Low-profile vertically-polarized omnidirectional antennas for V2X communications

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**NANYANG
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SINGAPORE

**LOW-PROFILE VERTICALLY-POLARIZED
OMNIDIRECTIONAL ANTENNAS FOR V2X
COMMUNICATIONS**

LIU ANKANG

SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING

2018

**LOW-PROFILE VERTICALLY-POLARIZED
OMNIDIRECTIONAL ANTENNAS FOR V2X
COMMUNICATIONS**

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School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2018

Statement of Originality

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

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Authorship Attribution Statement

This thesis contains material from 2 paper(s) published in the following peer-reviewed journal(s) where I was the first and/or corresponding author.

Chapter 3 is published as A.K. Liu, Y.L. Lu and L. Huang, “Low-profile patch antennas with enhanced omnidirectional gain in the horizontal plane for DSRC applications,” *IET Microwaves, Antennas & Propag.*, vol. 12, Iss. 2, pp. 246-253, Feb. 2018.

The contributions of the co-authors are as follows:

- Prof LU provided the initial project direction and edited the manuscript drafts.
- I prepared the manuscript drafts. The manuscript was revised by Dr Huang Ling.
- I conducted the antenna simulation. Dr Huang Ling and I measured the antenna.

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The contributions of the co-authors are as follows:

- Prof Lu provided the initial project direction and edited the manuscript drafts.
- I wrote the drafts of the manuscript. The manuscript was revised together with Dr. Huang Ling.
- I performed all the antenna simulation and measurement.

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Abstract

For many such wireless systems such as vehicle-to-everything (V2X) communication, vertically-polarized omnidirectional antennas are desired for the full coverage over the horizontal plane and being placed on a large ground. For these applications, monopole antenna and its variations are the primary choices. The height of the-state-of-the-art monopole based antenna is still too high for many applications. The main objective of this study is to innovate new low-profile vertically-polarized omnidirectional antenna designs with wider bandwidth and higher gain for better V2X communications. Based on the literature survey, the circular patch antenna (CPA) and the monocone antenna are two popular and promising candidates for this study. Therefore, much of the study is devoted to innovative designs based on these two antenna types.

The first design is a low-profile circular-patch antenna with over 3 dB gain enhancement in the horizontal plane for the emerging DSRC (Dedicated Short Range Communications) application. To increase the omni-directional gain in the horizontal plane, six groups of magnetic dipole-like resonators are introduced around a shorted circular patch-ring antenna. These resonators help to guide the wave propagating along the horizontal surface. In addition, the number of resonators in each group can be adjusted to control radiation

patterns in the horizontal plane.

The second design is a wideband circular patch antenna with omnidirectional gain enhancement in the horizontal plane. The gain enhancement over 1 GHz bandwidth is achieved by introducing low-profile I-shaped loading structures without increasing the thickness of antennas. The effective refractive index of the new structure is larger than that of the conventional substrate with the same material, therefore the new structure behaves as a lens to concentrate the energy and enhance the omnidirectional radiation performance. In addition, the proposed antenna design can achieve a fractional bandwidth of 19% by combining the multiple resonant frequencies of the antenna structures.

The third design is an ultrawideband low-profile monocone antenna with dielectric loading. The operating modes of the proposed antenna are different from those of conventional monocone antenna while similar to those of the circular patch antenna. The first mode is studied with the aid of the LC resonant circuit. To further improve the matching properties of the monocone antenna, an innovative dielectric structure is introduced. The dielectric loading can effectively change the impedance of the monocone antenna and provide additional freedom for increasing the fractional bandwidth.

In summary, three new low-profile vertically-polarized omnidirectional antenna designs are presented in the thesis with the theoretical study, computer-aided design and optimization, prototype fabrication, measurement verification, and analysis study. The thesis is concluded by pointing out a few recommendations for future investigations.

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Chapter 1

Introduction

1.1 Background

Nowadays, the two biggest problems on our roads are safety and congestion due to the rapid auto industry development. The formulation of intelligent transport systems (ITS), which can significantly enhance driving safety and improve the traffic efficiency, has caught growing attention from national governments, large industry players and outstanding universities [1–5]. Vehicle-to-everything (V2X) communications refer to passing information from a vehicle to any entity in the ITS that may affect the vehicle, and vice versa. V2X communications, at first, included vehicle-to-infrastructure (V2I) and vehicle-to-vehicle (V2V) communications and now develop to a wider range involving additional vehicle-to-pedestrian (V2P) and vehicle-to-network (V2N) communications [6].

Two potential solutions can be applied to support V2X communications: dedicated short-range communication (DSRC) and cellular network tech-

Table 1.1: DSRC standards in North America, Europe and Japan [7]

Region	Bands (MHz)
North America	5850 - 5925
Europe	5470 - 5725, 5795 - 5815, 5855 - 5925
Japan	755.5 - 764.5, 5770 - 5850

nologies. DSRC is a two-way short-range communication technology that is generally applied to automotive and ITS applications while cellular network technologies refer to the well-known 3G, 4G and 5G et al. Although DSRC is an essential technology for V2X communications, there is no globally-agreed definition for DSRC and its standards differ in North America, Europe and Japan. Table. 1.1 shows the different DSRC bands in North America, Europe, and Japan [7].

Since the traffic accidents may happen in a few seconds, DSRC is very attractive for vehicle communications due to its short delay. Because DSRC communication needn't pass by the base station (BS). However, the "short rang" characteristic of DSRC is its limitation. If vehicles move at high speed, they need to continuously switch their connecting roadside units (RSU) for Internet access due to the small coverage. Besides, the allocated narrow DSRC band cannot satisfy the high traffic demand from in-vehicle Internet access.

On the contrary, unlike DSRC, cellular network technologies have high network capability, wide cellular coverage range and widely-constructed infrastructures [7] which benefit the V2X communications. However, despite these advantages, the data delay is harmful to the applications of cellular network on vehicle communications, especially the safety applications. The

data whose final receiver is vehicles, needs to be passed to BS firstly, which dramatically increases the transmission time.

Due to the respective disadvantages of DSRC and cellular networks, the two technologies cannot satisfy all requirements of V2X communications alone. Therefore, some DSRC-cellular hybrid architectures are proposed and designed for reliable and sufficient V2X communications [8–10]. These architectures combine the advantages of both DSRC and cellular communications. Vehicles applying a DSRC-cellular hybrid architecture can enjoy V2V communications with a short delay, high-speed data and reliable Internet access [7] at the same time.

