the increase of ε_x . Similarly, the effects of ε_y and ε_z on the propagation characteristics of surface waves are also investigated. Figures 3.5(c) and (d) illustrate that increasing ε_y results in slight increases of β_y/k_0 and $|\beta_y/R_x|$ for quasi-TM modes, while the quasi-TE modes are nearly unaffected. As shown in Figures 5.5(e) and (f), the value of ε_z can significantly affect the propagation characteristics of quasi-TM modes, but hardly contributes to the quasi-TE modes. Since most substrates exhibit a nonnegligible amount of anisotropy during the process of fabrication, these effects should be seriously taken into account for engineering applications. It should be mentioned that although the relative permittivity tensor can affect the propagation constants along both x and y

Figure 3.5: Effects of (a), (b) ϵ_x , (c), (d) ϵ_y , and (e), (f) ϵ_z on β_y/k_0 and $|\beta_y/R_x|$ as a function of slab thickness for grounded anisotropic slab with slab width of $2a = 56.25$ mm.

directions, the slab width plays a major role in affecting the value of $|\beta_y/R_x|$, as displayed in Figure 3.4(b). In addition, we also employ the other two methods discussed above to analyze these effects as well as to validate our derivation further. The phenomena are in good agreement by using different methods, and these discussions are omitted here for the sake of brevity.

3.3 Wideband Flush-Mounted Surface Wave Antenna

3.3.1 Antenna Configuration

Figure 3.6 shows the configuration of our proposed wideband and low-profile surface wave antenna. It consists of a wideband probe-fed unidirectional surface wave launcher, a tapered grounded ceramic slab, and a tapered impedance transition. The launcher and slab are seamlessly joined together by tapering one end of both parts. The length and width of the antenna are denoted by L and W , respectively.

Figure 3.6: Configuration of the proposed wideband, low-profile, and compact surface wave antenna, (a) perspective view, (b) side view.

The surface wave launcher consists of a coaxial-to-parallel plate waveguide transition with a modified parabolic reflecting wall and a metallic post. The shape of the reflecting wall can be expressed as $|x| = (4py)^n$. The distance between the coaxial probe and the reflecting wall is p . The coaxial probe and metallic post are located along the symmetric plane of the launcher with a separation of d_p , and they protrude into the substrate with heights of h_f and h_p , respectively.

The grounded slab is smoothly tapered to transform the surface wave into radiated space wave as well as to improve the matching condition between the dielectric slab and free space. As shown in Figure 3.6(b), the taper is mainly determined by three segments AB , BC , and CD . The lengths of these three segments are represented by L_{AB} , L_{BC} , and L_{CD} , respectively. The thicknesses of the slab at points A , B , C , and D are denoted by h_A , h_B , h_C , and h_D , respectively. By selecting the proper length of each segment and thickness at each point, the curvature of the ground can be controlled. It should be mentioned that the ground curvature has a significant effect on the sidelobe levels of the radiation patterns, and it can also be controlled by employing more segments or other arc-shaped tapers. In our design, three linear segments are used in order to facilitate simulation and fabrication.

3.3.2 Design Considerations

3.3.2.1 Tapered Grounded Dielectric Slab

As discussed in Section 3.2, grounded dielectric slab with finite width can support quasi-TM surface wave mode. For our proposed surface wave antenna, it is generally required that most of the surface wave power can be retained inside the dielectric slab right after it enters the tapered grounded slab from a parallel plate waveguide. This can eliminate the strong reflection and potential diffraction occurring at the feeding junction, which may degrade the radiation pattern.

For surface waves propagating along the $+y$ direction of a grounded dielectric slab with the thickness of h and width w , the power can be derived as

$$\begin{aligned}
P &= \frac{1}{2} \int Re(\bar{\mathbf{E}} \times \bar{\mathbf{H}}^*) \cdot d\bar{\mathbf{S}} = \frac{1}{2} \int Re \begin{vmatrix} \hat{x} & \hat{y} & \hat{z} \\ E_x & E_y & E_z \\ H_x^* & H_y^* & H_z^* \end{vmatrix} \cdot d\bar{\mathbf{S}} \\
&= \frac{1}{2} \iint Re(E_z H_x^* - E_x H_z^*) dx dz \quad (3.20)
\end{aligned}$$

Therefore, the power inside the slab $P_1|_{z=0}^{z=h}$ and outside the slab $P_2|_{z=h}^{z=\infty}$ can be obtained

$$P_1|_{z=0}^{z=h} = \frac{1}{2} \int_0^h \int_{w/2}^{-w/2} Re[E_z^I (H_x^I)^* - E_x^I (H_z^I)^*] dx dz \quad (3.21)$$

$$P_2|_{z=h}^{z=\infty} = \frac{1}{2} \int_h^\infty \int_{w/2}^{-w/2} Re[E_z^{II} (H_x^{II})^* - E_x^{II} (H_z^{II})^*] dx dz \quad (3.22)$$

where the superscripts I and II denote the slab and air regions, respectively. Generally speaking, in order to minimize the reflection and diffraction at the feeding junction, the power inside the slab $P_1|_{z=0}^{z=h}$ over the total power ($P_1|_{z=0}^{z=h} + P_2|_{z=h}^{z=\infty}$) should satisfy [31]

$$C = \frac{P_1|_{z=0}^{z=h}}{P_1|_{z=0}^{z=h} + P_2|_{z=h}^{z=\infty}} > 0.9 \quad (3.23)$$

It is worth mentioning that all field components inside and outside the grounded slab can be calculated based on (3.5), (3.6), (3.9)-(3.12), and (3.17)-(3.19). These derivations are also suitable for the isotropic grounded slab, in which case we can set $\epsilon_x = \epsilon_y = \epsilon_z = \epsilon_r$ and $\mu_x = \mu_y = \mu_z = \mu_r$.

Although surface waves are bound waves, they can radiate into free space once a discontinuity occurs or a surface wave gap is produced. When looking straight down from the $+z$ direction, the grounded dielectric slab can be regarded as a transmission line, which is shorted at the bottom. For the tapered grounded slab, the surface impedance alters continuously as the height changes smoothly. When the slab height reaches $\lambda/4\sqrt{\epsilon_r}$, the short circuit at the bottom end is transformed into an open circuit at the interface. At this resonant point, the guided surface wave can be effectively transformed into a space wave that can be radiated into

space [21].

In order to effectively transform the surface wave into radiated space wave over a wide frequency band, a large range of thickness should be selected for the dielectric slab for a given ϵ_r . On the other hand, the slab thickness should be maintained low so that a compact and low-profile antenna can be designed, which indicates that a large ϵ_r should be selected. Finally, a tapered grounded dielectric slab with a high dielectric constant ϵ_r is employed to design a wideband and low-profile surface wave antenna.

The tapered grounded slab radiator can be designed based on the above understandings, and the procedures are summarized below.

- 1) First, the slab width W can be determined based on the desired beamwidth in the H-plane. For the TM surface wave propagating in the grounded dielectric slab, the half-power beamwidth (HPBW) of the radiation pattern in the H-plane is approximately [2]

$$\Theta_h = 114.6 \sin^{-1} \left(\frac{0.443\lambda}{W} \right) \text{ degrees} \quad (3.24)$$

In our example, the slab width is selected to be 29 mm so that the HPBW is around 45° at the center frequency $f = 12$ GHz.

- 2) Next, the antenna height h_A and dielectric constant ϵ_r of the tapered slab can be determined once the starting frequency f_{min} of the operating band is specified. In our design, the bandwidth from 6 to 18 GHz is desired and a large ϵ_r is preferred to maintain a low profile. h_A should satisfy

$$h_A > \lambda_{min} / 4\sqrt{\epsilon_r} \quad (3.25)$$

where λ_{min} is the free-space wavelength at the starting frequency f_{min} . Meanwhile, it is desirable that the power ratio C at the beginning part of the tapered slab is larger than 0.9. Therefore, we calculate the power ratio C in terms of slab thickness, dielectric constant and wavelength, which is shown in Figure 3.7. It is seen that the power ratios inside the slab are around 0.6 when the slab thickness is $h = \lambda / 4\sqrt{\epsilon_r}$ for different dielectric

constants. As the slab thickness increases, the power ratio C increases meaning that more power is concentrated inside the slab. Based on this analysis, we select a ceramic slab with a high dielectric constant $\epsilon_r = 25$ and height $h_A = 3\text{mm}$. In this case, the power ratio C is 0.92 so that the reflection and diffraction at the feeding junction can be negligible.

Figure 3.7: Relationship of power ratio C , slab thickness, dielectric constant, and wavelength ($W = 29\text{ mm}$).

- 3) Furthermore, the length L_s of the tapered slab can be initially determined by considering both the beam tilt-angle in the E-plane and the impedance matching condition between the dielectric slab and free space. The guided surface waves can be radiated into free space at the resonant position of the tapered grounded slab, where the slab thickness is $\lambda/4\sqrt{\epsilon_r}$. Therefore, the region starting from the resonant point to the end of the slab can be considered as an extended ground plane, whose length L_g affects the beam tilt-angle of the pattern in the E-plane [1]. As L_g increases, the beam angle tilts closer to the end-fire direction. When $L_g \cong 0.8\lambda$, the radiation beam is around 45° away from the end-fire direction.

Figure 3.8: Effects of L_s on VSWR of the linearly tapered grounded slab ($h_A = 3$ mm, $h_D = 0.2$ mm, $W = 29$ mm, $\epsilon_r = 25$).

On the other hand, in order to smoothen the impedance transition from the dielectric slab to free space, a longer length L_s is preferred, and h_D is better kept small to widen the bandwidth though it may be limited by the fabrication constraints. However, a long slab takes up more space and leads to a less compact structure. Therefore, a suitable length should be selected to compromise the matching condition and compactness. As a simple instance, a waveport is placed at the thick side of the tapered slab for excitation using ANSYS High Frequency Structure Simulator (HFSS). The initial parameters are $\epsilon_r = 25$, $h_A = 3$ mm, $h_D = 0.2$ mm, $W = 29$ mm. The effects of slab length L_s on the port's voltage standing wave ratio (VSWR) are shown in Figure 3.8. It is seen that a lower VSWR is obtained as L_s increases. When L_s is 47 mm, the VSWR value remains less than 2 within the entire frequency range. Thus, L_s can be initially selected to be 47mm considering both the beam angle and matching condition in our design.

- 4) Finally, the curvature of the ground can be carefully designed by considering the sidelobe level (SLL) of the radiation pattern. Since the ground shape affects the surface impedance variation along the tapered slab

and the field distribution along the surface, it plays a significant role on the radiation pattern, especially for the SLL [2], [22]. A simple way to control the curvature is to divide the tapered ground into several segments, and then vary the length and slope of each segment properly. In our design, three segments are employed in order to simplify the simulation. In the simulated model, the SLL of the tapered slab is treated as the objective of the optimization problem, which is a function of six parameters h_A , h_B , h_C , L_{AB} , L_{BC} , L_{CD} , and the operating frequency. Within a reasonable range of values, all these parameters are carefully tuned and finalized to obtain an acceptable SLL over the entire band. After optimization, the final parameters are $h_A = 3$ mm, $h_B = 2.1$ mm, $h_C = 1$ mm, $h_D = 0.2$ mm, $L_{AB} = 11$ mm, $L_{BC} = 15.5$ mm, and $L_{CD} = 20.5$ mm. It should be mentioned that using more segments or arc-shaped ground may also result in a satisfactory result. However, this can increase the complexity of design and fabrication, and the simulation process may also be time-consuming.

In order to illustrate the significances of the ground curvature on the radiation patterns, we summarized the comparison of sidelobe levels (SLLs) of the linearly tapered slab (#1) and the three-segment one (#2) in Table 3.1. The E-plane is the y - z plane, and the H-plane is the cut plane orthogonal to the E-plane and includes the maximum beam peak. At lower frequencies, the radiation beams are wide, and no sidelobes appear. Therefore, Table 3.1 only lists the SLLs at frequencies larger than 12 GHz. It is seen that the SLLs of the E-plane are large at higher frequencies for the linearly tapered grounded slab. In comparison, they are effectively improved for the three-segment taper. It is worth mentioning that the gain values of the three-segment taper are slightly larger than the linearly tapered one with the improvement in SLL at higher frequencies. In addition, slight change of the ground shape has little effect on the VSWR of the slab since a good impedance matching is mainly obtained by employing a sufficiently long tapered structure. Therefore, the VSWR is almost unchanged by employing a three-segment tapered slab compared to the linearly tapered one.

Table 3.1: Sidelobe levels (SLLs) comparison of the linearly tapered slab (#1: $h_A = 3$ mm, $h_D = 0.2$ mm, $L_s = 47$ mm) and three-segment tapered slab (#2: $h_A = 3$ mm, $h_B = 2.1$ mm, $h_C = 1$ mm, $h_D = 0.2$ mm, $L_{AB} = 11$ mm, $L_{BC} = 15.5$ mm, $L_{CD} = 20.5$ mm)

f (GHz)		<12	12	14	16	18
#1 SLL (dB)	E-plane	-	-7.5	-7	-2.6	-5
	H-plane	-	-9	-12.9	-13	-17
#2 SLL (dB)	E-plane	-	-8.5	-8.3	-9.5	-9.1
	H-plane	-	-12.7	-13.3	-17.8	-16.2

As discussed, surface waves are bound to the interface of the tapered grounded dielectric slab. However, they can be transformed into radiated space waves effectively in the tapered structure when the slab thickness reaches $\lambda/4\sqrt{\epsilon_r}$. As a proof of concept, a waveport is used to excite the tapered grounded dielectric slab. Figure 3.9 shows the tangential electric field distributions of the tapered grounded dielectric slab surface and in free space at different frequencies, which clearly illustrates the transformation from guided surface waves to radiated space waves via the tapered structure. As seen, the areas showing the strongest electric field intensity of the tapered slab effectively transform the surface waves into space wave, while the areas before that, starting from the waveport, guide the waves. The radiation occurs at different thicknesses of the tapered slab at different frequencies, the corresponding thickness is $\lambda/4\sqrt{\epsilon_r}$, as expected.

3.3.2.2 Surface Wave Launcher

The surface wave launcher is designed based on the following considerations. First, a broad bandwidth should be obtained when the structure maintains a very low profile. Second, in order to improve the launching efficiency, it is required that the phase differences between waves arriving at the waveguide aperture are minimized, and it is desirable to obtain a uniform distribution of the electric field at the waveguide aperture as well. Therefore, the cylindrical waves excited by the coaxial probe should be transformed to unidirectional quasi-plane waves effectively.

Figure 3.9: Tangential E-field distributions of the tapered grounded dielectric slab surface and in free space at different frequencies, (a) 6 GHz, (b) 12 GHz, (c) 18 GHz.

A few strategies are employed in our structure to meet the aforementioned requirements. The center frequency of our designed antenna is $f_0 = 12$ GHz. A Teflon sheet with a low permittivity of 2.1 is used as the slab in the parallel-plate waveguide because it has the same dielectric constant as the Teflon material inside the SMA feed. The thickness is selected to be 3mm, which is the same height as the grounded tapered slab. The top, bottom, and reflecting walls of the Teflon are covered with copper tapes to form the parallel-plate waveguide, which supports the propagation of the quasi-TEM wave.

Figure 3.10: Effects of metallic post on VSWR of the launcher ($\epsilon_r = 2.1$, $h = 3$ mm, $W = 29.1$ mm, $n = 0.5$, $h_f = 2.85$ mm, $p = 4.3$ mm).

1) Metallic Post: In the launcher, a metallic post is utilized to serve as a matching element, which helps to broaden the bandwidth of the launcher to a certain extent. The post partially penetrating into the waveguide acts as a shunt capacitive reactance, whose value is related to the height and diameter of the post. The mutual coupling between the coaxial probe and the post is determined by their spacing. By varying the length, diameter, and location of the post, the self-impedance of the metallic post and the mutual impedance between the probe and post can be adjusted. This can affect the input impedance of the launcher and help to improve the impedance matching. Figure 3.10 shows the simulated VSWR of

our proposed surface wave launcher with and without the metallic post. The VSWR is above 1.8 over frequency range without the post. However, the value is much lower with a metallic post of suitable height and at an appropriate location, which indicates that the impedance matching is significantly improved.

Figure 3.11: Effects of distance p on VSWR of the launcher ($\epsilon_r = 2.1$, $h = 3$ mm, $W = 29.1$ mm, $n = 0.5$, $h_f = 2.85$ mm, $d_p = 2.4$ mm, $h_p = 1.7$ mm.)

2) *Reflecting Wall*: The coaxial probe is fed at the focal point of the parabolic reflecting wall. The focal length p is generally chosen to be a quarter guided wavelength λ_g at the center frequency for matching purpose [45]. However, by varying p to be larger than a quarter but less than a half guided wavelength λ_g at the center frequency helps to widen the bandwidth [141]. Figure 3.11 shows the effects of the distance p on the VSWR of the launcher. It is seen that the operating frequency ranges from 7 to 15.8 GHz for VSWR less than 2 when $p = 0.25\lambda_g$. By increasing the distance p to be $0.4\lambda_g$, the frequency band from 6 to 18 GHz for VSWR ≤ 2 can be achieved. When p is larger than $0.5\lambda_g$, the impedance matching deteriorates.

A modified parabolic reflecting wall is employed to transform the cylindrical waves excited by the probe to quasi-plane waves at the aperture of the parallel plate waveguide. The shape of reflecting wall is optimized by carefully choosing the value n properly. It is well known that based on the geometric optics, when a source is positioned at the focal point and excited unidirectionally towards a large parabolic reflector, the reflected wave is planar [142]-[144]. However, different from the ideal case, a coaxial probe in our proposed structure excites an omnidirectional cylindrical wave rather than a unidirectional one. Therefore, there exists wave superposition of the reflected wave and the wave transmitted directly from the probe in the launcher, leading to an unavoidable phase distortion at the aperture of the parallel plate waveguide. Due to the very compact size of our launcher, using geometrical optics is insufficient to analyze the phase distortion caused by the wave superposition of the structure. To illustrate the importance of the shape of the reflecting wall, we plot the effects of n on the phase and magnitude errors provided by the launcher with the aid of HFSS, as shown in Figure 3.12.

Figure 3.12(a) plots the differences between the maximum and minimum phase values of the electric field distribution along aperture width of the parallel plate waveguide. It is seen that the phase differences are large for the ideally parabolic shape with $n = 0.5$. However, the phase distortion can be well improved by varying n properly. When n approaches 0.55, the phase differences can be maintained below 20° within the whole frequency band. Figure 3.12(b) shows the corresponding ratios of the maximum to minimum magnitude values as a function of n . It is noticed that the magnitude ratios follow almost the same variations as the phase difference, and the values are below 1.2 at different frequencies when $n = 0.55$. Therefore, it is concluded that quasi-plane waves are obtained in our proposed launcher for the case of $n = 0.55$ considering both the phase and magnitude errors. It is worth noting that choosing a certain value of n may not account for the lowest phase and magnitude differences at all frequencies. Therefore, an appropriate value of n should be chosen to yield a good compromise of phase and magnitude differences over the entire band.