1.2 Motivation

Antennas for V2X communications have some special considerations that should be taken into account. For example, vertical polarization and omnidirectional patterns are desired in V2X communications for the full coverage over the horizontal plane and the ability being placed on a large ground, e.g. the roof of the vehicles. Besides, low profile or being inserted in the roof is one essential characteristic for aesthetic, stable and space-saving purposes. Wideband antenna may be required for high data rate and the coverage of both DSRC band and cellular network band in the DSRC-cellular system for V2X communications. The widely-used monopole antenna is the primary choice at this stage for its low cost and easy fabrication. Based on literature review, circular patch antenna and monocone antenna, are also possible candidates for vehicle communications because of their simple structure and

easy fabrication. However, the three types of the state-of-the-art antennas cannot satisfy all these requirements due to their respective limitations.

1.2.1 Limitations of Monopole Antenna

In practice, a quarter-wavelength monopole mounted on a ground plane and fed by a coaxial line has been widely used [11]. However, the up to a quarter of wavelength profile is too high for some applications. One possible method to reduce the profile of monopole antenna is to bent partial conductor parallel to the ground while these antennas have poor omnidirectional radiating characteristics in the horizontal plane [12, 13]. Multi-elements are employed to reduce the ripples in the horizontal plane but make the antenna structure complicated and unstable [14]. Some researchers reduce the thickness of monopole antennas by loading a large top hat [15]. However, the radiation principle of these so-called monopole antennas is radiating through a magnetic slot, which is more like that of the circular patch antenna. In general, the low-profile monopole antenna is not the desired choice for vehicle communications.

1.2.2 Limitations of Circular Patch Antenna

Circular patch antennas are well known and widely used for their low cost, low weight, simple structure and easy installation [11]. But there are still some constraints of CPA applied to V2X communications.

1.2.2.1 Low gain in the horizontal plane

The low gain in the horizontal plane of CPA is mainly caused by two factors, the reflection of large ground and the low-profile in the vertical direction. The ground whose size is larger than that of CPA, reflects most energy upwards, leading to the low gain in the horizontal plane [16]. Besides, it is known that the gain is inversely proportional to the electrical size of the aperture [11]. Since the thickness of CPA is far smaller than the wavelength [17], the gain in the vertical plane is relatively small even under a symmetric case when the sizes of the ground and the patch are equal.

1.2.2.2 Narrow bandwidth

Conventional circular patch antennas have very narrow frequency bandwidth which is a few percents at most due to the high Q (sometimes larger than 100) [11]. It is found that the bandwidth can be improved through increasing the thickness and decreasing the dielectric constant of the substrate, whose effects, however, are limited [17]. On the other hand, V2X communications have multiband which cannot be covered by an antenna with such a narrow bandwidth. Therefore, some techniques should be employed to improve the bandwidth of CPA.

1.2.3 Limitations of Monocone Antenna

Generally speaking, the conventional monocone antenna is a transform of monopole antenna and, thus, not a low-profile structure. Some techniques like utilizing top hat and shorting pins have been applied to reduce the profile

of monocone antennas to below $0.08\lambda_{\text{low}}$ (λ_{low} refers to the wavelength at the lowest operating frequency in free space), but these antennas cannot achieve an over 140% fractional bandwidth [18,19]. Therefore, it is still a challenging task to design a low-profile structure with an ultra-wide bandwidth retained.

1.3 Objective

As mentioned above, two kinds of vertically-polarized omnidirectional antenna, circular patch antenna and monocone antenna, are promising candidates for the replacement of monopole antennas widely used in vehicle communications. However, these two types of antennas still suffer from various limitations such as low gain (CPA) and high thickness (monocone antenna). Much of the study is devoted to designing the new innovative antennas for alleviating these associated limitations. The main objectives of our research are summarized below:

- 1) Conduct a fruitful literature review covering the principles and the techniques for the associated limitations of CPA and monocone antenna. The review helps us understand the radiation theories, explore the principles of overcoming the limitations and accordingly propose innovative low-profile vertically-polarized omnidirectional antennas with novel techniques for V2X communications.
- 2) Propose circular patch antennas with gain enhancement in the horizontal plane. It is our target that the gain of the CPA in the horizontal plane is improved to 0 dB or higher, which should be comparable to

that of the monopole antenna. It is desirable that since V2X communications cover multiband, our proposed antenna could achieve a wide bandwidth while retaining the high gain in the horizontal plane. Besides, it is expected that our proposed CPAs can be inserted in the car roof without destroying the radiation performance, especially in the horizontal plane for aesthetic, stable and space-saving purposes.

- 3) Propose an ultrawideband low-profile monocone antenna for covering multiband. The proposed antenna is expected to have over 150% fractional bandwidth with a thickness of below $0.06\lambda_{\text{low}}$. Within the ultrawideband, the monopole-like radiation pattern should be achieved. In addition, the modes of the low-profile monocone antenna are expected to be investigated for explaining the radiating principles.
- 4) All the proposed antennas are expected to be fabricated and tested to verify the simulation result and the design principles.

1.4 Major Contributions of the Thesis

This thesis introduces three kinds of novel low-profile vertically-polarized omnidirectional antennas for wireless communications, especially V2X applications. Aligned with the objectives mentioned above, the proposed antennas are either CPA with gain enhancement in the horizontal plane or ultrawideband low-profile monocone antenna. The major contributions of this thesis are summarized as follows:

- 1) Propose two low-profile circular patch antennas with over gain enhance-

ment in the horizontal plane for DSRC applications. One novel method for gain enhancement, which is applying 6 groups of magnetic dipoles, is presented. The carefully designed magnetic dipoles resonate at a desired frequency and can guide the electromagnetic wave propagating along the horizontal plane. Parameters are also studied based on a circuit model, which offers a physical insight into the proposed antenna. Two antennas achieving omnidirectional and elliptically radiating patterns respectively are designed. Besides, the two antennas with perfect electric conductor (PEC) wall surrounded are simulated, fabricated and measured to verify that the proposed antenna can be inserted in the roof without destroying the high gain performance in the horizontal plane.

- 2) Propose a wideband circular patch antenna with gain enhancement in the horizontal plane through employing a novel artificial structure, “I-shaped” structure. The “I-shaped” structure is well designed so that the proposed structure has the refractive index larger than that of the conventional substrate. The artificial structure behaves as a lens to converge the energy and thus increase the antennas horizontal radiation gain by 0.5 dBi to 2 dBi. In addition, the bandwidth of the proposed antenna is increased to over 1 GHz through multiple resonances and one proximity feeding. Both simulation and measurement results show that even inserted in the vehicle roof, the proposed antenna has satisfied performance in terms of horizontal gain.

- 3) Propose an ultrawideband low-profile monocone antenna with dielec-

tric loading. It is found that the propagation modes of the low-profile monocone antenna in the antenna region are TM_{0N} mode, unlike that of the conventional monocone antenna. The first mode is studied with the aid of the epsilon-negative transmission line. To further improve the matching properties of the monocone antenna, an innovative method, dielectric structure, is proposed and discussed. The loaded dielectric can effectively change the impedance of the monocone antenna and provide additional freedom for increasing the fractional bandwidth. Both simulation and measurement results show that the proposed antenna with a thickness of $0.05\lambda_{low}$ can achieve an approximately 7.4:1 fractional bandwidth.

1.5 Organization of the Thesis

The thesis has been organized into 6 chapters. Chapter 1 introduces the background and the motivation of this thesis based on analyzing the requirements of V2X communications. The objective to overcome the limitations is presented in the following section. Finally, major contributions are summarized.

Chapter 2 reviews two kinds of vertically-polarized omnidirectional antennas, the CPA and the monocone antenna. In the beginning, the basic theory of the CPA is introduced based on the cavity model. Some effective methods are reviewed for improving the bandwidth and the gain of CPA. This chapter also presents the fundamentals of monocone antenna covering input impedance calculation, as well as near field and far field solutions. Besides,

the existing method for overcoming the challenges of monocone antenna like high thickness and high sidelobes are also introduced.

In chapter 3, low-profile patch antennas with enhanced horizontal omnidirectional gain for DSRC applications are proposed. Six groups of magnetic dipoles around the center-fed patch are introduced to increase the gain. Then parameters are studied based on a circuit model. Two antennas achieving omnidirectional and quasi-elliptical radiation patterns respectively are simulated, fabricated and measured.

In chapter 4, a wideband circular patch antenna with an I-shaped structure for horizontal omnidirectional gain enhancement is proposed. In this chapter, a type of I-shaped structure with the refractive index larger than the conventional substrate in a wide bandwidth is first proposed. Through loading the I-shaped structure, the gain of CPA in the horizontal plane can be increased up to nearly 2 dBi in 1 GHz bandwidth. The techniques for achieving wideband CPA are also presented. A prototype of the proposed antenna with gain enhancement is fabricated and measured for verifying our design and principles.

Chapter 5 first investigates the propagation modes of the low-profile monocone, unlike those of the conventional monocone antenna. Then a low-profile monocone antenna with dielectric loading is proposed. Next the methods for achieving ultrawide bandwidth and reducing the thickness of the antenna are presented. Simulated and measured results are both obtained and compared to verify the principles of the concepts.

Chapter 6 gives conclusions of the thesis and recommendations for future research.

Chapter 2

Literature Review

2.1 Introduction

As mentioned in Chapter 1, vertically-polarized omni-directional radiation patterns are desired for V2X communications. Due to the relatively low profile compared to the monopole antennas, circular patch antenna and monocone antenna are two promising candidates for this application. As an underlying basis for the studies in subsequent chapters, a fruitful review on both circular patch antennas and monocone antennas is conducted in this chapter. Section 2.2 introduces the fundamentals of the CPA as well as the method to increase CPA's bandwidth and gain. Section 2.3 firstly introduces radiation patterns, input impedance, as well as far field and near field solutions of the monocone antenna. Furthermore, the drawbacks of the monocone antenna are introduced and the corresponding techniques for overcoming these weaknesses are reviewed.

Figure 2.1: (a) Geometry of circular patch antenna; (b) cavity model and equivalent magnetic current.

2.2 Circular Patch Antenna

Circular patch antenna shown in Fig. 2.1(a) has received a lot of attention not only as a single element [20–26] but also in arrays [27–29] for its low profile, low cost and easy fabrication [20, 30]. Much work has been done to improve the bandwidth and increase the gain of the circular patch antenna.

2.2.1 Fundamentals of Circular Patch Antenna

The most popular methods for the analysis of the circular patch antenna are cavity model [31–33] and full wave model [34–36]. Only cavity model is applied to analyse the circular patch antenna in this chapter since this model gives the physical insight into the circular patch antenna and is relatively accurate. A cavity model is shown in Fig. 2.1(b) which is composed of two perfect electric conductor (PEC) bases at the bottom and the top to represent the ground and the patch respectively, and a cylindrical perfect magnetic conductor (PMC) periphery of the cavity [30]. Substrate is assumed to be

truncated beyond the extent of the patch.

Since the height of a CPA is very small ($h \leq \lambda_{\text{low}}$, λ_{low} refers to the wavelength at lowest operating frequency in the free space), the variation of the fields along the z direction can be ignored. Besides, due to the extremely low profile of circular patch antennas, the fringing of the fields at the edges of the patch is very small whereby the electric field is normal to the patch. Therefore, TM modes are the main modes only which will be considered for the low-profile CPA. The fields of the TM modes can be written as [30]:

$$\begin{aligned}
 E_\rho &= -j \frac{1}{\omega \mu \epsilon} \frac{\partial^2 A_z}{\partial \rho \partial z} \\
 E_\phi &= -j \frac{1}{\omega \mu \epsilon} \frac{1}{\rho} \frac{\partial^2 A_z}{\partial \phi \partial z} \\
 E_z &= -j \frac{1}{\omega \mu \epsilon} \left(\frac{\partial^2}{\partial z^2} + k^2 \right) A_z \\
 H_\rho &= \frac{1}{\mu} \frac{1}{\rho} \frac{\partial A_z}{\partial \phi} \\
 H_\phi &= -\frac{1}{\mu} \frac{\partial A_z}{\partial \rho} \\
 H_z &= 0
 \end{aligned} \tag{2.1}$$

which are subject to the boundary conditions of:

$$\begin{aligned}
 E_\rho(0 \leq \rho' \leq a; 0 \leq \phi' \leq 2\pi; z' = 0) &= 0 \\
 E_\rho(0 \leq \rho' \leq a; 0 \leq \phi' \leq 2\pi; z' = h) &= 0 \\
 H_\phi(\rho' = a; 0 \leq \phi' \leq 2\pi; z' = 0) &= 0
 \end{aligned} \tag{2.2}$$

The magnetic vector potential A_z can be expressed as [30]:

$$A_z = B_{mnp} J_m(k_\rho \rho') [A_2 \cos(m\phi') + B_2 \sin(m\phi')] \cos(k_z z') \tag{2.3}$$

Table 2.1: Roots of $J_m(x) = 0$

Modes(m, n)	0,0	1,1	2,1	0,2	3,1
χ'_{mn}	0	1.84118	3.05424	3.83171	4.20119

with the constraint of

$$(k_\rho)^2 + (k_z)^2 = (k_r)^2 = (\omega_r)^2 \mu \epsilon \quad (2.4)$$

$J_m(x)$ represents the Bessel function of the first kind of order m , and

$$k_\rho = \frac{\chi'_{mn}}{a} \quad (2.5)$$

$$k_z = \frac{p\pi}{h} \quad (2.6)$$

where $m = 0, 1, 2, \dots$, $n = 0, 1, 2, \dots$, $p = 0, 1, 2, \dots$ and χ'_{mn} represents the zeros of the derivative of the Bessel function $J_m(x)$. A few first values of χ'_{mn} are listed in Table. 2.1 in ascending order.