In order to illustrate the importance of the shape of reflecting wall more clearly, the tangential electric field distributions on the ground plane of the proposed surface wave launcher with $n = 0.5$ and 0.55 at different frequencies are plotted in Figure 3.13. It is seen that when $n = 0.5$, the field distributions at high frequencies are distorted. While for the case of $n = 0.55$, the cylindrical electromagnetic waves excited by the coaxial probe are effectively transformed into quasi-plane waves over the entire frequency band. In this case, the phase differences of the waves at the aperture of the parallel-plate waveguide are fairly small.

Based on the above discussions, the following procedures may be considered to design the proposed wideband and low-profile surface wave launcher.

- 1) Choose a substrate with the same dielectric constant ϵ_r as the insulating material in the feeding probe and the same height as the tapered grounded dielectric slab.
- 2) As an initial start, use the ideally parabolic wall with $n = 0.5$. The coaxial probe is fed at the focal point with $p = \lambda_g/4$, where λ_g is the guided wavelength at the center frequency f_0 . The focal length p can be larger than $\lambda_g/4$, but less than $\lambda_g/2$ to maximize the bandwidth. The parabolic shape may also be optimized by studying the variation of the phase and magnitude errors at the waveguide aperture at different frequencies.
- 3) Employ a metallic post to improve the impedance matching.

A few iterations may be invoked to obtain an optimum design. The final dimensions of our proposed surface wave launcher are as follows: $\epsilon_r = 2.1$, $h = 3$ mm, $n = 0.55$, $p = 6.8$ mm, $h_f = 2.85$ mm, $W = 29.1$ mm, $d_p = 2.4$ mm, $h_p = 1.7$ mm, $C_1 = 17.8$ mm, $C_2 = 30$ mm. The diameters of the coaxial probe and metallic post are both 1.3 mm.

Figure 3.12: Simulated (a) phase differences between the maximum and minimum phase values and (b) magnitude ratios of the maximum to minimum magnitude values of the electric field along the aperture width of the waveguide as a function of n at different frequencies ($\epsilon_r = 2.1$, $h = 3$ mm, $W = 29.1$ mm, $p = 6.8$ mm, $h_f = 2.85$ mm, $d_p = 2.4$ mm, $h_p = 1.7$ mm).

Figure 3.13: Tangential E-field distributions of surface wave launcher on the ground plane with $n = 0.5$ and 0.55 at different frequencies, (a), (b) 6 GHz, (c), (d) 12 GHz, (e), (f) 18 GHz.

3.3.2.3 Tapered Impedance Transition

In order to realize a smooth impedance transition between the Teflon slab and ceramic slab, one end of each slab is tapered, and the two parts are seamlessly connected, as shown in Figure 3.14. ϵ_1 and ϵ_2 represent the dielectric constants of

Teflon and ceramic materials, respectively. h_1 and h_2 denote the corresponding dielectric heights. L_{tr} is the length of the transition part.

Figure 3.14: Side view of the tapered impedance transition.

The propagation phase constant in the tapered transmission line can be approximately expressed as [145], [146]

$$\beta_g = \frac{2\pi}{\lambda_0} \sqrt{\varepsilon_a} \quad (3.26)$$

where

$$\varepsilon_a = \frac{\varepsilon_1 h_1 + \varepsilon_2 h_2}{h_1 + h_2} \quad (3.27)$$

is the volume average dielectric constant in the transition, λ_0 is the wavelength in free space.

The electrical length of the tapered transmission line is derived as

$$\phi = \int_0^{L_{tr}} \beta_g dy = \frac{4\pi(\varepsilon_2^{1.5} - \varepsilon_1^{1.5})}{3\lambda_0(\varepsilon_2 - \varepsilon_1)} L_{tr} \quad (3.28)$$

For the tapered transmission line, when the electrical length ϕ is greater than 3π , the reflection coefficient is quite small [147]. In our proposed structure, the center frequency f_0 is used to determine the transition length L_{tr} . Based on (3.28), L_{tr} is calculated to be 10.5 mm if ϕ is equal to 3π at $f_0 = 12$ GHz. Therefore, the length is selected to be $L_{tr} = 10.5$ mm in order to obtain a good matching as well as to maintain a relatively short length for the sake of compactness.

3.3.3 Experimental Verification

A prototype of our proposed surface wave antenna is shown in Figure 3.15. A ceramic material with a high dielectric constant of $\epsilon_r = 25$ and loss tangent of $\tan\delta = 0.0002$ is used for the grounded tapered radiator. The surface wave launcher is fabricated using a Teflon sheet. The Teflon and ceramic parts are seamlessly glued together and covered by copper tapes on both top and bottom sides of the antenna. The reflecting wall of the Teflon part is also covered with copper tape. The dimensions of our proposed antenna are designed to be: $n = 0.55$, $p = 6.8$, $L = 80.3$ mm, $W = 29.1$ mm, $C_1 = 16.7$ mm, $C_2 = 13.2$ mm, $C_3 = 12.2$ mm, $d_p = 2.4$ mm, $h_f = 2.85$ mm, $h_p = 1.7$ mm, $L_{tr} = 10.5$ mm, $h_A = 3$ mm, $h_B = 2.1$ mm, $h_C = 1$ mm, $h_D = 0.2$ mm, $L_{AB} = 11$ mm, $L_{BC} = 15.5$ mm, $L_{CD} = 20.5$ mm.

Figure 3.15: Prototype of the proposed surface wave antenna.

Figure 3.16 shows the simulated and measured VSWR results of the proposed surface wave antenna. It is seen that a wide bandwidth is achieved. The VSWR is less than 2 within the frequency range from 6.1 to 18 GHz for the simulated result and 6.3 to 18.8 GHz for the measured one. The discrepancy may be due to the fabrication and assembling tolerances and small air gaps between the Teflon and grounded ceramic slab in our prototype. In addition, if the conductor tape adhesive is not conductive enough, there exists a very thin layer of substrate between the ceramic slab and the ground plane, which can also attribute to the difference.

Figure 3.16: Simulated and measured VSWR of the proposed surface wave antenna.

Radiation patterns of the proposed antenna are measured in the anechoic chamber, as depicted in Figure 3.17. The results show that the simulated and measured patterns of both E- and H-plane are in good agreement. The radiation beam tilt-angles from the end-fire direction over the operation bandwidth are shown in Figure 3.18. It is seen that the beam pointing angles are about 45° away from the end-fire direction over the frequency range due to the existence of the finite ground. As discussed in Section 3.3.2, the finite ground has a significant influence on the radiation beam angle. The ratio of the ground length L_g that starts from the resonant point where the radiation occurs till the end of the slab to the free-space wavelength is almost identical at different frequencies. Therefore, the beam angles are almost stable since the ground effect is almost the same over the entire band.

The maximum gain values of the proposed surface wave antenna are also plotted in Figure 3.18. The simulated gain is 5.1 dBi at 6 GHz and increases to 12.8 dBi at 18 GHz. The measured gain values are slightly less than simulated ones and the gain discrepancies are within the range of 0.5 dB to 1 dB over the entire frequency band. It is expected that the insertion loss of the SMA connector

Figure 3.17: Simulated and measured radiation patterns of the proposed surface wave antenna at different frequencies.

Figure 3.18: Simulated and measured maximum gain values and beam tilt-angles from the end-fire direction of the proposed surface wave antenna.

is around 0.2 dB. The actual loss tangent of the commercial ceramic material is slightly larger than the one used in simulations, which also results in around 0.2 dB gain degradation. The other unpredictably slight discrepancy may be due to the fabrication and assembling tolerances, and the alignment error of our measurement setup. It should be mentioned that the radiation efficiency of the proposed antenna is above 88% over the entire operation bandwidth, which is calculated using G/D , where G denotes the gain and D is the directivity of the antenna.

In addition, it is worth noting that a larger ground plane can tilt the beam angle closer to the end-fire direction. If the ground plane is infinitely large, an ideally end-fire radiation pattern may be obtained. In order to examine the effects of the ground size on the antenna's radiation pattern, the ground plane of the proposed antenna is extended beyond the end of the slab along the +y direction, and the effect of different ground sizes on the radiation pattern is illustrated in Figure 3.19. It is seen that when the extended ground length increases from 1λ to 10λ ,

Figure 3.19: Simulated radiation patterns of the proposed surface wave antenna with different lengths of an extended ground plane.

Figure 3.20: Phase center variation of the proposed surface wave antenna in the symmetric plane of the structure at different frequencies.

Figure 3.21: Group delay of the proposed surface wave antenna.

the beam angle of the patterns in the E-plane can be gradually tilted from around 45° to 30° away from the end-fire direction.

Furthermore, the phase center of the proposed wideband surface wave antenna varies in the symmetric plane of the structure at different frequencies. As shown in Figure 3.20, the phase center moves closer to the antenna surface with the increase of the frequency. At higher frequencies, the position is relatively stable along the $+z$ direction, which is around 10 mm above the antenna surface. It is also seen that the phase center moves to a thinner position of the tapered grounded slab along the $+y$ direction as the frequency increases. Moreover, Figure 3.21 shows the group delay of the proposed surface wave antenna calculated by HFSS using two identical proposed antennas, which are placed at a distance of 1.5 meters away with the maximum radiation beam facing each other. It is seen that the group delay is within 1 ns over the whole operating bandwidth.

Table 3.2: Performance comparison of flush-mountable end-fire antennas

Reference	Bandwidth (VSWR $\leq$ 2)	Dimensions (λ_0) ($L \times W \times H$, λ_0 is the wavelength at the center frequency)	Volume (λ_0^3)	Gain at the center frequency (dBi)
[45]	19%	$1.56 \times 1.26 \times 0.1$	0.19	7.1
[70]	76%	$6.7 \times 3.8 \times 0.25$	6.365	7.8
[91]	10%	$6.7 \times 11.5 \times 0.17$	13.1	-
[92]	66.7%	$10 \times 4.8 \times 0.29$	13.92	10
[93]	120%	$9 \times 4.3 \times 0.32$	12.38	9.5
Our work	100%	$3.2 \times 1.16 \times 0.12$	0.45	9

Table 3.2 summarizes the performance comparison of several flush-mountable end-fire antennas. It is seen that although the antenna in [45] can maintain a slightly lower thickness and a compact size than our proposed antenna, the bandwidth is significantly narrower. Meanwhile, the design in [93] can achieve a wider bandwidth, but the size is big, and the thickness is much larger. Therefore,

our proposed antenna has the advantages of wide bandwidth, low profile as well as compact size over other flush-mountable end-fire antennas.

3.4 Summary

In this chapter, the characteristics of surface waves propagating on a grounded anisotropic dielectric slab have been investigated by taking into account the field variations along the direction transverse to the propagation direction. The propagation characteristics of surface waves supported by the grounded anisotropic slab have been analyzed in terms of the slab thickness, slab width, as well as the relative permittivity tensor of the anisotropic slab. Furthermore, different methods have been employed to compare the analyses as well as to validate our derivations. Distinct from the existing analyses which generally assume that the fields of surface waves uniformly distribute along the transverse direction of the infinitely large grounded slab, our method is more suitable for practical engineering applications.

A wideband, low-profile and compact surface wave antenna has also been proposed in this chapter. A wideband and low-profile unidirectional surface wave launcher has been realized by employing a probe-fed parallel-plate waveguide with a modified parabolic reflecting wall and a metallic post. A tapered grounded ceramic slab with high dielectric constant has been employed to radiate the surface wave over a wide frequency band as well as to maintain a low profile. The power of the surface waves inside and outside the dielectric slab has been analyzed based on our derived formulas. The guidelines for designing the proposed antenna have also been summarized. The proposed surface wave antenna can achieve a fractional bandwidth of 100% while maintaining an antenna thickness of only $0.12\lambda_0$. It is worth mentioning that the proposed antenna is potentially useful to be conformal and flush-mounted for airborne applications.

CHAPTER 4

WIDEBAND H-PLANE HORN ANTENNA

4.1 Introduction

In Chapter 3, we proposed a compact surface wave antenna employing a tapered grounded slab with high dielectric constant, which can achieve a wide frequency range from 6.1 to 18 GHz, and maintain the antenna thickness of only 3 mm. Nevertheless, it is challenging to further extend the antenna bandwidth since the higher-order surface modes in the grounded slab may radiate as the slab thickness varies, which can destroy the radiation pattern at higher frequencies.

A lot of other efforts have been made to explore suitable candidates for conformal end-fire antennas with a wider bandwidth. In [148], a balanced Vivaldi antenna element was designed to be mounted on a large cylindrical platform, achieving end-fire radiation over a wide bandwidth from 5 to 20 GHz. Unfortunately, its limitation is that the antenna performance will significantly deteriorate if the structure is placed very close to a conducting platform. In [92], a flared slot antenna can achieve a wide frequency band of 4:1. However, the antenna thickness is relatively thick. In [149], a low-profile log-periodic monopole array was proposed by employing the top-hat loading technique, achieving nearly end-fire radiation over a wide bandwidth of 4.53:1 while maintaining a very thin antenna profile, but the monopole array protrudes from the surface of platform, resulting in unnecessarily aerodynamic drag.

H-plane horn antenna is another excellent candidate to meet the requirements of low profile and end-fire radiation and can be easily extended to array configurations. It is well known that simultaneously achieving wide bandwidth and maintaining a low profile is a challenging task for H-plane horn antennas. Ridged waveguide and cone-shaped probe are exploited in the literature to enhance the bandwidth as well as to improve the impedance matching. In [72], a conformal wideband and low-profile H-plane horn antenna employing step

ridged substrate integrated waveguide was proposed to be mounted on a large conducting cylinder, which can cover the bandwidth from 6.1 to 19 GHz. However, further widening the bandwidth may introduce higher-order modes in the waveguide, which will result in high side-lobe levels of the radiation patterns at higher frequencies. Furthermore, although the antenna is low-profile, it is not embedded into the surface of the moving platform, which is not ideal considering the aerodynamic requirement.

In this chapter, both planar and conformal versions of a wideband and low-profile H-plane horn antenna are proposed. The conformal proposed antenna can be well embedded into the surface of the platform. Besides employing a tapered ridge and cone-shaped feeding probe to enhance the bandwidth and to improve the impedance matching, our proposed antenna also utilizes a pair of metallic posts in the coaxial-to-waveguide transition to suppress undesirable higher-order modes. In addition, an ellipse-shaped copper taper is extended along the horn aperture to obtain a smooth transition from the aperture to free space. A tapered grounded substrate is employed at the end of the proposed conformal structure so that the radiation can be gradually tilted away from the end-fire direction, making the proposed structure suitable to be fully embedded into the conducting cylinder.

Compared to existing designs in the literature, our proposed antenna has the following advantages. 1) The antenna can simultaneously achieve a very wide bandwidth and low profile. 2) Due to the suppression of higher-order modes, the antenna can obtain nearly end-fire radiation patterns over the entire bandwidth. 3) The proposed conformal antenna can be fully embedded into the cylindrical conductor so that the aerodynamic drag is minimized for moving platforms.

Figure 4.1: Configuration of the proposed planar wideband H-plane horn antenna, (a) perspective view, (b) top view, (c) left view.

4.2 Planar Wideband H-Plane Horn Antenna

4.2.1 Antenna Configuration

Figure 4.1 shows the configuration of the proposed H-plane horn antenna. It mainly consists of three parts: a coaxial-to-ridged waveguide transition, an H-plane horn with an ellipse-shaped copper extension along the aperture, and a large ground plane. The structure is implemented using two layers of substrates: the bottom layer is Teflon with a dielectric constant of ϵ_{r1} and thickness of h_1 , the top one is Rogers 5880 with a dielectric constant of ϵ_{r2} and thickness of h_2 . The total length and width of the proposed antenna are denoted by L and W , respectively. L_g is the length of the ground plane starting from the right side of the substrate to the ground edge.

A metallic ridge is employed in the coaxial-to-waveguide transition to enhance the bandwidth. The ridge is gradually tapered in order to get a smooth transition from waveguide to horn aperture. A cone-shaped feeding probe is employed in the waveguide and in direct contact with the ridge in order to improve the matching condition. A pair of metallic posts P_1 and P_2 are located at suitable positions along both sides of the feeding probe, which are essential for suppressing the unwanted higher-order mode so that good radiation patterns can be obtained over the entire bandwidth.

In order to smoothen the transition from the horn aperture to free space, an ellipse shape copper is printed on the top layer of the Rogers 5880 substrate, and extended along the horn aperture. Besides, another pair of metallic posts P_3 and P_4 are also employed in this part to improve the matching condition.

4.2.2 Design of the Proposed H-Plane Horn Antenna

In this section, the coaxial-to-waveguide transition is carefully studied first. Several techniques are employed in the transition in order to enhance the operation bandwidth as well as to suppress higher-order modes in the waveguide. After that, an H-plane horn antenna is designed based on the proposed wideband and low-profile transition. The analysis is conducted with the aid of HFSS.

4.2.2.1 Coaxial-to-Ridged Waveguide Transition

The coaxial-to-waveguide transition plays a crucial part in designing wideband and low-profile H-plane horn antenna. It is well known that it is difficult to simultaneously obtain a good impedance matching and maintain a low profile for waveguide structures. It is also a challenge to achieve a bandwidth over 3:1 for low-profile H-plane horn antennas. The main constraint is due to the fact that TE_{30} and other higher-order modes may appear as the frequency increases, which can degrade the radiation pattern. Based on these discussions, the following considerations should be seriously taken into account. First, a broad operating band of the transition should be achieved while maintaining a very low-profile structure. Second, higher-order modes of the narrow rectangular waveguide

In order to meet these requirements, we propose a wideband coaxial-to-waveguide transition, which is shown in Figure 4.2. A tapered ridge with a width of W_r and length of L_r is employed in this structure to enhance the bandwidth as well as to smoothen the transition from the waveguide to horn aperture. A cone-shaped feeding probe is in direct contact with the ridge to improve the matching condition. The upper and lower diameters of the cone are d_t and d_b , respectively. The distance between the feeding probe and the reflecting wall p is selected to be slightly larger than a quarter guided wavelength λ_g at the center frequency to widen the bandwidth [150]. A cylindrical air cavity is perforated in the Teflon

substrate to support the cone-shaped probe considering the actual implementation. It is known that a substrate with low dielectric constant is also beneficial for designing wideband structure, so we employ a Teflon substrate with relative dielectric constant of $\epsilon_{r1} = 2.1$ and height of h_1 in our design. A very thin Rogers 5880 substrate with $\epsilon_{r2} = 2.2$ and height of h_2 is placed atop of the Teflon because it has a close permittivity with Teflon and it is easy to fabricate the ellipse-shaped copper extension along the horn aperture using the printed circuit board technology.