As mentioned above, the height of the most CPA is very small (typically $\leq 0.05\lambda_{\text{low}}$) and hence the variation of the fields along z is very small which means $p = 0$ and $k_z = 0$ in (2.6). Therefore, the resonant frequency for the TM_{mn0} can be written using (2.5) as [30]:

$$(f_r)_{mn} = \frac{1}{2\pi\sqrt{\mu\epsilon}} \left(\frac{\chi'_{mn}}{a} \right) \quad (2.7)$$

However, the calculation of the resonant frequencies for the TM_{mn} modes by (2.7) doesn't take the fringing into consideration. Due to the fringing at the edges, the electrical size of the patch is larger than its physical size.

Hence, a correction is introduced by using an effective radius a_e , to replace the radius a in (2.7), which is given in [20]:

$$a_e = a \left\{ 1 + \frac{2h}{\pi a \epsilon_r} \left[\ln\left(\frac{\pi a}{2h}\right) + 1.7726 \right] \right\}^{1/2} \quad (2.8)$$

When a frequency is given, an approximation to the solution of the patch radius of a is to find a_e using (2.7) and to substitute that into (2.8) for a_e and for a in the logarithmic function. Doing this leads to

$$a = \frac{F}{\left\{ 1 + \frac{2h}{\pi F \epsilon_r} \left[\ln\left(\frac{\pi F}{2h}\right) + 1.7726 \right] \right\}^{1/2}} \quad (2.9)$$

where

$$F = \frac{1}{2\pi\sqrt{\mu\epsilon}} \left(\frac{\chi'_{mn}}{f} \right) \quad (2.10)$$

2.2.2 Increase the Bandwidth of Circular Patch Antenna

As reviewed, one of the major disadvantages of the circular patch antenna is narrow bandwidth which is typically only a fractional of several percents [11].

Since 1980s, many studies have been done to overcome the disadvantage of the narrow bandwidth of the microstrip patch antenna. It is pointed out that the bandwidth can increase with higher thickness and lower substrate permittivity in [37]. Based on the theory in [37], a gap is introduced below the patch to increase the height of the antenna and thereby reduce the effective permittivity of the cavity below the patch [38, 39]. In [40], the patch antenna with a thickness of $0.035\lambda_c$ achieves 24% fractional bandwidth

Figure 2.2: (a) Circular patch antenna with a gap between the probe and the patch [43]; (b) circular patch antenna with circular slot [45].

through adding four resonators at the sacrifice of the antenna size (quadruple increase). In this thesis, λ_c refers to the wavelength of the center frequency of the operating bandwidth.

However, for the thick patch antennas in [37–39, 41], high probe inductance causes the mismatching between the patch and the input connector [42]. For compensating the probe inductance, the use of kinds of capacitive gaps is introduced [43–45]. A gap can be generated with a thin layer substrate spacing the patch from the probe as shown in Fig. 2.2(a) [43] or sculptured on the patch as a circular slot as shown in Fig. 2.2(b) [45].

As shown in Fig. 2.3, L-shaped probe, which is an effective structure for improving the bandwidth of microstrip patch antenna, is firstly presented in [46] by K. M. Luk *et al.* The thickness of the CPA is only $0.1\lambda_c$, but the antenna with a compact structure has a bandwidth up to over 24%. L-shaped probe caught much attention immediately after being invented and

Figure 2.3: Circular patch antenna with L-shaped probe [50].

Figure 2.4: Circular patch antenna for pattern reconfiguration [60].

was the most impressive method of improving the bandwidth of the circular patch antenna in the next ten years. K. M. Luk *et al.* made plenty of ex-

Figure 2.5: Wideband circular patch ring antenna [62].

cellent work on employing L-shaped probe [25, 47–57]. These studies show that the L-shaped probe can be applied to increase the bandwidth of various kinds of patch antennas like circular patch antenna with conical-pattern radiation [25], circularly-polarized patch antenna [49], and patch antenna for pattern reconfiguration [56] (shown in Fig. 2.4), as well as dual-band patch antenna [53]. Though L-shaped probe can effectively increase the bandwidth of antennas, the thickness of these antennas in [25, 47–57] is generally over $0.1\lambda_{\text{low}}$, which is still too high and impairs the benefits of low profile of the CPA.

In recent years, it becomes popular that multi-resonance characteristic is excited for CPA as this method can markedly increase the bandwidth of the circular patch antenna when the low-profile property of CPA can be maintained. Kinds of slots, for example U-slot [58], half U-slot [59], rectangular U-slot [60] and V-slot [61], can introduce extra resonance and effectively increase the bandwidth of the microstrip patch antenna. In 2009, Al-Zoubi et

Figure 2.6: A low-profile and broadband microstrip monopolar patch antenna at TM_{01} and TM_{02} mode [64].

al. proposed a circular patch ring antenna on a substrate with a thickness of only $0.028\lambda_{\text{low}}$ shown in Fig 2.5 [62]. The antenna achieves 18% fractional bandwidth through adapting TM_{02} mode of the circular patch and an additional mode excited by a coupling ring. Similarly, a patch antenna with dual-ring couplers has a 27.1% bandwidth by generating three resonant frequencies [63]. In [64], TM_{01} mode and TM_{02} mode which have monopole-like radiation patterns, are excited simultaneously. The proposed antenna with a thickness of $0.024\lambda_{\text{low}}$ shown in Fig. 2.6 has a bandwidth of 18% fractional bandwidth. In addition, feeding open loops [65] and parasitic patch [66] can also increase the number of resonant frequencies to enhancing the matching properties of CPA.

2.2.3 Increase the Gain of Circular Patch Antenna

The low gain of circular patch antennas with either conical radiation patterns or broadside radiation patterns is another limitation of CPA. Most of the studies are devoted to increasing the gain of CPA with conical radiation patterns and several methods have been proposed to overcome this drawback.

One widespread effective approach is reducing the surface wave along the horizontal plane [67–74]. In [67], the surface wave is firstly suppressed by a superstrate material as a cover for maximizing the gain. Besides, it is found that the surface won't be excited if the substrate is thin enough in [68], which has been well known. Another method for reducing the surface wave is applying the synthesized substrate that lowers the effective dielectric constant of the substrate either under or around the patch [70,72]. In addition, shorted annular ring [69,74] and parasitic elements [71,73] can be applied to reduce the surface wave.