The height of the rectangular waveguide is so chosen as to suppress TE_{n1} mode ($n = 1, 3, 5 \dots$), so that $(h_1 + h_2)$ can be selected to be less than half guided wavelength at the highest frequency, which is

$$h_1 + h_2 < \frac{\lambda_h}{2\sqrt{\epsilon_r}} \quad (4.1)$$

where λ_h is the free space wavelength at the highest operating frequency. ϵ_r is the dielectric constant of the substrate. It should be mentioned that although the waveguide consists of two substrate layers, we can still use $\epsilon_r = \epsilon_{r1}$ for the initial approximation, because the dielectric constants of Rogers 5880 and Teflon are very close, and the thickness of Rogers 5880 h_1 is selected to be much thinner compared to the height h_2 .

The width of the waveguide W_1 of the proposed transition can be designed to suppress the second higher-order TE_{50} mode instead of the first higher-order TE_{30} mode, which can be expressed as

$$W_1 < \frac{5\lambda_h}{2\sqrt{\epsilon_{r1}}} \quad (4.2)$$

where λ_h is the free space wavelength at the highest operating frequency. It is worth mentioning that the first higher-order TE_{30} mode can be effectively suppressed by employing metallic posts at suitable positions of the waveguide, which will be discussed in detail later.

To obtain a smooth transition from the ridged waveguide to the horn aperture,

the ridge is tapered, and the width of the waveguide gradually increases from W_1 to W_2 . It should be noted that W_2 is also the width of the horn aperture. The length of the ridge L_r and aperture width W_2 can be designed to satisfy the following relationship so that the phase distribution of the electric field at the H-plane horn aperture is nearly uniform, which helps to obtain an optimal gain of the antenna [2].

$$W_2 \approx \sqrt{3\lambda_g L_r} \quad (4.3)$$

where λ_g is the guided wavelength at the center frequency.

Figure 4.3: Simulated VSWR of the proposed coaxial-to-ridged waveguide transition without and with tapering the feeding probe ($h_1 = 5$ mm, $h_2 = 0.508$ mm, $W_1 = 25$ mm, $W_2 = 57.5$ mm, $l_1 = 21.6$ mm, $l_2 = 40.2$ mm, $r_1 = 9$ mm, $W_r = 6.7$ mm, $L_r = 55.4$ mm, $p = 5.8$ mm, $t_1 = 6.8$ mm, $t_2 = 1.3$ mm, $d_1 = 1.3$ mm, $r_2 = 2.2$ mm, $d_r = 6.35$ mm, $h_r = 4.408$ mm).

The feeding probe that is extended from the inner conductor of a coaxial line is in direct contact with the ridge and tapered to a cone to improve the impedance matching since it can be used as a stub transformer to achieve a smooth transition from the coaxial line to ridged waveguide. Figure 4.3 shows the simulated VSWR

of the proposed structure with and without the taper. In this comparison, the upper and lower diameters of the tapered cone are $d_t = 2.2$ mm and $d_b = 1.3$ mm, respectively. The other dimensions remain the same. It is seen that the VSWR value with the taper is much lower than that without taper over almost the entire band, indicating that the taper effectively improves the impedance matching condition from the coaxial line to ridged waveguide. By introducing the cone-shaped probe, the VSWR is less than 2 within the frequency band of 3 to 18 GHz.

As mentioned earlier, the suppression of higher-order modes is essential for designing a good horn antenna. Although the proposed structure can achieve a wideband width of 6:1, it only inhibits the excitation of TE_{n1} , TE_{50} and their higher-order modes using (4.1) and (4.2). Therefore, TE_{30} mode can still be excited. In order to suppress TE_{30} mode, a pair of metallic posts P_1 and P_2 are employed at positions where the field of TE_{30} mode can be enforced to zero due to the existence of the metallic post. As shown in Figure 4.2(a), the diameter of the post is represented by d_1 , two posts are symmetrically located on each side of the cone-shaped probe with a distance of t_1 in the x direction. The distance between the posts and the probe is t_2 in the y direction. It is obvious that the inhibition of TE_{30} mode is the most effective when the posts are placed at $t_1 = W_1/3$, where the maximum fields of TE_{30} mode occur. On the other hand, it should be mentioned that the metallic posts also serve as inductors, which can affect the characteristic impedance of the structure. For example, when the posts move closer to the ridge in the y direction, the strong coupling between the posts and the ridge together with the inductance of the posts will introduce a new resonance, which can destroy the matching condition of the proposed transition. Therefore, the position of the posts should be carefully adjusted considering both suppression of TE_{30} mode and impedance matching. Based on these, in our design, the posts move a small distance away from the coaxial probe towards the reflecting wall in the y direction, and the distance t_1 is selected to be around $W_1/3.6$.

Figure 4.4: E-field distributions of the proposed transition at different frequencies, (a)-(c) without posts, (d)-(f) with posts. ($h_1 = 5 \text{ mm}$, $h_2 = 0.508 \text{ mm}$, $W_1 = 25 \text{ mm}$, $W_2 = 57.5 \text{ mm}$, $l_1 = 21.6 \text{ mm}$, $l_2 = 40.2 \text{ mm}$, $r_1 = 9 \text{ mm}$, $W_r = 6.7 \text{ mm}$, $L_r = 55.4 \text{ mm}$, $p = 5.8 \text{ mm}$, $d_1 = 1.3 \text{ mm}$, $r_2 = 2.2 \text{ mm}$, $d_r = 6.35 \text{ mm}$, $h_r = 4.408 \text{ mm}$, $d_b = 1.3 \text{ mm}$, $d_t = 2.2 \text{ mm}$, $t_1 = 6.8 \text{ mm}$, $t_2 = 1.3 \text{ mm}$.)

The significance of the posts for suppressing TE₃₀ mode can be illustrated by examining the E-field distributions of the proposed transition at different frequencies, as shown in Figure 4.4. At lower frequencies, only TE₁₀ mode is excited in the waveguide, so that the E-field distributions are similar for the transition with and without the posts, as displayed in Figures 4.4(a) and (d). At higher frequencies, it is obvious that in Figures 4.4(b) and (c), TE₃₀ mode appears for the case without the posts. However, in Figures 4.4(e) and (f), after introducing the posts, TE₃₀ mode is effectively suppressed, so that only the dominant mode is propagating over the entire bandwidth from 3 to 18 GHz.

Based on the above discussions, the guideline of designing the proposed wideband coaxial-to-waveguide transition is summarized.

- 1) Choose two layers of substrates with close and low dielectric constants. The top substrate can be selected to be very thin compared to the bottom one. Determine the height (h_1+h_2) and width of the waveguide W_1 based on the highest operating frequency using (4.1) and (4.2).
- 2) As an initial start, the coaxial probe is fed with the distance $p = \lambda_g/4$ away from the reflecting wall, where λ_g is the guided wavelength at the center frequency. The distance p can be slightly larger than $\lambda_g/4$ but less than $\lambda_g/2$ to maximum the bandwidth.
- 3) Employ a ridge in the waveguide to decrease the cutoff frequency of TE₁₀ mode. The width of the ridge W_r and its height h_r can then be initially determined when the cutoff frequency of TE₁₀ mode is below the required lowest operating frequency. The ridge is gradually tapered to smoothen the transition from the waveguide to horn aperture. The feeding probe is also tapered to improve the impedance matching. The width of the horn aperture can be determined based on (4.3).
- 4) A pair of metallic posts is placed at each side of the feeding probe with a distance of around $W_1/3$ initially, which may be further tuned considering both the suppression of TE₃₀ mode and the impedance matching.

A few iterations may be invoked to obtain an optimum design. In our design, the target frequency is from 3-18 GHz, and the optimized dimensions are shown in the caption of Figure 4.4.

4.2.2.2 H-Plane Horn Antenna

An ellipse-shaped copper is printed on the top layer of the Rogers 5880 substrate and extended along the horn aperture. As shown in Figure 4.5, the center point of the ellipse is O' , which has a distance of d_s away from the horn aperture. The half major and minor axes of the ellipse are represented by a_1 and a_2 , respectively. The ratio of the major axis and minor axis is around 1.5 in order to smoothen the transition from the horn aperture to free space over a wide frequency range. In addition, the printed copper also serves as a wave-guiding structure so that the wave can be guided along the forward direction, which helps to improve the front-to-back ratio of the radiation pattern.

Figure 4.5: The geometry of the proposed H-plane horn antenna from the top view.

Metallic posts P_3 and P_4 are utilized in the proposed H-plane horn antenna in order to improve the impedance matching because the posts serve as inductances

and the corresponding values are related to their diameters and locations. As shown in Figure 4.5, two posts with a diameter of d_2 are located at a distance of t_3 with respect to the symmetric plane of the antenna, the distance from the horn aperture to the posts along the y direction is denoted by t_4 . In order to illustrate the importance of these posts, we compare the simulated VSWR value of the proposed H-plane horn antenna with and without posts using the planar structure, as shown in Figure 4.6. It is seen that the VSWR is high especially at lower frequencies for the case without posts. However, when posts are employed at appropriate positions, the value of VSWR can maintain below 2.5 from 3.3 to 18 GHz.

Figure 4.6: Simulated VSWR of the proposed planar H-plane horn antenna with and without posts ($h_1 = 5$ mm, $h_2 = 0.508$ mm, $L = 267$ mm, $W = 100$ mm, $L_g = 110$ mm, $W_1 = 25$ mm, $W_2 = 56.5$ mm, $l_1 = 21.6$ mm, $l_2 = 40.2$ mm, $r_1 = 9$ mm, $W_r = 5.6$ mm, $L_r = 55.45$ mm, $p = 5.8$ mm, $t_1 = 6.8$ mm, $t_2 = 1.3$ mm, $d_1 = 1.3$ mm, $r_2 = 1$ mm, $d_r = 6.35$ mm, $h_r = 4.408$ mm, $d_b = 1.3$ mm, $d_t = 2.4$ mm, $d_s = 20.1$ mm, $a_1 = 46.5$ mm, $a_2 = 31$ mm, $d_2 = 2$ mm, $t_3 = 24.5$ mm, $t_4 = 15$ mm).

Figure 4.7: Effects of the ground plane on radiation patterns of the proposed antenna at different frequencies.

In addition, due to the finite length of the ground plane, the radiation pattern of the proposed antenna is tilted a certain angle away from the end-fire direction. In order to examine the influence of the ground on the antenna's pattern, the ground plane of the proposed antenna is extended beyond the end of the slab along the +y direction, and the effects of different length L_g on radiation pattern are illustrated in Figure 4.7. It is worth mentioning that the E-plane is the y-z plane, and the H-plane is orthogonal to the E-plane and passes through the maximum radiation pattern. It is clearly seen that when the extended ground length increases from 10 to 110 mm, the beam angle of the radiation patterns in the E-plane can be gradually tilted closer to the end-fire direction. It is expected that an end-fire radiation may be obtained for an infinitely large ground plane.

Figure 4.8: Phase center variation of the proposed planar H-plane horn antenna with the extended ground length $L_g = 10$ mm.

Furthermore, the position of the phase center varies in the symmetric plane of the proposed H-plane horn antenna at different frequencies. For the antenna structure with the extended ground length $L_g = 10$ mm, the phase center variation is plotted in Figure 4.8. It is seen that the phase center is from 98 mm to 116 mm away from the feeding point along the +y direction. In addition, it moves closer

to the ground plane as the frequency increases. The simulated group delay of the proposed horn antenna is shown in Figure 4.9. It is seen that the group delay is stable over the whole operating bandwidth.

Figure 4.9: Group delay of the proposed planar H-plane horn antenna with the extended ground length $L_g = 10$ mm.

4.2.3 Experimental Verification

The prototype of the proposed H-plane horn antenna is fabricated. As shown in Figure 4.10, the tapered ridge is realized by carving a groove on the substrates and covered with copper tapes. The printed copper on the top layer of the Rogers 5880 is fabricated using the printed circuit board technology. The Teflon and Rogers substrates are seamlessly glued together. In addition, the side walls of the transition part are also covered with copper tapes. The H-plane horn antenna is mounted on a large aluminum ground plane. The optimized parameters of the proposed conformal antenna are as follows: $h_1 = 5$ mm, $h_2 = 0.508$ mm, $L = 267$ mm, $W = 100$ mm, $L_g = 110$ mm, $W_1 = 25$ mm, $W_2 = 56.5$ mm, $l_1 = 21.6$ mm, $l_2 = 40.2$ mm, $r_1 = 9$ mm, $W_r = 5.6$ mm, $L_r = 55.45$ mm, $p = 5.8$ mm, $t_1 = 6.8$ mm, $t_2 = 1.3$ mm, $d_1 = 1.3$ mm, $r_2 = 1$ mm, $d_r = 6.35$ mm, $h_r = 4.408$ mm, $d_b = 1.3$ mm,

$d_t = 2.4$ mm, $d_s = 20.1$ mm, $a_1 = 46.5$ mm, $a_2 = 31$ mm, $d_2 = 2$ mm, $t_3 = 24.5$ mm, $t_4 = 15$ mm.

Figure 4.10: Prototype of the proposed wideband H-plane horn antenna.

Figure 4.11: Simulated and measured VSWR of the proposed antenna.

Figure 4.12: Simulated and measured radiation patterns of the proposed H-plane horn antenna at different frequencies.

Figure 4.13: Simulated and measured maximum gain values and beam tilt-angles from the end-fire direction of the proposed H-plane horn antenna.

Figure 4.11 shows the simulated and measured VSWR results of the proposed antenna. It is seen that the simulated and measured results agree very well, the VSWR below 2.5 can cover from 3.3 to 18.2 GHz and 3.3 to 17.9 GHz, respectively. A slight discrepancy at higher frequencies may be due to the fabrication and assembling tolerances.

Radiation patterns of the proposed H-plane horn antenna are measured in the anechoic chamber. The simulated and measured radiation patterns at different frequencies are also compared in Figure 4.12. It is seen that both results are in good agreement. In addition, due to the effective suppression of higher-order modes in the waveguide, the side-lobe levels are acceptable over the entire band, and no beam split is observed. Figure 4.13 shows the simulated and measured maximum gain values and beam tilt-angles from the end-fire direction of the proposed H-plane horn antenna. As seen, the corresponding angles tilt less than 40° from the end-fire direction. In addition, the gain value enhances as the frequency increases. The simulated gain is around 2.8 dBi at the lowest operating frequency of 3.2 GHz, and the value increases to be 17.7 dBi at 18 GHz. The measured values are slightly lower than simulated ones, which may be due to the

additional loss introduced by the SMA connector and the alignment error of the measurement setup.

4.3 Conformal Wideband H-Plane Horn Antenna

4.3.1 Antenna Configuration

The antenna configuration of the conformal wideband H-plane horn antenna is similar to the planar version in the previous section, as shown in Figure 4.14. The difference is that the proposed conformal antenna consists of a tapered conductor-backed dielectric slab at the end of the structure. The antenna is mounted on a curved conducting ground, and all side faces of the substrates are covered with metal. The radius of the conformal structure is represented by R . The other parameters are denoted in the same manner as those in the previous section.

Figure 4.14: Configuration of the proposed conformal wideband H-plane horn antenna, (a) perspective view, (b) left view.

From a practical point of view, it is desirable that the antenna can be embedded into the conducting platform so that the surfaces of the antenna and the platform can seamlessly become an integral part. This requires that the performance of the antenna is insensitive to the metal surrounding the structure. However, an H-plane horn antenna radiates along the axial direction, the metal at the right edge of the antenna can seriously destroy the antenna performance as it will block the radiation and cause strong backscattering. In order to solve this problem, the dielectric is gradually tapered from a thickness of (h_1+h_2) to h_t at the end of the structure, and the length of the taper is L_t . By smoothly tapering the conductor-backed substrate, the radiation can be slowly tilted at a small angle away from the end-fire direction. It should be mentioned that the length of the tapered structure L_t cannot be too small in order to smoothen the transition, a large L_t may also lead to a less compact structure. Therefore, L_t should be properly selected so that the antenna can be well embedded into a cylindrical platform.

Figure 4.15: Simulated VSWR of the proposed conformal H-plane horn antenna with different radius of the curvature ($h_1 = 5$ mm, $h_2 = 0.508$ mm, $L = 185.8$ mm, $W = 90$ mm, $W_1 = 25$ mm, $W_2 = 55$ mm, $l_1 = 22$ mm, $l_2 = 39.8$ mm, $r_1 = 9$ mm, $W_r = 5.5$ mm, $L_r = 55.6$ mm, $p = 5.8$ mm, $t_1 = 6.8$ mm, $t_2 = 1.3$ mm, $d_1 = 1.3$ mm, $r_2 = 1.4$ mm, $d_r = 6.25$ mm, $h_r = 4.408$ mm, $d_b = 1.3$ mm, $d_t = 2.3$ mm, $d_s = 20$ mm, $a_1 = 45.5$ mm, $a_2 = 30.5$ mm, $d_2 = 2$ mm, $t_3 = 24.5$ mm, $t_4 = 15$ mm, $L_t = 35$ mm, $h_t = 1$ mm).

4.3.2 Effects of the Conformal Platform

The effect of the radius of the cylindrical platform on the antenna characteristic is shown in Figure 4.15. When the antenna is bent with $R = 50$ mm and $R = 125$ mm, the VSWR values are around 4 and 3 at lower frequencies, respectively, indicating that the matching condition of the proposed antenna deteriorates as the radius decreases. As the radius increases to be 200 mm, the VSWR value can be maintained below 2.6 from 3.4 to 18 GHz, which is close to that of the planar structure. In our design, six proposed conformal antennas should be uniformly arranged and embedded into the surface of the conducting cylinder so that the radiation beams can cover the whole space. Therefore, we choose the curvature of the proposed antenna to be $R = 125$ mm, and it is obvious that its matching condition needs to be further improved by carefully optimizing the parameters of the antenna. The optimized dimension and simulated performance of the conformal antenna will be presented in the next section.

Figure 4.16: Structure of the proposed conformal antenna element embedded into the cylindrical platform.

Figure 4.17: Effects of the platform on radiation patterns of the conformal antenna element at different frequencies.

The platform can also affect the radiation pattern of the proposed conformal antenna because the surface of the platform serves as the ground plane of the antenna when the antenna is embedded in the platform. As shown in Figure 4.16, the conductor platform consists of a cylinder part with the radius of $R = 125$ mm and a cone part with the length of $L_c = 190$ mm. The distance from the right edge of the antenna to the cone is L_b . The radiation patterns at different frequencies of the original conformal antenna without being embedded in the platform are compared with the ones when the antenna is embedded. Simulated results are shown in Figure 4.17. It is seen that the radiation beam angles in the E-plane of the original conformal antenna tilt a bit far away from the end-fire direction especially at low frequency. For example, the angle is around 50° at 4 GHz. This is due to the existence of the tapered substrates and the finite ground plane beneath the antenna. When the antenna is embedded in the platform, the ground plane size of the antenna is enlarged, which can tilt the radiation beam closer to the end-fire direction. As the platform length L_b increases to be 140 mm, the offset beam angle from the end-fire direction is 40° at 4 GHz. When L_b further increases to be infinite, an end-fire radiation pattern may be obtained over the whole operating frequency band. It should be noted that the platform has insignificant effects on the radiation pattern in the H-plane.

Figure 4.18: Structure of the proposed conformal array configuration with six separate embedded antenna elements, (a) perspective view, (b) left view.

(a)

(b)

(c)

Figure 4.19: Radiation beam coverage of the proposed array configuration at different frequencies when each antenna element is excited independently, (a) $f = 4$ GHz, (b) $f = 12$ GHz, (c) $f = 18$ GHz.

Figure 4.18 shows the structure of the proposed conformal array configuration with six separate embedded antenna elements. When each antenna element is excited independently, it is desirable that the radiation beam can cover the whole space. In order to examine the radiation beam coverage of the array configuration, the radiation gain values of each antenna element in terms of both directions (θ , ϕ) are plotted in Figure 4.19. Due to the symmetry of the six-element array configuration, the radiation beam of each antenna element covers every 60° in the ϕ direction sequentially according to the coordinate system in Figure 4.18. It is seen that at low frequency $f = 4$ GHz, the gain values above 5 dBi are achieved within the region $22^\circ \leq \theta \leq 42^\circ$, where the gain ripple is only around 1 dB. In addition, the gain value in the end-fire direction ($\theta = 0^\circ$) is larger than 0 dBi. At higher frequencies at $f = 12$ GHz and $f = 18$ GHz, the radiation beams tilt closer to the end-fire direction and the maximum gain values above 15 dBi are observed within the regions $8^\circ \leq \theta \leq 22.5^\circ$ and $7^\circ \leq \theta \leq 24^\circ$, respectively. Although the gain ripples within these two regions increase compared to the corresponding value in Figure 4.19(a), the radiation gain values over these two regions are still above 5 dBi.

4.3.3 Experimental Verification

The prototype of the designed conformal H-plane horn antenna is fabricated. As shown in Figure 4.20 (a), the four side faces and the bottom side of the proposed conformal antenna are all covered with copper tapes. Figure 4.20(b) shows the details of the surface of the cylindrical platform, a sink with the same shape and height of the conformal antenna is carved out from the platform surface to support the embedded antenna. It should be noted that a tapered substrate is employed at the end of the proposed conformal antenna so that carving this part from the surface of the platform is very difficult because it is hard to control the fabrication accuracy of the curvature and the taper. In this case, we cut a slot to facilitate the fabrication processing. Figure 4.20(c) shows the assembled prototype of the conformal array configuration. Six elements are uniformly embedded into the aluminum cylinder with a radius of 125 mm. For each element, the inner

(a)

(b)

(c)

Figure 4.20: Fabricated prototype of the proposed antenna, (a) conformal antenna element, (b) details of the aluminum platform surface, (c) conformal array configuration with six antenna elements.

conductor of the SMA connector is connected with an aluminum cone, which touches the ridge from the bottom side of the antenna. The aluminum cylinder is also connected to a large aluminum cone on the top side. The total length of the fabricated platform is 520 mm. The optimized parameters of the proposed conformal antenna are as follows: $R = 125$ mm, $h_1 = 5$ mm, $h_2 = 0.508$ mm, $L = 175.8$ mm, $W = 90$ mm, $W_1 = 25$ mm, $W_2 = 55$ mm, $l_1 = 22$ mm, $l_2 = 38.8$ mm, $r_1 = 9$ mm, $W_r = 5.6$ mm, $L_r = 54.5$ mm, $p = 5.8$ mm, $t_1 = 6.8$ mm, $t_2 = 1.3$ mm, $d_1 = 1.3$ mm, $r_2 = 1.2$ mm, $d_r = 6.28$ mm, $h_r = 4.408$ mm, $d_b = 1.3$ mm, $d_t = 2.4$ mm, $d_s = 20.7$ mm, $a_1 = 43.6$ mm, $a_2 = 31.2$ mm, $d_2 = 2$ mm, $t_3 = 24.6$ mm, $t_4 = 14.5$ mm, $L_t = 35$ mm, $h_t = 1$ mm.

Figure 4.21: Simulated and measured VSWR of a single proposed conformal antenna element embedded into the platform.

In the array configuration, the polarization of each antenna element is different because six elements are arranged at different positions of the platform. Due to the coverage of 360° radiation beam in the whole space, the array configuration can receive electromagnetic waves from different directions when each element is excited independently. Therefore, the proposed circular array is very suitable for the direction of arrival (DOA) estimation to determine the direction of the target and it is a good candidate for MIMO applications. Because of the symmetry of the array configuration, we only measure the radiation characteristics

Figure 4.22: Simulated and measured radiation patterns of the proposed conformal antenna element embedded into the platform at different frequencies.

Figure 4.23: Simulated and measured maximum gain values and beam tilt-angles from the end-fire direction of the proposed conformal antenna element embedded into the platform.

Figure 4.24: Measured coupling between each antenna element in the array configuration.

of a single antenna element embedded into the platform. Figure 4.21 shows the simulated and measured VSWR and proposed conformal antenna. At lower frequencies, the VSWR values are a bit high because the profile of the structure is extremely low compared to the corresponding free-space wavelength. However, both simulated and measured VSWR values can still remain below 2.5 from 3.4 to 18 GHz.

Figure 4.22 shows the simulated and measured radiation patterns of a single antenna element embedded into the platform. It is seen that the radiation beams in the E-plane tilt a small angle away from the end-fire direction due to the finite length of the ground plane. Owing to the effective suppression of higher-order modes in the waveguide, the side-lobe levels are acceptable and no beam split is observed. In addition, it is worth mentioning that the cross-polarization of the proposed antenna element at the beam-pointing direction is very low, which is below -25 dB over the entire bandwidth. The simulated and measured maximum gain values and beam tilt-angles of the proposed conformal antenna element embedded into the platform are shown in Figure 4.23. It is seen that the radiation beam tilts closer to the end-fire direction as the frequency increases. The simulated gain is around 3.5 dBi at 3.4 GHz and increases to 16.2 dBi at 18 GHz. The measured gain values are slightly less than simulated ones, which may be attributed to the additional loss introduced by the SMA connector and the alignment error of the measurement setup.

Figure 4.24 shows the measured coupling coefficients between each port of the six antenna elements in the array configuration. Because of the symmetry of the structure, $|S_{21}|$ is equal to $|S_{61}|$, and $|S_{31}|$ is equal to $|S_{51}|$. It is seen that the coupling between each port is almost lower than -40 dB over the entire frequency band, indicating that the couplings from one port to the others are very weak.

Table 4.1 lists the performance comparison of various conformal end-fire antennas available in the literature, where λ_L is the free-space wavelength at the lowest operating frequency. It is seen that although the design in [93] is suitable to be embedded into a conformal platform and has a wide bandwidth, the antenna height is relatively thick. Meanwhile, the antenna in [148] can achieve a wide

bandwidth of 4:1 and maintain a very low profile, the ground plane has a significant effect on the antenna performance and the structure cannot be embedded into the platform. Compared to the existing conformal end-fire antennas, our proposed design has the advantages of wide bandwidth and low profile while possessing the ability to be embedded in conducting platforms.

Table 4.1: Performance comparison of various conformal end-fire antennas

Ref.	VS WR	Bandwidth	Height (λ_L)	Gain at the center frequency (dBi)	Structure
[72]	<2.5	3.1:1	0.065	11.6	Non-embedded
[74]	<2.5	2.1:1	0.112	10	Embedded
[92]	<1.4	2:1	0.21	10	Embedded
[93]	<2.6	4:1	0.127	9.5	Embedded
[148]	<2	4:1	0.053	-	Non-embedded
[150]	<2	2.95:1	0.064	9	Non-embedded
Our work	<2.5	5.29:1	0.062	14	Embedded

4.4 Summary

In this chapter, both planar and conformal versions of a wideband and low-profile H-plane horn antenna have been proposed. The guidelines for designing the proposed coaxial-to-waveguide transition have been summarized based on the considerations of wide bandwidth, low profile, suppression of higher-order modes, as well as the phase distribution of the electric field along the horn aperture. Several strategies such as a cone-shaped probe, elliptical printed copper, and metallic posts have been utilized in the H-plane horn to improve the impedance matching. In addition, the conformal version of the proposed antenna is fully embedded in a large cylindrical platform and has been extended to build a conformal array configuration.

It has been shown in the simulated results that both planar and conformal antennas can achieve very a wide bandwidth from 3.4 to 18 GHz for $VSWR \leq 2.5$ while retaining the antenna profile of only 5.508 mm ($0.062\lambda_0$ at the lowest operating frequency). Nearly end-fire radiation has been obtained over the entire frequency band. Prototypes of the planar antenna and conformal array configuration have been fabricated and tested; measured results are in good agreement with simulated ones.

CHAPTER 5

SUPERGAIN SLOT ANTENNA

5.1 Introduction

As reviewed in Chapter 2, most supergain antennas are related to dipole and monopole structures, little has been reported on supergain slot antennas. In this chapter, we propose three supergain slot antennas based on the inverted microstrip-fed cavity-backed structure [151]. One is a planar back-to-back structure, and the other two are conformal to cylindrical and conical platforms, respectively.

In the first proposed configurations, two separate substrates are placed on both the top and bottom walls of a rectangular cavity. Different from the conventional cavity-backed structure where the cavity is employed to eliminate the undesirable bidirectional radiation [152], the proposed antenna utilizes the bidirectional radiation to obtain a strong radiation in the end-fire direction. Three slots are closely etched out on the broad walls of the cavity in parallel. The required magnitude and phase of the excited electric field for achieving supergain can be obtained by adjusting the position of each slot. Slots with appropriate lengths are employed to generate multiple resonances as well as to decrease the Q value of the cavity, which broadens the bandwidth effectively. Furthermore, off-centered microstrip feedlines are utilized to improve the impedance matching and excite the cavity mode. Finally, a compact microstrip balun based on the rat-race coupler is designed and integrated into the side wall of the cavity for measurement convenience and verification purpose.

Based on the first structure, we also propose conformal supergain slot antennas, which operate in the X band and are mounted on cylindrical and conical platforms, respectively. A large ground is utilized to obtain the near end-fire radiation pattern for the cylindrical platform, while an end-fire radiation can be achieved

when the proposed supergain slot antenna is embedded into the surface of a conical platform.

Our proposed structures have the following advantages. 1) The proposed antenna effectively alleviates the limitations of traditional supergain antennas and can achieve a compact size, enhanced bandwidth, high gain, and high radiation efficiency. 2) The proposed conformal antennas are very low-profile and can be fully embedded into metallic platforms. 3) The proposed conformal structure only consists of a substrate and a cavity, which is very simple and low-cost. 4) End-fire radiation is obtained when the proposed antenna is embedded into a conical platform. 5) Six conformal antenna elements are wrapped around the conical platform to form an array configuration. The array structure is immune to electromagnetic waves with arbitrary polarization, which is polarization-independent and potentially very useful for airborne applications.

5.2 Planar Cavity-Backed End-Fire Slot Antenna

5.2.1 Antenna Configuration

The configuration of our proposed antenna is shown in Figure 5.1. Two substrates with a relative permittivity of ϵ_r and thickness of h are employed on both the top and bottom walls of the cavity, whose inner size is denoted by $W_c \times L_c \times H_c$, with the wall thickness of t . Three slots with different lengths and equal width are etched out in parallel on the inner conducting layers of both substrates, facing inward the cavity. The width of each slot is W_s . The lengths of slots 1-3 are represented by L_1 , L_2 , and L_3 , respectively, which are adjusted to achieve multi-resonant frequencies and to decrease the Q value of the cavity, leading to an increase of the operation bandwidth. Two open-ended microstrip feedlines are printed on both outer layers of substrates. Both feedlines are off-centered in order to obtain a good impedance matching. The spacings between slots 1 and 2 and slots 2 and 3 are denoted by d_l and d_r , respectively. The distance from slot 2 to the left side wall of the cavity is represented by d_s . By varying the position of each slot on broadside walls of the metallic cavity, the required magnitude and phase of the electric field along the slot surfaces to achieve supergain can be

obtained. As shown, ports 1 and 2 are fed with equal amplitude and opposite phase to achieve a strong radiation in the end-fire direction.

Figure 5.1: Configuration of an inverted microstrip-fed cavity-backed slot supergain antenna, (a) 3D view, (b) top view, (c) side view.

5.2.2 Antenna Analysis

The proposed structure is analyzed by HFSS. Figures 5.2(a) and (b) show the simulated electric field distribution of the cavity and the corresponding current distribution on the cavity wall, respectively. It is seen that TE_{120} mode is strongly excited at 5.8 GHz in the cavity and the y-directed current flowing on the broad wall of the cavity is disturbed by the slots. In this manner, the transverse electric field across the slot surfaces can be properly excited for the radiation purpose. In order to stimulate TE_{120} mode in the cavity, the microstrip feedlines are placed

across slot 2 (the middle slot). It should be mentioned that TE_{120} mode is perturbed due to the existence of etched slots and microstrip feedlines. The operation bandwidth of the proposed antenna lies between the resonant frequencies of TE_{210} and TE_{120} mode of the cavity, which are 5.35 GHz, and 6.4 GHz, respectively. However, the perturbed TE_{120} mode dominates within the entire bandwidth and is responsible for exciting the slots.

The effects of each slot length on the impedance bandwidth and maximum gain in the end-fire direction are shown in Figure 5.3. It is seen that three resonant frequencies $f_{\text{slot } 1}$, $f_{\text{slot } 2}$, $f_{\text{slot } 3}$ are obtained at around 5.56 GHz, 5.8 GHz and 6 GHz, respectively. Two gain peaks are also displayed near $f_{\text{slot } 1}$ and $f_{\text{slot } 3}$.

Figure 5.2: (a) E-Field distribution of the cavity, (b) current distribution on the broad wall of the cavity ($f = 5.8$ GHz, $H_c = 12$ mm, $W_c = 48$ mm, $L_c = 66$ mm, $W_s = 3$ mm, $L_1 = 23$ mm, $L_2 = 18.3$ mm, $L_3 = 17.3$ mm, $d_s = 25.96$ mm, $d_l = 7$ mm, $d_r = 6.08$ mm).

Figure 5.3(a) implies that when L_1 increases, $f_{\text{slot } 1}$ decreases while $f_{\text{slot } 2}$ and $f_{\text{slot } 3}$ are almost unchanged, which enlarges the impedance bandwidth. Meanwhile, the gain ripple between two peaks increases while the first gain peak near $f_{\text{slot } 1}$ increases and moves to a lower frequency.

(a)

(b)

(c)

Figure 5.3: Effects of slot lengths L_1 , L_2 , and L_3 on $|S_{11}|$ and gain values in the end-fire direction of the proposed antenna ($H_c = 12$ mm, $W_c = 48$ mm, $L_c = 66$ mm, $h = 0.787$ mm, $W_s = 3$ mm, $d_s = 25.96$ mm, $d_l = 7$ mm, $d_r = 6.08$ mm, $t = 5$ mm, $W_f = 2.43$ mm, $a = 12.43$ mm, $b = 4.065$ mm).

Figure 5.3(b) indicates that when L_2 increases, $f_{\text{slot } 2}$ moves to a lower frequency while $f_{\text{slot } 1}$ and $f_{\text{slot } 3}$ remain unaffected, so the impedance bandwidth is hardly changed. The gain peak near $f_{\text{slot } 3}$ increases and the one near $f_{\text{slot } 1}$ decreases with the increase of L_2 , which somehow balances two gain peaks.

Figure 5.3(c) shows that when L_3 increases, $f_{\text{slot } 1}$, and $f_{\text{slot } 2}$ are almost invariable while $f_{\text{slot } 3}$ decreases, which reduces the impedance bandwidth. At the same time, the gain ripple between the two peaks decreases while the second gain peak near $f_{\text{slot } 3}$ increases and moves to a lower frequency.

As seen from the above discussions, there is a tradeoff between bandwidth and gain. Three resonant frequencies are introduced by using three different slot lengths. Slots 1 and 2 and slots 2 and 3 are responsible for the first and second gain peaks, respectively.

Table 5.1: Magnitude and phase differences of the electric field along the back-to-back slot structure at $f_{\text{slot } 1}$ and $f_{\text{slot } 3}$

f	$f_{\text{slot } 1}$			$f_{\text{slot } 3}$		
Slots	Slot 1	Slot 2	Slot 3	Slot 1	Slot 2	Slot 3
Phase	151.3°	0°	-42°	-36.2°	0°	-149.8°
Magnitude	1.1	1	0.45	0.22	1	1.48

Table 5.1 shows the simulated relative values of the magnitude and phase for the electric field along the slot surfaces at $f_{\text{slot } 1}$ and $f_{\text{slot } 3}$, where the gain peaks occur. In order to facilitate the comparison, the corresponding magnitude and phase of slot 2 are normalized to be 1 and 0°, respectively. It is seen that the phase differences of the electric field along the surfaces of slots 1 and 2 and slots 2 and 3 are almost equal, with the values of 151.3° at $f_{\text{slot } 1}$, and 149.8° at $f_{\text{slot } 3}$, respectively. The corresponding magnitude of the first slot pair (slots 1 and 2) is

higher than that of slot 3 at $f_{\text{slot } 1}$. Similarly, the magnitude of the second slot pair (slots 2 and 3) is larger than that of slot 1 at $f_{\text{slot } 3}$.

5.2.3 Experimental Verification

5.2.3.1 Design Example

In order to characterize the radiation performance of the proposed antenna, a prototype is fabricated. In this design, Duroid RT 5880 substrates with permittivity of 2.2 and thickness of 0.787 mm are used to etch the slots and print the feedlines. The cavity is made up of aluminum plates, whose inner dimension is $48 \times 66 \times 12 \text{ mm}^3$ and wall thickness is 5 mm. The lengths of three slots are designed to be $L_1 = 23 \text{ mm}$, $L_2 = 18.3 \text{ mm}$, and $L_3 = 17.3 \text{ mm}$, respectively. The width of each slot is chosen to be $W_s = 3 \text{ mm}$. Interelement spacing values of d_l and d_r are 7 mm ($0.135\lambda_0$) and 6.08 mm ($0.12\lambda_0$), respectively. The other dimensions are as follows: $d_s = 25.96 \text{ mm}$, $W_f = 2.43 \text{ mm}$, $a = 12.43 \text{ mm}$, and $b = 4.065 \text{ mm}$.