Electromagnetic bandgap (EBG) structure is such a structure that has the ability to open a bandgap to forbid the propagation of electromagnetic waves. Therefore, applying EBG is another effective method to reduce the surface wave of the CPA in the substrate. Such an idea is first proposed and applied by D. Sievenpiper on a CPA in [75]. Based on this approach, the performance of the CPA is improved using a new cylindrical electromagnetic bandgap substrate as shown in Fig. 2.7 [76]. A cylindrical electromagnetic bandgap substrate, based on the mushroom-like substrate, can suppress the surface wave and increase the gain. For further increasing the gain of the patch antenna, there are many kinds of EBG structures employed to patch

Figure 2.7: CPA incorporated with the cylindrical EBG substrate [76].

antennas [77–79]. However, it is very difficult for these patch antennas to achieve the gain higher than 8 dBi.

Another technique to achieve the high gain of the patch antenna is by exciting two modes at the same band. The approach is firstly proposed by P. Juyal et al. by stacking a parasitic patch on the driven patch [80]. Fig. 2.8 shows that the two patches work at the TM_{11} and the TM_{13} modes respectively. Similarly, in [81] the TM_{12} mode is applied to suppress the sidelobe of the TM_{14} mode to enhance the antenna’s gain. What is more attractive is that the antenna is designed on a single layer, which reduces the profile of the antenna compared to the antenna in [80]. But the two antennas [80, 81] are both narrow bandwidth. As shown in Fig. 2.9, P. Squadrito and S. Zhang

Figure 2.8: High-gain single-feed dual-mode microstrip disc radiator [80].

et al. proposed an antenna whose unwanted modes can be excited not efficiently and thus will not distort the gain, sidelobe level and impedance bandwidth [82]. The matching property of the antenna can be improved by moving the position of the feeding.

Unlike increasing the gain of the CPA with conical radiation patterns, much less work has been done to increase the gain of the CPA with broadside radiation patterns, especially in the horizontal plane. As shown in Fig. 2.10, the circular patch antenna with corrugated ground was proposed with a diameter of $3.2\lambda_{\text{low}}$ and a thickness of $0.135\lambda_{\text{low}}$ [83]. The corrugated ground can guide the electromagnetic wave along the horizontal plane and increase the gain in the horizontal plane to 0.55 dB. The size of the CPA in [83] is too large for some applications. To reduce the size of the CPA with gain enhancement, a shorted ring structure (Fig. 2.11) can increase the gain of a textile CPA in the horizontal plane to -0.55 dB in [84], which is even lower than that of the antenna in [83]. Besides, the antenna size is still large, which has a diameter of $1.98\lambda_{\text{low}}$ and a thickness of $0.058\lambda_{\text{low}}$. Although the two

Figure 2.9: Wideband low-profile CPA with high gain and sidelobe suppression [82].

antennas in [83,84] introduced above can increase the gain in the horizontal plane to around 0 dBi, they both have two drawbacks, narrow bandwidth and large size, which are waiting for further investigation.

2.3 Monocone Antenna

Monocone antenna, which is half of the biconical antenna on an infinite ground, can be regarded as a transformation of the monopole antenna. It is pointed that *the bandwidth of an antenna (which can be closed within a sphere of radius a) can be improved only if the antenna utilizes efficiently, with its geometrical configuration, the available volume within the sphere* [11]. As shown in Fig. 2.12, apparently, the monopole antenna poorly utilizes the volume within the sphere with a radius of r , while the utilizing-volume efficiency of monocone antenna is significantly improved. Therefore, the band-

Figure 2.10: High-gain circular patch antenna with corrugated antenna [83].

width of the monocone antenna is much wider than that of the monopole antenna. Monocone antenna is one popular and promising candidate for its wide bandwidth and relatively low profile.

2.3.1 Electromagnetic Characteristics of Wide-Angle Monocone Antenna

The characteristics of the wide-angle monocone antenna with $\theta_0 \geq 30^\circ$ are reviewed for its widespread utilization. In the antenna region $\rho_l \leq r \leq a$, $\theta_0 \leq \theta \leq \pi/2$, it can be assumed that only TEM mode as the dominant

Figure 2.11: High-gain textile circular patch antenna [84].

Figure 2.12: (a) monocone antenna; (b) monopole antenna.

mode exists in the antenna region and higher complementary modes can be neglected in comparison to the TEM mode [85,86].

2.3.1.1 Input Impedance of Monocone Antenna

Based on the above-mentioned approximation, the input impedance of a monocone antenna with a length of a and a flare angle of θ_0 is expressed as [86]:

$$Z_{in} = Z_0 \frac{1 - \beta/\alpha}{1 + \beta/\alpha} \quad (2.11)$$

where

$$Z_0 = 60 \ln \cot \frac{\theta_0}{2} \quad (2.12)$$

$$\frac{\beta}{\alpha} = e^{-2jka} \frac{1 + j \frac{60}{Z_0} \sum_{n=1}^{\infty} \frac{2n+1}{n(n+1)} [P_n(\cos \theta_0)]^2 \xi_n(ka)}{-1 + j \frac{60}{Z_0} \sum_{n=1}^{\infty} \frac{2n+1}{n(n+1)} [P_n(\cos \theta_0)]^2 \xi_n(ka)} \quad (2.13)$$

and

$$\xi_n(ka) = \frac{h_n^{(2)}(ka)}{h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka)} \quad (2.14)$$

n is odd integral values. $h_n^{(2)}$ is the spherical Hankel function of the second kind, and $P_n(\cos \theta)$ is the Legendre polynomial of order n . Z_0 is the characteristic impedance of the monocone antenna. β/α is the ratio of the amplitudes of the reflected and the outwardly propagating TEM waves in the antenna region. $\xi(ka)$ is a complex auxiliary function of the real variable of ka .

2.3.1.2 Electric Fields of Monocone Antenna

Since only TEM mode needs to be considered in the antenna region, the θ component of the electric field in the antenna region, which is the dominant component of the electric field, can be expressed as [87]:

$$E_\theta(r, \theta) = \sqrt{\frac{\mu}{\epsilon}} \frac{\alpha e^{-jkr} - \beta e^{jkr}}{r \sin \theta} \quad (2.15)$$

where α and β are the constants of the outwardly propagating and the reflected TEM wave respectively as mentioned in equations (2.13).