Figure 5.4: Prototype of the proposed antenna with integrated rat-race balun, (a) perspective view, (b) details inside the cavity.

A feeding balun is also designed to be integrated with the proposed antenna for measurement convenience, as shown in Figure 5.4. The feeding network is based on a rat-race coupler, which provides the output power with equal amplitude, opposite phase, and good isolation [79]. The rat-race balun is printed

on a Duroid RT 5880 substrate with the dimensions $58 \times 13.574 \times 0.787 \text{ mm}^3$, which can be seamlessly integrated into the side wall of the cavity. It consists of one input, one isolation port, two outputs, and a microstrip ring. All four ports are designed to be $50 \text{ } \Omega$ with the width of 2.34 mm, while the impedance of microstrip ring is $70.7 \text{ } \Omega$. The isolation port is terminated by a $50\text{-}\Omega$ resistor, and the outputs are perpendicularly connected to port 1 and port 2 in Figure 5.1(a) by copper tapes directly.

5.2.3.2 Simulated and Measured Results

S-parameters of the proposed antenna are measured using a calibrated Agilent vector network analyzer N5230A. First, return loss and isolation of the two port antenna are measured without the feeding balun. Two SMA connectors are soldered to the microstrip feedlines for measurement purpose. It should be mentioned that $|S_{11}|$ and $|S_{22}|$ are equal because of the symmetric structure of the antenna. The measured results are compared with simulated ones, and they are shown in Figures 5.5 and 5.6. Figure 5.5 shows that the -10-dB impedance bandwidth ranges from 5.51 to 6.08 GHz for the simulated result and from 5.48 to 6.05 GHz for the measured one, indicating that a fractional bandwidth of 9.8% is obtained. The measured $f_{\text{slot } 1}$ and $f_{\text{slot } 2}$ are slightly lower compared to simulated ones, which may be due to the small fabrication and assembly tolerances. Due to the back-to-back feeding structure sharing the same metallic cavity, the isolation between two ports is not high, as expected. Figure 5.6 shows that a coupling of -10 dB is observed for both simulated and measured results. Next, the return loss is measured after the antenna is integrated with the feeding balun. The impedance matching of the whole antenna is affected by the introduced network, which makes the return loss result different from the one without a balun, as shown in Figure 5.5. However, the bandwidth for 10-dB return loss changes a bit, ranging from 5.51 to 6.07 GHz for the simulated result and 5.45 to 6.0 GHz for the measured one. The measured fractional bandwidth is 10.3%, which is slightly larger than the simulated one. Again, this may be attributed to the fabrication and assembly errors.

Figure 5.5: Measured and simulated $|S_{11}|$ of the proposed back-to-back slot supergain antenna with and without feeding balun.

Figure 5.6: Measured and simulated $|S_{12}|$ of the proposed back-to-back slot supergain antenna without feeding balun.

Figure 5.7: Simulated and measured radiation patterns of the proposed back-to-back slot supergain antenna at $f = 5.8$ GHz, (a) yz -plane, (b) xy -plane.

Radiation performance of the proposed antenna is measured in the anechoic chamber with the feeding balun. Figure 5.7 shows the simulated and measured radiation pattern at 5.8 GHz in the yz -plane and xy -plane. The maximum radiation is obtained in the end-fire direction. The front-to-back ratio is 8.5 dB at 5.8 GHz. Figure 5.8 depicts the simulated and measured gain results of designed antenna in the end-fire direction. Simulated result shows that a maximum gain of 8.4 dBi can be obtained; the gain ripple is 0.9 dB over the bandwidth. The measured maximum gain is 8.2 dBi, which is slightly lower than the simulated one. This discrepancy may be attributed to the transition losses introduced by the SMA connector and microstrip feeding network. It should also be mentioned that the radiation efficiency of our proposed antenna is greater than 90% over the operation bandwidth.

Figure 5.8: Measured and simulated gain results in the end-fire direction of the proposed antenna.

5.3 Conformal Cavity-Backed Slot Antenna Mounted on Cylindrical Platform

5.3.1 Antenna Configuration

The configuration of the proposed antenna is shown in Figure 5.9. Different from the previous back-to-back slot antenna, three slots are etched on only one side of the cavity here, which is more suitable to be mounted on metallic platforms. In this structure, a curved Duroid RT 5880 substrate with permittivity of $\epsilon_r = 2.2$ and thickness of $h = 0.508$ mm is placed on the top wall of a shallow metallic cavity, whose inner dimension is $W_c \times L_c \times H_c$, and the wall thickness is denoted by t . Three slots are closely etched out in parallel on the inner layer of the substrate. The width of each slot is designed to be W , the length of three slots are L_1 , L_2 , and L_3 , respectively, which can achieve multi-resonance to broaden the bandwidth. The required magnitude and phase of the excited electric field along the slot surfaces to achieve supergain can be obtained by adjusting the positions of slots. An off-centered microstrip feedline with stubs is printed on the outer layer of the substrate. The proposed cavity-backed antenna is mounted on a large cylindrical platform with a radius of R . It is worth mentioning in the simulated model, we only use a small portion of cylindrical platform surface with the width of $(L_c + 2t)$ and radius of R to serve as the ground plane, this is because the radiation pattern of the antenna tilts close to the end-fire direction so that the other portion of the cylindrical platform can hardly affect the antenna performance. The extra length of the ground plane is L_g , which can affect the radiation beam angle of the proposed conformal antenna.

In order to obtain the required magnitude and phase conditions of the electric fields along slot surfaces for achieving supergain, the slot lengths L_n ($n = 1, 2, 3$), inter-element spacing $d_{l(r)}$, and the distance from slot 2 to the left side wall of the cavity d_s can be appropriately chosen as

$$L_n \approx \frac{\lambda_0}{2\sqrt{\epsilon_r}} \quad (5.1)$$

$$d_{l(r)} \approx 0.1\lambda_0 \quad (5.2)$$

$$d_s \approx 0.5W_c \approx 0.5\lambda_0 \quad (5.3)$$

where λ_0 is the free-space wavelength at the center frequency.

Figure 5.9: Configuration of the proposed conformal cavity-backed slot supergain antenna, (a) 3D view, (b) top view, (c) side view.

5.3.2 Antenna Analysis

In this analysis, the antenna performance with different ground shapes and sizes are studied. The planar structure can be considered as the one with an infinite radius, which is denoted as $R = \infty$.

Figure 5.10: Effects of the radius R on $|S_{11}|$ and maximum gain of the proposed conformal structure ($H_c = 3$ mm, $W_c = 28$ mm, $L_c = 28$ mm, $h = 0.508$ mm, $W = 2$ mm, $L_1 = 14.5$ mm, $L_2 = 10.6$ mm, $L_3 = 10.55$ mm, $d_s = 13$ mm, $d_l = 3.7$ mm, $d_r = 2.9$ mm, $d_f = 10.45$ mm, $t = 5$ mm, $W_f = 1.55$ mm, $W_s = 2.5$ mm, $L_s = 1.2$ mm, $P_s = 7.4$ mm, $a = 6.025$ mm, $b = 1$ mm, $Lg = 90$ mm).

The effects of the radius R on the return loss and realized gain of the proposed antenna are shown in Figure 5.10. It is clearly seen that for the planar structure with $R = \infty$, three resonances are achieved, which correspond to slots 1, 2, and 3, respectively. Two gain peaks around 8.7 dBi near $f_{\text{slot 1}}$ and 9 dBi near $f_{\text{slot 3}}$ are also obtained, and they are mainly determined by slots 1 and 2 and slots 2 and 3, respectively. When the radius decreases to be $R = 100$ mm, the structure is bent gradually so that the electrical length of each slot slightly decreases, which results in a frequency shift to a higher band. When the radius further decreases to be $R = 50$ mm, the structure is bent severely, which degrades the impedance matching of the antenna. In addition, the gain changes accordingly when the ground shape varies. The gain values remain above 7.8 dBi when the radius is larger than 100 mm, and gain ripples are less than 1 dB over the bandwidth.

Figure 5.11: Effects of the extended ground length L_g on $|S_{11}|$ and maximum gain of the proposed conformal structure with $R = 100$ mm.

Figure 5.11 shows the effects of the extended ground length L_g on the return loss and maximum gain of the proposed conformal structure. It is seen that the ground length L_g has insignificant influence on the return loss. While the gain value slightly increases as the length L_g extends. The effects of L_g on radiation patterns of the proposed antenna are illustrated in Figure 5.12. The E-plane is the y - z plane, and the H-plane is orthogonal to the E-plane and contains the peak of the radiation pattern. As shown in Figure 5.12(a), the length L_g plays a significant role in beam angle of the radiation pattern in the E-plane. When the radius is fixed to be 100 mm, the beam angles are $\theta = 32^\circ$, 40° , 50° , and 54° for the case of $L_g = 0$, 10 mm, 50 mm, and 90 mm, respectively. It is worth mentioning that if L_g is infinite, an end-fire radiation pattern may be obtained. However, due to the cavity-backed structure and finite ground size, the maximum radiation unavoidably tilts away from the end-fire direction. In addition, compared to the length L_g , the radius R has less effects on the radiation pattern of the proposed antenna, which will not be discussed here in detail.

Figure 5.12: Effects of the extended ground length L_g on radiation patterns of the proposed conformal structure with $R = 100$ mm at $f = 10$ GHz, (a) E-plane, (b) H-plane.

Table 5.2: Magnitude and phase differences of the electric field along the conformal slot structure at $f_{\text{slot 1}}$ and $f_{\text{slot 3}}$

f	$f_{\text{slot 1}}$			$f_{\text{slot 3}}$		
Slots	Slot 1	Slot 2	Slot 3	Slot 1	Slot 2	Slot 3
Phase	146.3°	0°	-67.7°	-11.3°	0°	-149.3°
Magnitude	1.2	1	0.4	0.2	1	1.65

Table 5.2 shows the simulated relative values of the magnitude and phase for the electric field along the conformal slot surfaces at $f_{\text{slot 1}}$ and $f_{\text{slot 3}}$ with $R = 200$ mm, where the gain peaks occur. Similar to Section 5.2.2, in order to facilitate the comparison, the corresponding magnitude and phase of slot 2 are normalized to be 1 and 0°, respectively. As expected, the phase differences of the electric field along the surfaces of slots 1 and 2 at $f_{\text{slot 1}}$ and slots 2 and 3 at $f_{\text{slot 3}}$ are almost equal. The corresponding magnitude of the first slot pair (slots 1 and 2) is higher than that of slot 3 at $f_{\text{slot 1}}$. Similarly, the magnitude of the second slot pair (slots 2 and 3) is larger than that of slot 1 at $f_{\text{slot 3}}$.

5.3.3 Experimental Verification

A prototype of the proposed conformal cavity-backed slot antenna mounted on a cylindrical platform is fabricated as shown in Figure 5.13. Microstrip feedline and three slots are printed and etched on the top and bottom layer of the Rogers 5880 substrate, respectively. A cavity is carved out on the surface of a cylindrical aluminum platform with the radius of $R = 200$ mm. The Rogers substrate is fixed at the top broad wall of the cavity using screws. The extended ground length L_g is 90 mm in our prototype. The proposed antenna is fed using SMA connector whose inner connector is soldered with the microstrip line. The other parameters are designed as follows: $H_c = 3$ mm, $W_c = 28$ mm, $L_c = 28$ mm, $h = 0.508$ mm, $W = 2$ mm, $L_1 = 14.5$ mm, $L_2 = 10.6$ mm, $L_3 = 10.55$ mm, $d_s = 13$ mm, $d_l = 3.7$ mm,

$d_r = 2.9$ mm, $d_f = 10.45$ mm, $t = 5$ mm, $W_f = 1.55$ mm, $W_s = 2.5$ mm, $L_s = 1.2$ mm, $P_s = 7.4$ mm, $a = 6.025$ mm, $b = 1$ mm.

Figure 5.13: Prototype of the proposed conformal cavity-backed slot antenna.

Figure 5.14: Simulated and measured $|S_{11}|$ and the maximum gain of the proposed conformal antenna mounted on cylindrical platform.

Figure 5.15: Simulated and measured radiation patterns of the proposed conformal antenna mounted on cylindrical platform at $f = 10$ GHz, (a) E-plane, (b) H-plane.

Figure 5.14 shows the simulated and measured reflection coefficients and gain of the proposed conformal antenna. Simulated results show that the -10-dB impedance bandwidth ranges from 9.26 to 10.76 GHz, indicating that a fractional bandwidth of 15% is obtained. The measured impedance bandwidth ranges from 9.2 to 10.71 GHz, slightly lower than the simulated one, which may be due to the fabrication and assembling tolerance. In addition, a maximum gain of 9.1 dBi can be obtained for the simulated result. The maximum gain is measured to be 8.5 dBi, which is 0.6 dB less than the simulated one, the discrepancy may be attributed to the loss introduced by the transition between SMA and microstrip feedline. In addition, the comparison of the simulated and measured radiation patterns at $f = 10$ GHz is shown in Figure 5.15. It is seen that simulated and measured results agree very well with each other, the beam pointing angle is 38° from the end-fire direction.

5.4 Conformal Cavity-Backed Slot Antenna Embedded in Conical Platform

5.4.1 Antenna Configuration

The configuration of the proposed planar antenna is shown in Figure 5.16. A substrate with relative permittivity of ϵ_r and dimension of $W_a \times L_a \times h$ is placed on the top broad wall of a square cavity with the size of $W_c \times W_c \times H_c$. This antenna configuration is based on the planar version in Section 5.3.1. The difference is that we employ a coaxial-to-microstrip transition here so that the antenna can be fed from the bottom side of the aluminum platform. The distances from the coaxial feedline and the starting point of the microstrip feedline to the nearest side wall of the cavity are represented by L_f and L_t , respectively. The other parameters are denoted in the same manner as those in Section 5.3.1 in order to facilitate the analysis.

Figure 5.17 shows the conformal structure of the proposed antenna. The cavity-backed slot antenna is fully embedded into the conical platform. The conical platform consists of an aluminum supporter and a radome made of Teflon. The radius and total length of the cone are R and L , respectively. The half-angle of the cone tip is denoted by α . The length of the aluminum part is represented by l_1 along the y direction. The thicknesses of the aluminum and Teflon layers are t_1 and t_2 , respectively.

5.4.2 Antenna Analysis

5.4.2.1 Planar Structure

The planar version of the proposed structure is studied first. Based on the previous section, the parameters are designed as follows: $H_c = 3$ mm, $W_c = 28$ mm, $\epsilon_r = 2.2$, $h = 0.508$ mm, $W_a = 45$ mm, $L_a = 50$ mm, $W = 2$ mm, $L_1 = 14.5$ mm, $L_2 = 10.6$ mm, $L_3 = 10.55$ mm, $d_s = 13$ mm, $d_l = 3.7$ mm, $d_r = 2.9$ mm, $d_f = 10.45$ mm, $W_f = 1.55$ mm, $W_s = 2.5$ mm, $L_s = 1.8$ mm, $P_s = 7.4$ mm, $a = 6.025$ mm, $b = 0.95$ mm, $L_f = 6$ mm, $L_t = 6.8$ mm.

Figure 5.18: Simulated $|S_{11}|$ and maximum gain of the proposed planar structure.

Table 5.3: Magnitude and phase differences of the electric field along the planar cavity-backed slot structure at $f_{\text{slot 1}}$ and $f_{\text{slot 3}}$

f	$f_{\text{slot 1}}$			$f_{\text{slot 3}}$		
Slots	Slot 1	Slot 2	Slot 3	Slot 1	Slot 2	Slot 3
Phase	145.3°	0°	-61.7°	2.61°	0°	-147.3°
Magnitude	1.2	1	0.3	0.2	1	1.65

Figure 5.18 shows the simulated $|S_{11}|$ and maximum gain of the proposed planar structure. It is seen that three resonances are achieved, which correspond to slots 1, 2, and 3, respectively. Two gain peaks around 8.4 dBi near $f_{\text{slot 1}}$ and 8.7 dBi near $f_{\text{slot 3}}$ are also obtained. In addition, the radiation beam of the proposed antenna tilts around 52° away from the end-fire direction due to the finite ground length. If the ground plane extends further along the +y direction, the radiation beam can tilt closer to the end-fire direction. Furthermore, the magnitude and phase differences of the electric fields along the slot surfaces at $f_{\text{slot 1}}$ and $f_{\text{slot 3}}$ are listed in Table 5.3. Similar to previous analyses, in order to facilitate the comparison, the magnitude and phase values of slot 2 are normalized to be 1 and 0°, respectively.

5.4.2.2 Conformal Structure

The conical platform can have significant effects on the radiation performance of the conformal antenna. Figure 5.19 shows the effects of the radius R on $|S_{11}|$ and maximum gain of the conformal version. For a small radius $R = 50$ mm, the structure is bent severely, which degrades the matching condition of the conformal antenna. As R increases, the operating bandwidth of the conformal antenna shifts to a lower band. This is because the electric length of each slot slightly increases with the increase of R . Meanwhile, the gain values also vary and gain peaks shift accordingly. In addition, radius R , length L , and angle α of the cone satisfy the following relationship

$$\tan(\alpha) = R/L \quad (5.4)$$

Therefore, it is expected that the angle α changes with different values of R if the length L is kept fixed, which can tilt the radiation beam into different directions. Figure 5.20 shows the effects of R on radiation patterns of the conformal antenna. It is worth mentioning that the end-fire direction is along the $+y$ direction, which is denoted as $\theta = 90^\circ$ and $\phi = 90^\circ$. It is seen that when the radius R of the platform is 80 mm, the radiation beam is along the end-fire direction. When the radius changes to be $R = 50$ mm and $R = 110$ mm, the beams are pointing along $\theta = 75^\circ$ and 110° , respectively.

Figure 5.19: Effects of the radius R on $|S_{11}|$ and maximum gain of the proposed conformal structure ($H_c = 3$ mm, $W_c = 28$ mm, $\epsilon_r = 2.2$, $h = 0.508$ mm, $W_a = 45$ mm, $L_a = 50$ mm, $W = 2$ mm, $L_1 = 14.9$ mm, $L_2 = 11$ mm, $L_3 = 10.8$ mm, $d_s = 13$ mm, $d_l = 3.7$ mm, $d_r = 2.9$ mm, $d_f = 10.725$ mm, $W_f = 1.55$ mm, $W_s = 2.6$ mm, $L_s = 2$ mm, $P_s = 7.5$ mm, $a = 6.525$ mm, $b = 0.95$ mm, $L_f = 4.2$ mm, $L_t = 5.2$ mm, $L = 120$ mm, $l_1 = 55$ mm, $t_1 = 2$ mm, $t_2 = 4$ mm).

Figure 5.20: Effects of the radius R on radiation patterns of the proposed conformal structure at $f = 10$ GHz, (a) E-plane, (b) H-plane.

The length of the aluminum part l_1 plays a significant role in antenna radiation performance because the aluminum platform serves as the ground of the conformal antenna. As shown in Figure 5.21, the impedance characteristic is almost unaffected by changing the length l_1 , but the gain values are enhanced with the increase of l_1 . As a comparison, we also plot the antenna performance for a whole aluminum platform, in which case the lengths l_1 and L are equal, and the peak gain can reach above 10 dBi. In addition, the effects of the ground length l_1 on radiation patterns of the proposed conformal structure are shown in Figure 5.22. It is seen that l_1 can also influence the radiation beam angle in the E-plane. For the case of $l_1 = 40$ mm, 55 mm, 70 mm, and the whole aluminum cone, the corresponding beam angles are $\theta = 78^\circ$, 90° , 94° , and 102° , respectively.

Figure 5.21: Effects of the ground length l_1 on $|S_{11}|$ and maximum gain of the proposed conformal structure with $R = 80$ mm.

Furthermore, the length of the cone L also affects the radiation beam angle of the conformal antenna. For example, when $R = 80$ mm and $l_1 = 55$ mm, the beam angles are $\theta = 84^\circ$, 90° , and 96° for the case of $L = 140$ mm, 120 mm, and 100 mm, respectively. In addition, the thickness of the Teflon t_2 only influences the radiation pattern in the E-plane slightly, which will not be discussed here in detail for the sake of brevity.

Figure 5.22: Effects of the ground length l_1 on radiation patterns of the proposed conformal structure at $f = 10$ GHz with $R = 80$ mm, (a) E-plane, (b) H-plane.

From the analyses above, it is concluded that the radius R and angle α mainly affect the impedance matching and the radiation beam angle of the proposed conformal antenna, respectively. Meanwhile, the ground length l_1 can influence both gain and radiation beam of the conformal antenna.

Figure 5.23: Conformal array configuration with six supergain slot antenna elements embedded into the conical platform, (a) perspective view, (b) left view.

Figure 5.24: Radiation beam coverage of the proposed supergain slot antenna array configuration at $f = 10$ GHz when each antenna element is excited independently.

Figure 5.23 shows the structure of the proposed conformal array configuration with six separate embedded antenna elements. It should be mentioned that the polarization of each antenna element is different due to different positions of six elements with respect to the axis of the cone. When each element is excited independently, the array configuration can receive signals irrespective of polarization in the end-fire direction. The radiation gain values in terms of both directions (θ , ϕ) of each antenna element are plotted in Figure 5.24. Due to the symmetry of the six-element array configuration, the radiation beam of each antenna element covers every 60° in the ϕ direction sequentially according to the coordinate system in Figure 5.23. It is clearly seen that the maximum gain value can reach above 9 dBi and the radiation beam is pointing along the end-fire direction $\theta = 0^\circ$.

Figure 5.25: Prototype of the proposed conformal antenna array configuration, (a) perspective view, (b) details of the cone surface, (c) view from the bottom side of the cone.

5.4.3 Experimental Verification

Figure 5.25 shows the prototype of the proposed conformal array configuration. Six separate antenna elements are uniformly embedded into the conical platform. The conical platform consists of two parts, an aluminum supporter and a Teflon radome. They are seamlessly combined together. The cavities are carved out from the surface of the aluminum part. Rogers 5880 substrate is used to fabricate the slots and microstrip feedline, and all substrate layers are wrapped on the top side of the cavity. Each antenna element is fed from the bottom side of the aluminum cone using SMA connector. The parameters are designed as follows: $H_c = 3$ mm, $W_c = 28$ mm, $\epsilon_r = 2.2$, $h = 0.508$ mm, $W_a = 45$ mm, $L_a = 50$ mm, $W = 2$ mm, $L_1 = 14.9$ mm, $L_2 = 11$ mm, $L_3 = 10.8$ mm, $d_s = 13$ mm, $d_l = 3.7$ mm, $d_r = 2.9$ mm, $d_f = 10.725$ mm, $W_f = 1.55$ mm, $W_s = 2.6$ mm, $L_s = 2$ mm, $P_s = 7.5$ mm, $a = 6.525$ mm, $b = 0.95$ mm, $L_f = 4.2$ mm, $L_t = 5.2$ mm, $L = 120$ mm, $R = 80$ mm, $l_1 = 55$ mm, $t_1 = 2$ mm, $t_2 = 4$ mm.

Figure 5.26: Simulated and measured $|S_{11}|$ and maximum gain of the proposed antenna element embedded into the platform.

Figure 5.27: Simulated and measured radiation patterns of the proposed conformal antenna element embedded into conical platform at $f = 10$ GHz, (a) E-plane, (b) H-plane.

The radiation characteristics of a single antenna element embedded into the platform are measured. As shown in Figure 5.26, the 10-dB return loss bandwidth ranges from 9.3 to 10.81 GHz and from 9.18 to 10.7 GHz for simulated and measured results, respectively. The frequency shifting may be attributed to the errors of the fabrication and assembling. It is also shown that the simulated gain values within the operating bandwidth are from 8.8 to 9.5 dBi.

The measured gain values are a bit lower than simulated ones, which may be due to the alignment error of the measurement setup and the loss introduced by the SMA connector. In addition, the radiation patterns at $f = 10$ GHz are plotted in Figure 5.27. It is seen that the simulated and measured results are in good agreement, showing that good end-fire radiation patterns are obtained. It is worth mentioning that the simulated radiation efficiency of the proposed conformal antenna ranges from 92.5% to 97.8% within the entire operating bandwidth.

Figure 5.28: Measured coupling between each antenna element in the array configuration.

Furthermore, the coupling between each element in the array configuration is also measured. As shown in Figure 5.28, the S-parameters between each port are all lower than -40 dB, which indicates that the couplings among the antenna elements are very weak. It should be mentioned that due to the symmetry of the structure, $|S_{12}|$ is equal to $|S_{16}|$, and $|S_{13}|$ is equal to $|S_{15}|$.

Table 5.4 summarizes the performance comparison of various antennas of high gain. Although the antenna in [94] can achieve the maximum gain of 9.7 dBi and bandwidth of 9.67% using three monopole elements, the radiation beam tilts away from the end-fire direction and the structure is not suitable to be mounted on moving platforms. From the comparison, it is clearly seen that our proposed antenna exhibits the advantages in terms of bandwidth and gain. More importantly, the proposed conformal antenna can achieve end-fire radiation as well as minimize the aerodynamic drag for moving platforms.

Table 5.4: Performance comparison of various antennas of high gain

Ref.	Element Type	Element No.	Element Spacing (λ_0)	Bandwidth ($ S_{11} < -10$ dB)	Max. Gain (dBi)	Radiation
[100]	Monopole	2	0.1	5.8%	8.9	Tilted
[94]	Monopole	3	0.053	9.67%	9.7	Tilted
[98]	Dipole	2	0.02	1.05%	6.6	End-fire
[97]	Dipole	3	0.02	1.9%	7.12	End-fire
[153]	Slot	3	0.12	10.3%	8.4	End-fire
This work	Slot	3	0.1	15.2%	9.1	End-fire

5.5 Summary

In this chapter, three cavity-backed supergain slot antennas have been presented. In the first configuration, two substrates on both broad walls of a metallic cavity are employed to form a back-to-back structure. With suitable positions and lengths of three slots, the antenna has achieved a maximum gain of 8.4 dBi in the end-fire direction and a fractional impedance bandwidth of 9.8%. Based on the

back-to-back supergain slot antenna, conformal structures that are mounted on cylindrical and conical platforms have also been presented. The conformal antenna element has achieved a maximum gain above 9 dBi and a fractional bandwidth over 15% when mounted on large cylindrical and conical conducting platforms. The proposed antenna has also been extended into an array configuration with six antenna elements embedded into the conical platform.

The proposed slot supergain antennas have achieved the radiation efficiency of above 90% over the operating bandwidth. The limitations of traditional superdirective antennas have been alleviated effectively. The performance of high gain, end-fire radiation, and enhanced bandwidth performance has been verified by experimental results showing that the proposed antennas are promising for many wireless communication applications.

CHAPTER 6

PLANAR HELICAL ANTENNA OF CIRCULAR POLARIZATION

6.1 Introduction

Several end-fire antennas with various characteristics such as wide bandwidth, supergain, and low profile have been proposed in the previous chapters. However, they are all linearly polarized. In some airborne applications, circularly polarized (CP) antennas with a low profile are also in great demand to avoid the polarization mismatch and suppress the multipath reflections.

A rectangular planar helix has found many applications in slow-wave structures [123]-[126] and linearly polarized antennas [127]-[132], but little literature has been reported on CP end-fire antennas so far. It is well known that in order to realize circular polarization, two orthogonal electric fields with equal amplitude and phase difference of 90° are required. For rectangular planar helical antennas, two orthogonal electric fields can be generated by the currents traveling along the perpendicular sides of the helix, respectively. It is expected that if one side of the rectangle is much longer than the other, the amplitudes of two orthogonal electric fields will be of great difference. Therefore, it remains a challenge to achieve CP end-fire radiation for a low-profile planar helical antenna with a rectangular cross-section, which may be essential for achieving a low-profile structure.

In this chapter, a planar rectangular helical antenna with straight-edge connections implemented by plated via-holes is presented, which is very low-profile and can obtain CP end-fire radiation. Besides the currents flowing along the strips and via-holes of the helix contributing to the horizontal and vertical polarizations, respectively, the current on the inner edge of the ground plane plays an important role in weakening the strong amplitude of the horizontal electric field generated by the current on the strips, which is critical for achieving

circular polarization. In addition, tapered helix and conducting side-walls are utilized in the antenna to improve the axial ratio bandwidth and the radiation pattern. The proposed planar helical antenna has the following advantages. 1) The antenna can achieve wide impedance and axial ratio bandwidths. Excellent CP performance with end-fire radiation is obtained over a 34% frequency band. 2) The antenna has a very low profile, which is only $0.11\lambda_0$ at the center frequency. 3) The structure is very simple, lightweight, and low-cost. It can be easily fabricated using the printed circuit board technology. 4) It is very suitable to be mounted on the wings of the unmanned aerial vehicle (UAV) and aircraft.

6.2 Antenna Configuration

The configuration of the proposed planar helical antenna is shown in Figure 6.1. The helix has a rectangular cross-section, which is formed using printed strips with straight-edge connections implemented by plated via-holes. As shown in Figure 6.1(a), the length and width of the antenna are denoted by L and W , respectively. ϵ_r and h represent the dielectric constant and thickness of the substrate, respectively. The diameter of the via-hole of the helix is d_v , and the pitch angle between the top and bottom strip is α . The width of each printed strip is uniform, which is represented by W_s . The length of the strip on the bottom layer of the substrate is denoted by L_{si} ($i = 1, 2, 3, 4$) for the i th turn of the helix. In order to improve the axial ratio and radiation pattern, the helix is tapered according to the following ratio

$$r = \frac{L_{s2}}{L_{s1}} = \frac{L_{s3}}{L_{s2}} = \frac{L_{s4}}{L_{s3}} \quad (6.1)$$

A coaxial-to-grounded coplanar waveguide (GCPW) transition is employed in the antenna to feed the helix. As shown in Figure 6.1(b), W_f is the width of the center strip of the GCPW. g_1 and g_2 represent the gaps between the center strip and its adjacent ground planes. d_f is the distance from the edge of the center strip of the GCPW to the coaxial probe. If a substrate is chosen, the characteristic impedance of the GCPW is determined by the strip width W_f and the gap g_1 , while

the matching condition from the coaxial probe to the GCPW can be tuned by varying d_f and g_2 .

Figure 6.1: Configuration of the proposed planar helical antenna, (a) perspective view, (b) top and (c) bottom layers of the substrate.

Generally speaking, a large conducting plane that is vertical to the axis of the helix should be incorporated in the helical antenna to obtain a good CP radiation in the end-fire direction. However, it is not suitable for the low-profile structures. In our proposed antenna, two metallic plates are printed on both the top and

bottom layers of the substrate to form the horizontal ground planes. Grounded via-holes with diameter d and spacing s are perforated along the inner edge of the ground planes and around the GCPW, and they satisfy the following criteria [154]

$$0.05\lambda_g < s < 0.25\lambda_g, \quad s < 2d \quad (6.2)$$

to prevent the power from leaking out through the via-gaps, where λ_g is the guided wavelength. These via-holes can be regarded as a small vertical conducting wall of the helical antenna, and they can also help to obtain a good RF ground connection between the top and bottom layers of the substrate and suppress the unwanted parallel-plate waveguide mode. It should be noted that the vertical wall is U-shaped, which consists of the vertical back-wall (AA') and side-walls. The side-walls in the structure can help to increase the gain and to improve the radiation pattern. Furthermore, L_t and L_b in Figures 6.1(b) and (c) represent the distances from the vertical back-wall (AA') to the inner edge of the top and bottom grounds, respectively. They can affect the current distribution on the inner edge of the horizontal ground plane.

6.3 Antenna Analysis

6.3.1 One-Turn Helix

The proposed low-profile helical antenna with one-turn helix is designed and analyzed in this part. The substrate used is the Duroid RT 5880 substrate with a relative permittivity of $\epsilon_r = 2.2$ and thickness of $h = 3.175$ mm. The strip length of the helix is represented by L_s , the other dimensions are the same as those indicated in Figure 6.1. The effective total length of one-turn helix L_{eff} is chosen to be around one free-space wavelength in order to achieve CP radiation in the end-fire direction, which can be expressed as

$$L_{eff} = 2(h \times \sqrt{\epsilon_{eff1}} + L_s \times \sqrt{\epsilon_{eff2}}) \approx 1\lambda \quad (6.3)$$

where $\epsilon_{eff1} \approx \epsilon_r$ and $\epsilon_{eff2} \approx (\epsilon_r + 1)/2$ [155], [156].

Figure 6.2: Current distributions of the low-profile one-turn helical antenna at different times $t = 0, T/4, T/2$, and $3T/4$ at $f = 10 \text{ GHz}$, where T is the period of time. (a), (c), (e), (g), horizontal currents on the ground plane and strips of the helix. (b), (d), (f), (h) Vertical currents along the via-holes of the helix.

(a)

(b)

(c)

Figure 6.3: (a) Magnitude and (b) phase values of surface current density integration of each segment on the helix and edge of the ground. (c) Magnitude and phase values of combined total surface current density integrations of x - and z -directions.

In general, circular polarization is generated by two orthogonal electric fields, which are equal in amplitude and 90° different in phase. For the proposed helical structure, the currents flowing on the strips and along the via-holes of the helix contribute to the horizontal electric field E_ϕ and vertical electric field E_θ , respectively. It is expected that the magnitude of E_ϕ is much stronger than that of E_θ if the height of the via-hole h is very small compared to the strip length L_s , which makes it difficult to meet the criteria of achieving circular polarization. However, in our proposed helical antenna, the current flowing on the inner edge of the ground plane can help to weaken the strong magnitude of E_ϕ generated by the current on the strips. By this way, it is possible to obtain circular polarization in the end-fire direction over a wide frequency range.

To verify the principle of achieving the circular polarization, the current distribution along the proposed one-turn helical antenna is analyzed with the aid of HFSS, as shown in Figure 6.2. The parameters of the one-turn helical antenna are as follows: $W = 37$ mm, $L = 38$ mm, $h = 3.175$ mm, $W_s = 3$ mm, $L_s = 10$ mm, $\alpha = 28^\circ$, $d_v = 2$ mm, $d = 1$ mm, $s = 1.5$ mm, $W_f = 3$ mm, $g_1 = 0.8$ mm, $g_2 = 1.5$ mm, $d_1 = 1.7$ mm, $d_2 = 1.4$ mm, $d_f = 1$ mm, $L_t = 4.7$ mm, $L_b = 3.1$ mm, $L_f = 9.17$ mm, $d_t = 2.77$ mm, $L_1 = 15.3$ mm, $b = 19.6$ mm. When $t = 0$, it is observed from Figure 6.2(a) that strong horizontal currents appear on the bottom and top strips, and also on the inner edge of the ground plane, which are denoted by I_b , I_t , and I_{g1} , respectively. On the other hand, the vertical currents flowing along the via-holes are very small, as shown in Figure 6.2(b). Therefore, it is concluded that E_ϕ is dominated along the end-fire direction in the far-field region. It should be mentioned that I_{g1} is $-x$ directed, which is opposite to the $+x$ direction of I_b and I_t , so that it can somehow weaken the strong amplitude of E_ϕ generated by I_b and I_t . This is critical for achieving circular polarization, as mentioned earlier. Figures 6.2(c) and (d) show the current distributions at $t = T/4$, where T is the period of time. It is seen that strong vertical currents I_{v1} and I_{v2} flow along the via-holes of the helix in the $+z$ direction, while the horizontal currents, as indicated in Figure 6.2(c) are weak compared to I_{v1} and I_{v2} . In addition, the currents either on the edge of the ground plane or the strips exhibit two components that are 180° out

of phase, so that E_ϕ is very small and E_θ dominates in the far field. The same phenomenon is observed in Figures 6.2(e)-(h). As seen, the current distributions at $t = T/2$, and $t = 3T/4$ are opposite to those when $t = 0$ and $t = T/4$, which also generate the dominant but out-of-phase E_ϕ and E_θ , respectively. As a result, left-hand circular polarization in the end-fire direction can be obtained under the combined influences of the current distributions along the helix as well as on the ground plane of the proposed antenna.

Figure 6.4: Simulated magnitude ratio and phase difference of the horizontal electric field E_ϕ and vertical electric field E_θ , and axial ratio of the proposed one-turn helical antenna in the end-fire direction.