In the far field ($kr \rightarrow \infty$), the E_θ is given by the following expression [87]:

$$E_\theta(r, \theta) = K \frac{e^{-jkr}}{r} \sum_{n=1}^{\infty} E_n(r, \theta) \quad (2.16)$$

where

$$E_n(r, \theta) = P_n(\cos \theta_0) P_n^1(\cos \theta) \frac{2n+1}{n(n+1)} \frac{i^n}{h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka)} \quad (2.17)$$

and

$$K = \sqrt{\frac{\mu}{\epsilon}} \frac{\alpha e^{-jkr} - \beta e^{jkr}}{r \sin \theta} \quad (2.18)$$

All these variables and functions are defined as those in Sec. 2.3.1.1. Through dividing equation (2.16) by $(E_\theta(r, \pi/2))$, the normalized electric distribution function can be obtained [87]:

$$\begin{aligned} N(\theta) &= \frac{E_\theta(r, \theta)}{E_\theta(r, \pi/2)} \\ &= \frac{\sum_{n=1}^{\infty} P_n(\cos \theta_0) P_n^1(\cos \theta) \frac{2n+1}{n(n+1)} \frac{i^n}{h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka)}}{\sum_{n=1}^{\infty} P_n(\cos \theta_0) P_n^1(0) \frac{2n+1}{n(n+1)} \frac{i^n}{h_{n-1}^{(2)}(ka) - \frac{n}{ka} h_n^{(2)}(ka)}} \end{aligned} \quad (2.19)$$

2.3.2 History of Monocone Antenna

The biconical antenna coming into people's notice could date back to the late of nineteenth century. Such an antenna was firstly proposed by Sir Oliver Lodge in 1899 [88]. Due to the lack of practical applications, the biconical antenna was ignored until the late of the 1930s, when Philip Carter rediscovered this type of antenna [89]. In the following several years, many

Figure 2.13: Biconical antenna with reduced size [91].

theoretical studies have been made as reviewed in Section 2.3.1, which covers radiation principles, input impedance calculation as well as far field and near field solutions of the monocone antenna [85–87, 90].

Considering the high thickness of monocone antennas, the design of low-profile monocone antennas with wide bandwidth is a hot topic in recent years. As shown in Fig. 2.13, the techniques, top hat loaded, shorting pins and low-cost plastic molding process, which have been widely employed at present, are firstly discussed for reducing the size of biconical antenna in [91]. The proposed antenna achieved a height reduction of 60% with a bandwidth from 3.1 GHz to 10.6 GHz. With a top-hat loading and shorting to the ground, a antenna for base station application is proposed in [18] and the antenna could achieve the height of $0.071\lambda_{\text{low}}$. In addition, one dual-sleeve structure could enhance the fractional bandwidth to about 137% through introducing another two resonant modes. For further reducing the thickness of the monocone antenna, a wideband antenna with a shorted top hat is proposed with a height of $0.068\lambda_{\text{low}}$ for unmanned aerial vehicles in [92].

Figure 2.14: Design of biconical antenna with abrupt structure [98].

Then it is found that capacitive loading [19], matching layer [93] and shallow cavity [94] can further reduce the thickness or the size of the monocone antenna, however, at the sacrifice of the wide bandwidth.

In addition, the high side-lobes of the monocone antenna at high frequency also greatly limit its applications. And it catches much attention to overcome this drawback. A structure combining a biconical antenna with a dielectric material creates a frequency independent main-lobe in the E-plane and suppress side-lobes [95]. This technique with various dielectric structures is successfully employed to ensure the frequency independent radiation characteristics in [96, 97]. However, the dielectric loading introduces dielectric losses and high reflections at the dielectric-to-air transitions which reduce antenna's efficiency. In [98], it is found that abrupt structure which is shown in Fig. 2.14, can reduce the side-lobes and create a frequency independent main lobe.

2.4 Summary

In this chapter, an overview of two kinds of the low-profile vertically polarized antennas, the CPA and the monocone antenna is conducted including their fundamentals and some techniques for solving their drawbacks. The basic theories of the CPA and the monocone antenna like radiation principles and field distributions lay the theoretical foundation for the corresponding antenna design. The methods of the CPA for increasing the bandwidth and the gain have been reviewed. In addition, the techniques of the monocone antenna to overcome its drawbacks like high thickness and high side lobes are also introduced.

Although the reviewed techniques can alleviate the limitations of the existing CPA and monocone antenna in some degree, there still exist a lot of challenges waiting for further exploration. Concerning the CPA, a little work has been reported to increase the omnidirectional gain in the horizontal plane of the CPA. This topic is very challenging due to the reflection effects of the ground and the small aperture in the vertical direction. Even current work has the ability to increase the gain in the horizontal plane, which, however, is still quite low. Besides, there are two significant drawbacks of the existing work, narrow bandwidth and large size. As for the monocone antenna, it is tough to reduce the thickness to below $0.07\lambda_{\text{low}}$ while the antenna can still enjoy the benefits of wide bandwidth because of the high Q value of low-profile antennas. Overcoming all these challenges are our objectives in the following chapters.

Chapter 3

Low-profile Patch Antennas with Enhanced Horizontal Gain

3.1 Introduction

DSRC technology is an emerging and promising technology enabling cars to communicate with other devices like other vehicles, traffic lights and roadside units [99] and vice versa. DSRC contributes to reducing road accidents and improving time and energy, efficiency especially for self-driving vehicles. For DSRC systems applied for V2X communications, the communication path is mainly built in the horizontal plane. Therefore the gain in $\theta = 90^\circ$ should be large enough to ensure reliable communication. Besides, a low-profile antenna is more desired for aesthetic purpose while maintaining smooth radiation in the horizontal plane (less than 3 dB).

The circular patch antenna which has a low profile and the monopole-like radiation patterns is another more promising candidate for the vehi-

cle antenna. In [71, 100–104], patch antennas were studied and they provided monopole-like radiation patterns and vertical polarization. However, for these antennas, the gains in the horizontal plane are all below -5 dB.

As reviewed in Chapter 2, although a lot of attention is paid to improve the gain of the broadside radiation pattern, few works aim to increase the gain of the conical radiation pattern, especially in the horizontal plane. In [83] circular patch antenna with corrugated ground for guided wave was proposed with a diameter of $3.2\lambda_{\text{low}}$ and a thickness of $0.14\lambda_{\text{low}}$. For reducing the size of the CPA with the gain enhancement for practical application, a textile antenna with surface wave enhancement is proposed in [84]. On the one hand, the radius of the antenna is effectively reduced by 35.3% and the thickness is decreased by 57.1% compared to the antenna proposed in [83], but on the other hand, the gain in the horizontal plane of the antenna in [84] is only about -0.5 dBi.

In this paper, a low-profile antenna with higher gain in the horizontal plane and reduced size compared to the antennas in [83, 84] is proposed for DSRC applications. Shorted circular patch-ring antenna (SCPRA) as shown in Fig. 3.1(a) is firstly proposed based on circular patch-ring antenna (CPRA) in [62]. Due to the low horizontal gain of CPRA, 12 shorting pins are applied in SCPRA to increase the horizontal gain by 0.5 dB. To further increase the gain in the horizontal plane, half-width microstrip lines with two flanks shorted and one edge open were employed, which work as magnetic dipole antennas [105], to guide the wave along the horizontal surface in our proposed antenna. The half-width microstrip line works in EH_1 mode and details of the half-width microstrip line have been analyzed by Liu et al., in [106]. The gain

enhancement in the horizontal surface ($\theta = 90^\circ$) is 3.10 dB from -2.04 dBi to 1.06 dBi with the aid of the gain enhancement elements. Besides, to better understand the effects of the annular ring and the conductive pins on the matching property, we use the equivalent circuit method to analyze the SCPRA. The proposed antennas are designed using FR4, a low-cost printed circuit board (PCB), with a diameter of $1.65\lambda_{\text{low}}$ (84 mm) and a thickness of $0.03\lambda_{\text{low}}$ (1.6 mm) at 5.9 GHz.