It is well known that the electric field in the far-field is determined by integrating the surface current densities over the distributed areas, which can be represented as [2]

$$\mathbf{E}_i \approx -j \frac{\omega\mu}{4\pi} \iint \mathbf{J}_{S_i} \frac{e^{-jkR'_i}}{R_i} dS_i \quad (6.4)$$

where i represents each segment on the helix and the edge of the ground for our one-turn helical antenna. These segments include the bottom and top strips, the ground edges on each side of the CPW, and two via-holes, which are denoted by b , t , $g1$, $g2$, $v1$, and $v2$, respectively. S_i denotes the surface area of segment i , and $\mathbf{J}_{S_i}$ is the surface current density distributed on S_i . k is the wavenumber. R_i and R'_i are the distances from the current source to the far-field for amplitude and phase terms, respectively. It is noticed that if we assume that there are no amplitude and phase differences between each current source to the far-field region, the ratio of the horizontal E-field and vertical E-field approximately equals to that of the combined total integrations of surface current densities over the distributed areas of x component $\iint \mathbf{J}_{S_x} dS$ and z -component $\iint \mathbf{J}_{S_z} dS$ in the end-fire direction. With the aid of the fields calculator of HFSS, the magnitude and phase values of surface current density integration of each segment on the helix and the edge of the ground are plotted in Figures 6.3(a) and (b). We use $Int_J_{S_i}$ ($i = b, t, g1 \dots$) to represent the integration of surface current density over the area of each segment. The input power is set as 1 Watt. As shown, $Int_J_{S_b}$ and $Int_J_{S_t}$ are almost in-phase, which are nearly 180° out-of-phase with $Int_J_{S_{g1}}$ near the center frequency of 10 GHz. In addition, $Int_J_{S_{v1}}$ and $Int_J_{S_{v2}}$ are almost in-phase, which have phase difference around 90° with $Int_J_{S_b}$ and $Int_J_{S_t}$ near the center frequency. These phenomena are in accordance with the current distribution displayed in Figure 6.2. The magnitude and phase values of the x - and z -components of the combined total surface current density integrations are also depicted in Figure 6.3(c). It is seen that the magnitudes of x - and z -components are very close, and the phase difference is around 90° near the center frequency of 10 GHz. It is worth noting that no phase differences between each current source to the far-field region is assumed for the analysis in order to basically clarify the physical concept of achieving circular polarization. By taking the phase variations into account, the magnitude ratio and phase difference between the horizontal electric field E_ϕ and vertical electric field E_θ , and axial ratio of the proposed one-turn helical antenna in the end-fire direction are plotted

in Figure 6.4. It is seen that the magnitude ratio of E_ϕ and E_θ is close to one and the phase difference between two orthogonal fields is almost 90° near the center frequency of 10 GHz. Resultantly, a good circular polarization in the end-fire direction is obtained and the 3-dB AR bandwidth ranges from 9.6 to 10.45 GHz.

Additionally, it should be noted that the vertical conducting wall made up of via-holes can prevent the power from leaking out through the via-gaps. Resultantly, the current flowing on the ground plane outside the vertical wall is very small. Therefore, the ground length L_1 starting from the wall to the left edge of the substrate and the ground width W , as shown in Figure 6.2(a), are of insignificance to the antenna performance, which makes the proposed antenna suitable to be mounted on a large conducting platform.

Furthermore, the helical antennas can achieve wide bandwidth when they operate in the axial mode. This is because the helical antenna is a traveling-wave structure; the outgoing current at the input end and the reflected current at the open end of the helix decay rapidly [19]. In our proposed structure, a coaxial-to-GCPW transition is employed to feed the antenna. By choosing proper parameters W_f , g_1 , g_2 , and d_f , a good impedance match can be obtained. Figure 6.5 shows the simulated input impedance and reflection coefficient $|S_{11}|$ of the proposed one-turn helical antenna. It is seen that the input impedance can maintain a small variation within a wide frequency range, and the 10-dB return loss bandwidth is from 7.2 to 12.5 GHz.

6.3.2 Parametric Study

A parametric study is conducted in order to examine the effects of the key parameters on the radiation performance and help to facilitate the practical design of the proposed low-profile planar helical antenna. The important parameters include the number of helical turns N , the length of the strip L_s , the pitch angle between the top and bottom strips α , the distances from the vertical back-wall to the inner edge of the top ground L_t and bottom ground L_b , and the strip-length ratio r . Other parameters such as the width of the strip W_s and the diameter of the via-hole d_v are of less importance although they also slightly affect the radiation

performance, and they will not be discussed in detail below. In the parametric study, only one parameter is varied at one time while the others remain unchanged as follows: $W = 40$ mm, $L = 60$ mm, $h = 3.175$ mm, $W_s = 3$ mm, $d_v = 2$ mm, $d = 1$ mm, $s = 1.5$ mm, $W_f = 2.2$ mm, $g_1 = 0.6$ mm, $g_2 = 2$ mm, $d_1 = 2$ mm, $d_2 = 1.5$ mm, $d_f = 1.1$ mm, $L_f = 8.75$ mm, $d_t = 2.25$ mm, $L_1 = 17.5$ mm, and $b = 22$ mm.

1) *Number of Turns N* : Basically, the helical antenna can achieve better circular polarization performance and higher gain if more turns of the helix are employed. Figures 6.6 and 6.7 show the effects of the number of turns N on the axial ratio (AR) and radiation pattern of the proposed antenna, respectively. It is seen that the fractional AR bandwidths ($AR \leq 3$ dB) are 8.5, 16, 26, and 28.3% for $N = 1, 2, 4$, and 6, respectively. In addition, the gain values at the center frequency $f = 10$ GHz are 6.2, 6.9, 7.3, and 7.5 dBi for $N = 1, 2, 4$, and 6, respectively. These results indicate that although the axial ratio bandwidth and gain can be both enhanced with the increase of N , the improvement is small when N is larger than 4. This is due to the fact that the current decays when it travels along the helix [107]. When N is large, the current along the last few turns of the helix is small, slightly contributing to the performance of the antenna. Based on this study, we choose $N = 4$ in our proposed structure.

2) *Strip Length L_s* : As mentioned earlier, the effective circumference of a helix L_{eff} should be around one free-space wavelength in order to achieve CP radiation in the end-fire direction. Therefore, the strip length L_s is a key parameter to determine the operating frequency band of circular polarization if the substrate thickness is fixed. The effect of L_s on the AR of the proposed helical antenna is shown in Figure 6.8. It is observed that the 3-dB AR frequency band shifts to a lower one as the length L_s increases. When $L_s = 10.5$ mm, the 3-dB AR bandwidth can cover from 8.6 to 11.2 GHz.

Figure 6.5: Simulated input impedance and $|S_{11}|$ of the proposed one-turn helical antenna.

Figure 6.6: Effects of the number of turns N on the axial ratio of the proposed helical antenna ($L_s = 10.5$ mm, $\alpha = 28^\circ$, $L_t = 4$ mm, $L_b = 2.3$ mm).

Figure 6.7: Effects of the number of turns N on the radiation pattern of the proposed helical antenna at $f = 10$ GHz, (a) xy -plane, (b) yz -plane ($L_s = 10.5$ mm, $\alpha = 28^\circ$, $L_t = 4$ mm, $L_b = 2.3$ mm).

Figure 6.8: Effects of the strip length L_s on the axial ratio of the proposed helical antenna ($N = 4$, $\alpha = 28^\circ$, $L_t = 4$ mm, $L_b = 2.3$ mm).

Figure 6.9: Effects of the angle α on the axial ratio of the proposed helical antenna ($N = 4$, $L_s = 10.5$ mm, $L_t = 4$ mm, $L_b = 2.3$ mm).

3) *Pitch Angle α* : The pitch angle α plays an important role on the current distribution on the top and bottom strips so that it can affect the axial ratio of the proposed helical antenna. As shown in Figure 6.9, the 3-dB AR bandwidth is 12.5% for a small angle $\alpha = 20^\circ$, and it can be enhanced by increasing the value α appropriately. When the angle $\alpha = 28^\circ$ and $\alpha = 32^\circ$, the AR bandwidths are both around 26%. On the other hand, it should be mentioned the gain values in the end-fire direction are 7.6, 7.4, 7.3, and 7 dBi for $\alpha = 20^\circ$, 24° , 28° , and 32° , respectively, which indicates that a larger α may slightly degrade the gain of the antenna. This is due to the fact that the current flowing on the strips can be divided into x - and y -directed components, a larger α results in an increase of the y -directed component and decrease of the x -directed one, which can enhance the sidelobe of the radiation pattern and reduce the gain in the $+y$ direction. Therefore, taking both axial ratio and gain into account, we choose $\alpha = 28^\circ$ in the proposed antenna.

4) *Distances L_t and L_b* : The distances L_t and L_b exhibit significant effects on AR performance of the antenna since they can influence the current distributions on the inner edge of the top and bottom ground planes. As shown in Figure 6.10, it is found that the AR values of the antenna are sensitive to the variations of L_t and L_b , and the best AR performance of the proposed structure with four-turn uniform helix can be obtained when $L_t = 4$ mm and $L_b = 2.3$ mm. Therefore, the values of L_t and L_b should be carefully optimized in order to achieve good CP performance.

5) *Strip-Length Ratio r* : The significance of the strip-length ratio r on the antenna's AR performance is shown in Figure 6.11. In the parametric study, the length of the first strip L_{S1} is kept as 12.5 mm, and the pitch angle is maintained as $\alpha = 28^\circ$. The planar helix is uniform when $r = 1$, and is tapered as r decreases. It is seen that the values of the AR at higher frequencies are improved and the AR bandwidth is enhanced when the helix gradually tapers. The best AR performance of the proposed structure is obtained when $r = 0.89$. It should also be noted that r cannot be too small to avoid the overlapping of both edge ends of printed strips.

Figure 6.10: Effects of the distance (a) L_t and (b) L_b on the axial ratio of the proposed helical antenna ($N = 4$, $L_s = 10.5$ mm, $\alpha = 28^\circ$).

6.3.3 Effects of the Tapered Helix and Conducting Side-Wall

The radiation pattern of the antenna can be improved by utilizing tapered helix and conducting side-walls. Figure 6.12 shows the simulated radiation patterns of three antenna designs, which include Ant. #1: uniform helix without a side-wall, Ant. #2: tapered helix without side-wall, and Ant. #3: tapered helix with side-walls. It is seen that the sidelobe levels (SLLs) of the radiation patterns of Ant. #1 are relatively high. In comparison, by tapering the helix as Ant. #2, the SLLs are improved, and the gain values increase accordingly at higher frequencies. Furthermore, comparing the radiation patterns of Ant. #3 and Ant. #2, it is concluded that employing side-walls can further enhance the gain and improve the SLLs of the proposed antenna. In addition, the AR values of the three antennas are compared in Figure 6.13. It is observed that the AR bandwidths of Ant. #2 and Ant. #3 are almost the same, and they are wider than that of Ant. #1. Therefore, tapered helix and conducting side-walls are both employed in our final design considering the improvement of AR, antenna gain as well as the radiation pattern.

Figure 6.11: Effects of the strip-length ratio r on the axial ratio of the proposed helical antenna ($N = 4$, $L_{s1} = 12.5$ mm, $\alpha = 28^\circ$, $L_t = 4$ mm, $L_b = 2.3$ mm).

Figure 6.12: Radiation pattern comparison of three planar helical antenna designs at different frequencies. (a) Ant. #1: uniform helix without side-wall ($r = 1$, $L_{s1} = 10.5$ mm), (b) Ant. #2: tapered helix without side-wall ($r = 0.89$, $L_{s1} = 12.5$ mm), and (c) Ant. #3: tapered helix with side-walls ($r = 0.89$, $L_{s1} = 12.5$ mm).

Figure 6.13: Simulated axial ratio values of three planar helical antenna designs (Ant. #1: uniform helix without side-wall, Ant. #2: tapered helix without side-wall, and Ant. #3: tapered helix with side-walls.)

6.4 Experimental Verification

A prototype is fabricated to characterize the performance of the proposed planar helical antenna, as shown in Figure 6.14. The substrate used in this design is Duroid RT 5880 substrate with permittivity of 2.2, loss tangent of $\tan\delta = 0.0009$, and thickness of 3.175 mm. The outer conductor of a 50 Ω SMA connector is soldered on the bottom layer of the substrate; the inner conductor is connected to the center strip of the GCPW to feed the antenna. For the matching purpose, the center strip width of the GCPW is designed to be $W_f = 3$ mm, the gaps between the center strip and its adjacent ground planes are $g_1 = 0.8$ mm and $g_2 = 1.5$ mm, respectively. The distance from the edge of the center strip of the GCPW to the coaxial probe is $d_f = 1.2$ mm. The corresponding characteristic impedance of the GCPW is 73.1 Ω . Based on the analysis conducted in Section 6.3, the other dimensions are: $W = 44$ mm, $L = 52$ mm, $W_s = 3.2$ mm, $L_{s1} = 12.5$ mm, $r = 0.89$, $\alpha = 28^\circ$, $d_v = 2$ mm, $d = 1$ mm, $s = 1.5$ mm, $d_1 = 1.83$ mm, $d_2 =$

1.45 mm, $L_t = 4.9$ mm, $L_b = 3.2$ mm, $L_f = 9.66$ mm, $d_t = 1.86$ mm, $b = 24.32$ mm,
 $L_1 = 15.9$ mm, $L_2 = 16.3$ mm, $t_1 = 4.67$ mm, $t_2 = 2.13$ mm, $t_3 = 3.5$ mm.

(a)

(b)

Figure 6.14: Prototype of the proposed planar helical antenna, (a) top view, (b) bottom view.

Figure 6.15: Simulated and measured $|S_{11}|$ results of the proposed antenna.

Figure 6.16: Simulated and measured axial ratio of the proposed antenna.

Figure 6.17: Simulated and measured radiation patterns of the proposed antenna at different frequencies. (a) 8.5 GHz, (b) 9.5 GHz, (c) 10.5 GHz, (d) 11.5 GHz.

S-parameter of the proposed antenna is measured using a calibrated Agilent vector network analyzer N5230A. The comparison between the simulated and measured $|S_{11}|$ is shown in Figure 6.15. It is seen that the 10-dB return loss bandwidth ranges from 7.4 to 12.8 GHz for the simulated result, and from 7.2 to 12.9 GHz for the measured one. The corresponding fractional bandwidths are 54% and 57%, respectively. The slight difference may be due to the fabrication tolerance.

Axial ratio and radiation pattern of the proposed antenna are measured in an anechoic chamber. Figure 6.16 shows the simulated and measured axial ratio values in the end-fire direction. The simulated result indicates that a 3-dB axial

ratio bandwidth of 34% can be achieved for the proposed antenna, which covers from 8.2 to 11.6 GHz. The measured result agrees well with the simulated one. The simulated and measured radiation patterns in the xy -plane and yz -plane at different frequencies are depicted in Figure 6.17. As seen, end-fire radiations are obtained over the frequency range from 8.5 to 11.5 GHz, simulated and measured results are in good agreement as well. It is noted that the radiation patterns are not very symmetric respect to end-fire direction, which is due to the asymmetry of the structure. In addition, the sidelobe level (SLL) and front-to-back ratio (F/B) are relatively high at the lowest frequency of $f = 8.5$ GHz. Two reasons may account for this phenomenon. First, the normal mode of the helical antenna may appear at the lowest frequency, which can distort the radiation pattern of the axial mode. Second, it is well known that a large conducting wall vertical to the axis of the helix is generally employed in the traditional helical antennas to improve the SLL and F/B . However, our structure only employs a small vertical via-wall rather than a large one for the sake of low profile. The height of the vertical wall is very small compared to the operating wavelength, which degrades the radiation pattern, especially at lower frequencies. In addition, the simulated and measured gain values in the end-fire direction are depicted in Figure 6.18. The simulated gain is 3.1 dBi at 8.5 GHz and increases to 10.5 dBi at 11.5 GHz. At frequencies lower than 8.5 GHz and higher than 11.5 GHz, the gain decreases with the degradation of the end-fire radiation patterns. The measured gain values are slightly less than the simulated ones. This discrepancy may be due to the additional loss introduced by the SMA connector and the alignment error of the measurement setup. Furthermore, it should be mentioned that perfect electric conductors are assumed in the simulated model. By taking the loss of the substrate into account, the simulated radiation efficiency is also plotted in Figure 6.18. It is seen that the proposed antenna achieves a very high radiation efficiency.

Table 6.1 summarizes the performance comparison of existing low-profile helical antennas. It is seen that although the antennas in [128] and [131] retain much lower thickness than the others, they achieve linear polarization only. The planar helix antenna in [136] can achieve axial ratio bandwidth of 30% and gain of 10.3 dBi. However, the antenna consists of two separate layers of substrates,

Figure 6.18: Simulated and measured gain values in the end-fire direction and simulated radiation efficiency of the proposed antenna.

Table 6.1: Performance comparison of existing low-profile helical antennas

Ref.	Structure	Impedance Bandwidth ($ S_{11} < -10$ dB)	AR Bandwidth (AR < 3 dB)	Height (λ_0 is the wavelength at the center frequency)	Gain at the center frequency (dBi)	Polarization	Fabrication Complexity
[107]	Non-planar	-	12%	$0.19 \lambda_0$	9	Circular	Medium
[115]	Non-planar	-	14.6%	$\sim 0.15 \lambda_0$	9.2	Circular	Complex
[120]	Planar	20.5%	Narrow	$0.22 \lambda_0$	-	Circular	Medium
[128]	Planar	15%	-	$0.02 \lambda_0$	-	Linear	Simple
[131]	Planar	6.1%	-	$0.013 \lambda_0$	0	Linear	Simple
[136]	Planar	34%	30%	$0.27 \lambda_0$	10.3	Circular	Simple
Our work	Planar	54%	34%	$0.11 \lambda_0$	8	Circular	Simple

and the antenna thickness is $0.27\lambda_0$, which is relatively thick. In addition, a ground plane vertical to the axis of the helix is employed in the structure, which is unsuitable for surface-mounted applications. Compared to the other circularly polarized low-profile helical antennas, our proposed antenna clearly exhibits the overwhelming advantages in terms of impedance and AR bandwidths as well as the antenna thickness. In addition, our proposed planar helical antenna is very simple and is suitable for mass production with low cost.

6.5 Summary

A planar helical antenna has been presented for achieving wideband end-fire radiation of circular polarization while maintaining a very low profile. The helix is formed using printed strips with straight-edge connections implemented by plated via-holes. The operating principle of achieving the circular polarization has been analyzed based on the current distribution along the proposed one-turn helical antenna. The effects of key parameters on the antenna performance have also been carefully investigated. Furthermore, a tapered helix and conducting sidewalls have been employed to broaden the axial ratio bandwidth as well as to improve the end-fire radiation pattern.

The proposed planar helical antenna operates at the center frequency of 10 GHz. It has been shown in the simulated results that the antenna can achieve a wide impedance bandwidth ($|S_{11}| < -10$ dB) from 7.4 to 12.8 GHz (54%) and 3-dB axial ratio bandwidth from 8.2 to 11.6 GHz (34%) while retaining a thickness of only $0.11\lambda_0$ at the center frequency. A prototype of the proposed antenna has been fabricated and tested showing that the measured results are in good agreement with simulated ones.