An omnidirectional vertically polarized circular patch antenna is firstly presented and the horizontal gain improves markedly compared to the conventional circular patch antenna without gain enhancement elements. Moreover, in DSRC system, the communication distance between front and back vehicles in the same lane/different lanes is supposed to be much longer than that between vehicles in different lanes when they are side by side. Therefore, elliptical radiation patterns in horizontal plane are more desirable than omnidirectional radiation patterns in some cases. The number of the gain enhancement elements of two opposite groups on the proposed antenna is reduced to achieve quasi-elliptical radiation patterns for these cases.

3.2 Antenna Design and Operating Principles

3.2.1 Antenna's Configuration

As illustrated in Fig. 3.1, the circular patch antenna is center-fed by a coaxial probe with a radius of 0.64 mm. An annular ring is concentrically coupled

Figure 3.1: Structures of SCPRA and the proposed omnidirectional circular patch antenna with gain enhancement: (a) Top view of SCPRA; (b) Top view of the proposed omnidirectional circular patch antenna with gain enhancement; (c) Cross view of the proposed omnidirectional circular patch antenna with gain enhancement.

Table 3.1: Parameters of the proposed omnidirectional antenna

Parameters	Value	Parameters	Value
R_1	13.4 mm	R_2	14.7 mm
R_3	19.6 mm	R_g	42 mm
h	1.6 mm	R_r	6.8 mm
R_e	6.4 mm	S_0	0.3 mm
S_1	0.7 mm	S_2	1.5 mm
α_1	66.4°	D_s	1.8 mm

to the circular patch, which improves the antenna's matching characteristics. Besides, the ring increases the distance between the gain enhancement elements and the center, therefore, the number of groups of gain enhancement elements can be increased. 12 shorting pins are applied on the patch to increase the horizontal gain by 0.5 dB. As introduced in Chapter 2, the theoretical radius of the centric circular patch working at the TM_{02} mode is calculated by (2.9).

In order to further increase the gain in the horizontal plane, 6 groups of gain enhancement elements are added around the ring. The gain enhancement element with three edges shorted and one long edge open behaves like a magnetic element providing vertical polarization and enhanced gain in the horizontal plane [105]. The three shorted edges are realized by closely-spaced shorting pins. It should be noted that during the antenna optimization, the shorting pins can be replaced by PEC walls to reduce the simulation time [107]. Through the optimization of the length and width, the gain enhancement element can resonate at the desired frequency to guide the TM wave along the surface and thus increase the gain in the horizontal plane.

The gaps S_0 , S_1 and S_2 are also critical parameters in the antenna design.

Figure 3.2: Structures of curved and straight gain enhancement elements: (a) curved gain enhancement element; (b) straight gain enhancement element.

S_0 should be small for avoiding large energy leakage from the gap when the antenna matching property is maintained. S_1 and S_2 should not be larger than the thickness of the substrate to realize strong coupling between the gain enhancement elements and the ring. The details of the antenna parameters are shown in Table 3.1.

3.2.2 Gain Enhancement Element

In this section, the gain enhancement element is analyzed based on both the calculated and the simulated results. Basically, the gain enhancement element can be regarded as a half-width microstrip line in EH_1 mode with two flanks shorted [106, 108] as shown in Fig. 3.2. It works as a magnetic dipole, and thus the length of the half-width microstrip line L_e is $\lambda_x/2$ in the

Figure 3.3: Normalized E_z versus varying distance from the source.

x direction (Fig. 3.2(b)), where the wavelength λ_x can be expressed as:

$$\lambda_x = 2\pi/\beta \quad (3.1)$$

To reduce the complexity, the straight gain enhancement element (Fig. 3.2(b)) is firstly analyzed rather than the curved one (Fig. 3.2(a)).

The propagation wavenumber for half-width microstrip lines in EH_1 have been fully analyzed in [106]. At the resonant frequency 5.9 GHz, the phase constant calculated by [106] based on the Liu-Oliner's method [108] is

$$\beta = 1.06k_0 \quad (3.2)$$

Therefore, the calculated wavelength at 5.9 GHz is 47.97 mm.

To verify the calculated results above and offer a simple way to obtain the gain enhancement element's length, a 200 mm half-width microstrip line (much longer than 1 wavelength) is simulated in HFSS and the normalized E_z versus varying distance from the source is shown in Fig. 3.3. The simulated

wavelength at 5.9 GHz can be calculated by:

$$\lambda_x = \frac{200 \times (0.762 - 0.036)}{3} = 48.4 \text{ mm} \quad (3.3)$$

The simulated wavelength λ_x is 48.40 mm which is close to the calculated value.

The width R_e can be varied to make the gain enhancement element resonate at the desired frequency when the length L_e is fixed at $\lambda_x/2$. We use HFSS with “Eigenmode” solution type to calculate the resonant frequency of the gain enhancement element. It is found that when $R_e = 5.7$ mm, the gain enhancement element resonates at 5.9 GHz.

The parameters of the straight gain enhancement element have been solved, while only the curved elements are suitable for our proposed antenna. The propagation wave number of the curved half-width microstrip line approaches that of the straight half-width microstrip line [109]. Therefore, the arc length of the curved element and the length of the straight element have close values; the angle of curved gain enhancement element α_1 can be estimated by:

$$\alpha_1 = \frac{360^\circ L_e}{2\pi R_c} \quad (3.4)$$

In addition, the width R_e of the curved gain enhancement element after optimization in the final design is 5.4 mm which is close to that of the straight element (5.7 mm).

Figure 3.4: E-field distributions in the vertical plane at 5.9 GHz: (a) the proposed antenna; (b) SCPRA.

3.2.3 Gain Enhancement Principles

To verify the principles of enhancing the gain in the horizontal plane, we compared the E-field distributions of SCPRA and our proposed circular patch antenna with enhanced gain. As shown in Fig. 3.4a, the gain enhancement elements of our proposed antenna resonate at the 5.9 GHz and guide the wave along the horizontal plane, while in Fig. 3.4b, the beam of SCPRA is tilted away from the horizontal plane due to the effects of the ground. In this case, more energy can be retained in the horizontal plane of our proposed antenna, while the energy of the SCPRA will mostly be reflected upwards. In some degree, each group of gain enhancement elements can be regarded as one Yagi-Uda antenna. Therefore, the proposed antenna can achieve a higher omnidirectional gain in the horizontal plane, compared to that of SCPRA without gain enhancement elements.