CHAPTER 7

CONCLUSIONS AND RECOMMENDATIONS

7.1 Conclusions

This thesis focuses on overcoming the limitations of traditional end-fire antennas and proposing novel and effective solutions to achieve low-profile and end-fire antennas, which can be mounted on metallic platforms for airborne applications. Traditional end-fire antennas such as log-periodic, Yagi, and Vivaldi antennas can achieve wide bandwidth, but they are not suitable for flush-mounted applications since the mounting platforms have significant effects on the antenna performance. H-plane horn antenna can alleviate the ground effect, but it is very challenging to achieve a wide bandwidth while maintaining a very low profile. In addition, although superdirective antennas can obtain end-fire radiation, they suffer from high mismatch loss, narrow bandwidth, and low radiation efficiency. Furthermore, circularly polarized antennas are also in great demand in airborne applications. However, very few candidates can meet the requirements of end-fire radiation with very low profile. Although many attempts have been made to achieve circular polarization using planar helical antennas, the cross-section of the helix is usually square, which is undesirable for low-profile applications.

Surface wave antennas based on grounded dielectric slab are very promising candidates to meet the requirements of end-fire radiation, wide bandwidth, and low profile. In Chapter 3, the characteristics of surface waves propagating along the anisotropic grounded slab have been investigated first. Different from most open literature that generally assumes that the fields of surface waves are uniformly distributed along the transverse direction of the infinitely large grounded slab, we have thoroughly considered the field variations in the direction transverse to the wave propagation direction, and the corresponding propagation characteristics of surface waves have been derived on the basis of boundary conditions and Maxwell's equations. Three different methods have been compared to analyze the surface waves on grounded anisotropic slab as well as

to validate our derivations. In addition, the effects of the relative permittivity tensor on the propagation characteristics of surface waves have also been numerically studied. More importantly, our method has taken into account the finite-width grounded slab with the field distributions non-uniform along the transverse direction of the slab, which is more suitable for practical engineering.

Following the theoretical analysis of the surface waves on a grounded dielectric slab, a wideband surface wave antenna based on tapered grounded ceramic slab has been proposed, which can achieve nearly end-fire radiation over the frequency range from 6.1 to 18 GHz while maintaining the antenna thickness of only $0.12\lambda_0$ at the center frequency. The surface wave is radiated at the position of the tapered grounded slab where the slab thickness is equal to $\lambda/4\sqrt{\epsilon_r}$. A unidirectional surface wave launcher with a broad bandwidth has been realized by utilizing a probe-fed parallel-plate waveguide with a modified parabolic reflecting wall and a metallic post. The post has been employed close to the feeding probe to serve as a matching element, which is essential for improving the impedance matching of the launcher. The shape of the reflecting wall has been carefully modified so as to obtain a more uniform amplitude and phase distribution over a wide bandwidth. In addition, the procedures of designing proposed surface wave antenna have also been provided.

Next, an H-plane horn antenna has been proposed in Chapter 4, which can achieve a wider bandwidth from 3.4 to 18 GHz, and the antenna thickness is only 5.508 mm. In order to suppress the higher-order modes in the rectangular waveguide, the width and height of the waveguide have been carefully selected, and a pair of metallic posts has been introduced at positions where the maximum field of TE_{30} occurs. A cone-shaped probe and a metallic ridge have been employed in the coaxial-to-rectangular waveguide in order to improve the impedance matching as well as to enhance the operating bandwidth of TE_{10} mode. An ellipse-shaped copper taper and another pair of metallic posts have also been utilized in the H-plane horn antenna, which effectively smoothen the transition from the horn aperture to free space and improve the impedance matching of the antenna. Both planar and conformal versions of the proposed H-plane horn

antenna have been investigated in this chapter. In the conformal antenna, a tapered grounded substrate has been employed at the end of the structure, making the antenna suitable to be embedded into a cylindrical conducting platform. The effects of the platform on the conformal antenna performance have also been analyzed. In addition, the proposed conformal antenna has been extended into a conformal array configuration.

In Chapter 5, three cavity-backed supergain slot antennas have been proposed for end-fire radiation. In the first structure, we have employed a back-to-back cavity structure to achieve the end-fire radiation. In the other two conformal structures, the effects of platforms on the radiation beam angles have been carefully investigated, nearly end-fire radiation and end-fire radiation have been obtained using cylindrical and conical platforms, respectively. In the proposed structures, the bandwidth has been effectively widened by utilizing three slots with different lengths. The required magnitude and phase of the excited electric field for achieving supergain have been obtained by adjusting the inter-element spacing and the position of each slot. Furthermore, the array configuration of the proposed antenna with six separate antenna elements embedded into the conical platform has also been fabricated and tested. The array structure is immune to electromagnetic waves with arbitrary polarization, which is potentially very useful for airborne applications.

In addition to the above mentioned linearly polarized end-fire antennas, a planar helical antenna of circular polarization has been proposed in Chapter 6. The planar helix has been formed using printed strips and straight-edge connections implemented by plated via-holes. In the proposed antenna, the current on the ground plane plays an essential role in weakening the strong amplitude of the horizontal electric field generated by the one on the strips, while the vertical polarization is contributed by the current flowing along the via-holes. The operating principle of achieving circular polarization has been analyzed based on the current distribution of the proposed antenna with one-turn helix. Parametric studies have been conducted to analyze the influences of the key parameters on the antenna performance. Furthermore, the effects of the tapered

helix and conducting side-walls on the radiation pattern and axial ratio have been studied.

In summary, we have proposed a number of low-profile end-fire antennas. All antennas have overcome the existing problems of traditional end-fire antennas. Their wide bandwidth, high gain, end-fire radiation, and circular polarization have been verified by experimental results. The proposed antennas are very promising to be mounted on the surface of airborne applications and can be extended to conformal arrays.

7.2 Recommendations for Future Work

This thesis has presented several low-profile end-fire antennas for airborne applications with novel concepts and improved performance. However, it still offers many opportunities for us to have a further investigation. Some of the suggested studies are listed as follows.

- 1) In Chapter 3, a surface wave antenna employing tapered grounded ceramic slab is proposed. The surface wave is radiated at the position of the tapered grounded slab where the slab thickness is equal to $\lambda/4\sqrt{\epsilon_r}$. In the future, we may design surface wave antennas based on metasurface to achieve wide bandwidth and low profile. Characteristics of surface waves propagating along and radiating from metasurface should be carefully investigated. A gradient metasurface is expected to radiate the surface wave within a wide frequency range.
- 2) In Chapter 5, supergain antennas have been proposed using three closely spaced slots and a backed cavity. Generally speaking, more slots are desirable to be employed in order to take full advantage of the cavity space. However, it is very challenging to realize supergain in practice when employing more slots, because it will suffer from the severe sensitivity of the excitation amplitude and phase for each slot. In addition, when more slots are employed closely, the radiation resistance of the

antenna will decrease significantly, leading to a very low radiation efficiency. These problems may be addressed in the future.

- 3) Although the planar helical antenna proposed in Chapter 6 can achieve CP performance with end-fire radiation over a 34% frequency band, it may not be enough for some wideband applications. Our future investigation may focus on designing novel low-profile antennas that can achieve a wide 3-dB axial ratio fractional bandwidth of 100% in the end-fire direction. By utilizing the concept of frequency independent structure, it may be possible to realize the wideband CP performance.
- 4) When designing conformal antennas, the effects of the platform on the antenna performance should be taken into account. In our conformal designs in Chapters 4 and 5, parametric studies have been conducted to investigate the effects of platforms on the return loss and radiation pattern. In the future, the suitable theoretical analysis may be conducted to understand these effects.
- 5) The recent growth in the ambit of modern wireless communication has increased the demand for multi-band antennas that can satisfy the requirement of multiple functionalities. Designing multi-band antenna can also further reduce the production cost. Therefore, researching on conformal dual-band and multi-band end-fire antennas may be the topics for our further investigation.

APPENDIX A

DERIVATION OF THE FIELD DISTRIBUTIONS AND DISPERSION RELATIONS IN ANISOTROPIC SLAB

For the anisotropic slab with the relative permittivity tensor $\bar{\bar{\epsilon}}$ and relative permeability tensor $\bar{\bar{\mu}}$, the Maxwell's equations can be written as

$$\nabla \times \bar{E} = -j\omega\mu_0\bar{\bar{\mu}} \cdot \bar{H} \quad (\text{A.1a})$$

$$\nabla \times \bar{H} = j\omega\epsilon_0\bar{\bar{\epsilon}} \cdot \bar{E} \quad (\text{A.1b})$$

where $\bar{\bar{\epsilon}}$ and $\bar{\bar{\mu}}$ are expressed as (3.1). We assume that surface waves propagate along the +y direction with a propagation factor of $e^{-j\beta y}$. By solving the above two equations, the vector equations of all E-fields and H-fields can be expressed as

$$-j\beta_y E_z - \frac{\partial E_y}{\partial z} = -j\omega\mu_0\mu_x H_x \quad (\text{A.2a})$$

$$-j\beta_y H_z - \frac{\partial H_y}{\partial z} = j\omega\epsilon_0\epsilon_x E_x \quad (\text{A.2b})$$

$$\frac{\partial E_x}{\partial z} - \frac{\partial E_z}{\partial x} = -j\omega\mu_0\mu_y H_y \quad (\text{A.2c})$$

$$\frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} = j\omega\epsilon_0\epsilon_y E_y \quad (\text{A.2d})$$

$$\frac{\partial E_y}{\partial x} + j\beta_y E_x = -j\omega\mu_0\mu_z H_z \quad (\text{A.2e})$$

$$\frac{\partial H_y}{\partial x} + j\beta_y H_x = j\omega\epsilon_0\epsilon_z E_z \quad (\text{A.2f})$$

From (A.2b) and (A. 2e), E_x and H_z can be derived as

$$E_x = \frac{1}{j(\epsilon_x \mu_z k_0^2 - \beta_y^2)} [\beta_y \frac{\partial E_y}{\partial x} - \omega \mu_0 \mu_z \frac{\partial H_y}{\partial z}] \quad (\text{A.3a})$$

$$H_z = \frac{1}{j(\epsilon_x \mu_z k_0^2 - \beta_y^2)} [\beta_y \frac{\partial H_y}{\partial z} - \omega \epsilon_0 \epsilon_x \frac{\partial E_y}{\partial x}] \quad (\text{A.3b})$$

Similarly, H_x and E_z can be derived from (A.2a) and (A. 2f) as

$$H_x = \frac{1}{j(\epsilon_z \mu_x k_0^2 - \beta_y^2)} [\beta_y \frac{\partial H_y}{\partial x} + \omega \epsilon_0 \epsilon_z \frac{\partial E_y}{\partial z}] \quad (\text{A.3c})$$

$$E_z = \frac{1}{j(\epsilon_z \mu_x k_0^2 - \beta_y^2)} [\beta_y \frac{\partial E_y}{\partial z} + \omega \mu_0 \mu_x \frac{\partial H_y}{\partial x}] \quad (\text{A.3d})$$

Therefore, all field components can be found once E_y and H_y are determined. In addition, by substituting the corresponding field expressions into (A.2c) and (A.2d), the dispersion relations in anisotropic media as indicated in (3.10) can then be directly obtained. It should be mentioned that the field distributions and dispersion relations in air regions can be also derived following the same procedures.

APPENDIX B

DERIVATION OF SURFACE WAVE MODES UNIFORMLY DISTRIBUTED ON ANISOTROPIC GROUNDED SLAB

For the grounded slab with infinite width as shown in Figure 3.1, pure TM or TE modes can be supported if the field distribution of surface wave along the transverse direction of the slab is assumed to be uniform ($R_x = 0$).

For TM surface wave, substituting $B_I = B_{II} = 0$ into (3.5a)-(3.6d) and applying continuities of all the tangential fields across the interface lead to

$$k_z \tan(k_z t) = \varepsilon_y k_{oz} \quad (\text{B.1})$$

By solving Maxwell's equations based on (A.2a) and (A.2b), the dispersion relations in Regions I and II can be derived as

Region I

$$\beta_y^2 + \frac{\varepsilon_z}{\varepsilon_y} k_z^2 = \varepsilon_z \mu_x k_0^2 \quad (\text{B.2})$$

Region II

$$\beta_y^2 - k_{oz}^2 = k_0^2 \quad (\text{B.3})$$

For TE surface wave, $A_I = A_{II} = 0$. Similarly, the boundary conditions and the dispersion relations in the anisotropic medium lead to

$$k_z \cot(k_z t) = -\mu_y k_{oz} \quad (\text{B.4})$$

$$\beta_y^2 + \frac{\mu_z}{\mu_y} k_z^2 = \varepsilon_x \mu_z k_0^2 \quad (\text{B.5})$$

$$\beta_y^2 - k_{oz}^2 = k_0^2. \quad (\text{B.6})$$

APPENDIX C

DERIVATION OF SURFACE WAVE MODES BASED ON MARCATILI'S METHOD

In [140], Marcatili introduced an approximate method to describe how the guided waves propagate through rectangular dielectric waveguides. Based on his method, the surface waves are divided into E_z and E_x modes. For E_z mode, the field component E_z is strong, while H_z is assumed to be zero, which represents the quasi-TM mode in our discussion. Similarly, the E_x mode represents the quasi-TE mode, in which case the field E_x is strong and H_x is assumed to be zero. In addition, the fields in Regions V and VI as shown in Figure 3.2(b) are assumed to be negligible. Based on these assumptions, the quasi-TM and quasi-TE modes can be analyzed separately.

For quasi-TM mode, H_z and H_x of Regions I, II, III, and IV can be expressed as

$$H_z^I = H_z^{II} = H_z^{III} = H_z^{IV} = 0 \quad (C.1a)$$

$$H_x^I = A_I(e^{-R_x x} + r_1 e^{R_x x})e^{-j\beta_y y} \cos(k_z z) \quad (C.1b)$$

$$H_x^{II} = A_{II}(e^{-R_x x} + r_1 e^{R_x x})e^{-j\beta_y y} e^{-k_{0z} z} \quad (C.1c)$$

$$H_x^{III} = A_{III}e^{-R_{0x} x} e^{-j\beta_y y} \cos(k_z z) \quad (C.1d)$$

$$H_x^{IV} = A_{IV}e^{R_{0x} x} e^{-j\beta_y y} \cos(k_z z) \quad (C.1e)$$

Based on the fields H_z and H_x , all field distributions and dispersion relations in four regions can then be derived by solving Maxwell's equations.

The continuity conditions of E_y and H_x at $z = t$ give

$$k_z \tan(k_z t) = \varepsilon_y k_{0z} \quad (C.2)$$

In addition, E_z and H_y are continuous at $x = \pm a$, which leads to

$$R_{0x}\mu_y(\varepsilon_y\mu_x k_0^2 - k_z^2) \frac{e^{-R_x a} + e^{R_x a}}{e^{-R_x a} - e^{R_x a}} = \varepsilon_y\mu_x R_x(k_0^2 - k_z^2) \quad (C.3)$$

The dispersion relations in four regions are derived as

$$\beta_y^2 - \frac{\mu_x}{\mu_y} R_x^2 + \frac{\varepsilon_z}{\varepsilon_y} k_z^2 = \varepsilon_z\mu_x k_0^2 \quad (C.4)$$

$$\beta_y^2 - R_x^2 - k_{0z}^2 = k_0^2 \quad (C.5)$$

$$\beta_y^2 - R_{0x}^2 + k_z^2 = k_0^2 \quad (C.6)$$

It is seen that β_y , R_x , R_{0x} , k_z , and k_{0z} can be solved for given k_0 , t , $\bar{\varepsilon}$ and $\bar{\mu}$ based on (C.2)-(C.6).

For quasi-TE mode, H_z and H_x in Regions I, II, III, and IV can be expressed as

$$H_x^I = H_x^{II} = H_x^{III} = H_x^{IV} = 0 \quad (C.7a)$$

$$H_z^I = B_I(e^{-R_x x} + r_2 e^{R_x x})e^{-j\beta_y y} \sin(k_z z) \quad (C.7b)$$

$$H_z^{II} = B_{II}(e^{-R_x x} + r_2 e^{R_x x})e^{-j\beta_y y} e^{-k_{0z} z} \quad (C.7c)$$

$$H_z^{III} = B_{III}e^{-R_{0x} x} e^{-j\beta_y y} \sin(k_z z) \quad (C.7d)$$

$$H_z^{IV} = B_{IV}e^{R_{0x} x} e^{-j\beta_y y} \sin(k_z z) \quad (C.7e)$$

Similarly, based on Maxwell's equations and boundary conditions, five equations with five unknowns β_y , R_x , R_{0x} , k_z , and k_{0z} can be derived and they are summarized below

$$-k_z \cot(k_z t) = \mu_y k_{0z} \quad (C.8)$$

$$R_x(k_0^2 + R_0^2) \frac{e^{-R_x a} - e^{R_x a}}{e^{-R_x a} + e^{R_x a}} = R_{0x}\varepsilon_x(\varepsilon_y\mu_z k_0^2 + R_x^2) \quad (C.9)$$

$$\beta_y^2 - \frac{\varepsilon_x}{\varepsilon_y} R_x^2 + \frac{\mu_z}{\mu_y} k_z^2 = \varepsilon_x\mu_z k_0^2 \quad (C.10)$$

$$\beta_y^2 - R_x^2 - k_{0z}^2 = k_0^2 \tag{C.11}$$

$$\beta_y^2 - R_{0x}^2 + k_z^2 = k_0^2. \tag{C.12}$$

AUTHOR'S PUBLICATIONS

Journal papers

1. **Z. Chen** and Z. Shen, "Planar helical antenna of circular polarization," *IEEE Trans. Antennas Propag.*, vol. 63, no. 10, pp. 4315 - 4323, Oct. 2015.
2. **Z. Chen** and Z. Shen, "Wideband flush-mounted surface wave antenna of very low profile," *IEEE Trans. Antennas Propag.*, vol. 63, no. 6, pp. 2430-2438, Jun. 2015.
3. **Z. Chen** and Z. Shen, "A compact cavity-backed endfire slot antenna," *IEEE Antennas Wireless Propag. Lett.*, vol. 13, pp. 281–284, 2014.

Conference papers

- 1 **Z. Chen** and Z. Shen, "Conformal cavity-backed slot antenna embedded in a conical platform for end-fire radiation," *IEEE Antennas Propag. Soc. Int. Symp.*, 2017, Accepted.
- 2 **Z. Chen** and Z. Shen, "Design considerations of conformal wideband end-fire antenna mountable on cylindrical platforms," *Progress in Electromagnetics Research Symp. (PIERS)*, 2016.

- 3 **Z. Chen**, Z. Hu, Y. Zhao, and Z. Shen, “Wideband and low-profile end-fire antennas mountable on metallic platforms,” *IEEE Int. Conf. on Antenna Measurements Applications (CAMA)*, 2015, pp. 1 - 2.
- 4 **Z. Chen** and Z. Shen, “Broadband and low-profile H-plane ridged horn antenna,” *IEEE Antennas Propag. Soc. Int. Symp.*, 2015, pp. 2319 - 2320.
- 5 **Z. Chen** and Z. Shen, “A conformal cavity-backed supergain slot antenna,” *IEEE Antennas Propag. Soc. Int. Symp.*, 2014, pp. 1288–1289.

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