Figure 3.5: Comparison of horizontal radiation patterns between antennas with 4 groups (dashed line) and 6 groups (solid line) of gain enhancement elements.

3.2.4 Number of Gain Enhancement Element's Groups

There are several considerations to determine the number of gain enhancement elements:

A. Due to the limited space and fixed size of gain enhancement elements, the number of groups cannot increase without limit. The number of groups should satisfy:

$$N \leq \frac{2\pi R_c}{L_e} \quad (3.5)$$

where R_c and L_e have been labelled in Fig. 3.2. Therefore, the value of N cannot be larger than 6. Besides, (3.5) shows that larger R_c means more

groups of gain enhancement elements. That's another reason that we bring in an annular ring.

B. The number of groups should ideally be as large as possible for increasing the gain in the horizontal plane. As shown in Fig. 3.5, the horizontal radiation pattern of the antenna with six groups of gain enhancement elements (the largest number of groups allowed in our design) is compared with that of the antenna with four groups. Apparently, the gains of the antenna with six groups (Max: 1.1 dB; Min: -0.5 dB) are higher than those of the antenna with four groups (Max: 0 dB; Min: -1.6 dB).

In our design, the number of gain enhancement elements is optimized to be six.

3.2.5 Number of Elements in one group

Since one group of the magnetic dipoles can be regarded as one Yagi-Uda antenna, it can be expected that the horizontal gain can be increased when more gain enhancement elements in one group are added. The simulated result verify our idea as shown in Fig. 3.6. When the number of magnetic dipoles increase to 0, the horizontal gain increases from -2 dBi to 1 dBi.

3.3 Equivalent Circuit of SCPRA

Our proposed antenna's matching property mainly depends on the parameters of SCPRA, as the gain enhancement elements have little effects due to the relatively weak coupling from the feed. Therefore, we only focus on

Figure 3.6: Horizontal gain versus different number of elements in one group.

Figure 3.7: Equivalent circuit of SCPRA.

Figure 3.8: Simplified equivalent circuit of SCPRA.

the influence of shorting pins and the annular ring on the antenna's input reflection coefficients.

An equivalent circuit model has been built which gives physical insight to the behavior of SCPRA and helps us match the antenna at desired frequency. Fig. 3.7 shows the proposed equivalent circuit model of SCPRA. Both the circular patch and the annular ring are represented by RLC circuit ((R_p, L_p, C_p) and (R_r, L_r, C_r) respectively) [110, 111]. C_g is the coupling capacitance between the circular patch and the shorted annular ring. The shorting pins are represented by an inductance.

However, since there is only one resonance over the operating frequency, the circuit model in Fig. 3.7 can be simplified as a parallel LC circuit as shown in Fig. 3.8. The shorted patch performs as an inductance L_{in} . The circuit of the annular ring and the gap between the ring and the patch are regarded as an entirety and HFSS simulation shows that it is capacitive with a capacitance C_{in} within the antenna operating frequency. The resonant frequency can be calculated by [111]:

$$f_0 = 1/(2\pi\sqrt{LC}) \quad (3.6)$$

Figure 3.9: Parametric study of reflection coefficients: (a) simulated reflection coefficients of SCPRA with varying gap; (b) simulated reflection coefficients of SCPRA with varying ring width.

Therefore, we can find the varying tendency of the resonant frequency, since based on (3.6), the resonant frequency is inversely proportional to the value of L_{in} and C_{in} .

Annular ring. The gap between the circular patch and the annular ring is a critical parameter in the antenna's impedance property. As illustrated in Fig. 3.9(a), with the increase of the gap, the resonant frequency shifts to a higher value. This is attributed to the fact that the larger gap reduces the value of the capacitance C_g , meaning a decrease in C_{in} , which surely increases the resonant frequency.

In addition, the width of the annular ring is another important factor

Figure 3.10: Simulated reflection coefficients of SCPRA with varying pins' radius.

affecting the reflection coefficient of SCPRA. Decreasing the width reduces C_r , the capacitive reactance of the ring, meaning the value of the capacitive loads of SCPRA decreases. Therefore, as shown in Fig. 3.9(b), the resonant frequency of SCPRA increases with the ring width decreasing.

Shorting pins. The six vias connecting the circular patch and the ground play a critical role in SCPRA's impedance matching. As illustrated in Fig. 3.10, the reflection coefficients of SCPRA are also sensitive to the radius of the shorting pins. The increased radius of shorting pins reduces its inductance L_s , which decreases the inductance of the equivalent circuits. Therefore, the resonant frequency increases when the radius of shorting pins increases.

Figure 3.11: The fabricated antennas: (a) omnidirectional antenna; (b) elliptically radiating antenna.

3.4 Experiment Results

In this section, two antennas, omnidirectional and elliptically radiating antennas with enhanced gain, are simulated and measured. Both simulated and measured results show that the proposed antennas have good matching property and about 3 dB enhanced gain in the horizontal plane within the DSRC band.

3.4.1 Omnidirectional Patch Antenna

The fabricated omnidirectional antenna is shown in Fig. 3.11(a). The return loss and the radiation patterns of the proposed antenna are shown below.

The reflection coefficients are shown in Fig. 3.12(a). The simulated bandwidth of the proposed antenna covers 5.86 GHz to 6.07 GHz, while the measured bandwidth starts from 5.80 GHz to 5.99 GHz. Both the simulated and

Figure 3.12: Simulated and measured results of the proposed omnidirectional antenna: (a) reflection coefficients; (b) radiation patterns in vertical plane; (c) radiation patterns in horizontal plane.

measured bandwidth cover the whole DSRC band (5.85 GHz to 5.925 GHz).

The radiation patterns of the proposed omnidirectional antenna at 5.9 GHz are shown in Fig. 3.12(b) and (c). The monopole-like radiation patterns are realized and the measured radiation patterns match well with the simulated results. The maximum gain in the horizontal plane is 0.97 dB. The simulated cross polarization is as low as a value below -40 dB and the measured cross polarization is only below -20 dB. The difference between the simulated and measured results may be mainly due to the fabrication error and the dielectric coefficient variations of PCB. Both the simulated and measured gains in the H-plane of the omnidirectional antenna with a finite ground are shown in the Table 3.2. The simulated results of the proposed antenna with an infinite ground is also shown for comparison. The gains in the H-plane of the omnidirectional antenna with the finite ground are all around 1 dB, and increase over 3 dB compared to the SCPRA within the DSRC band. Because of the diffraction effects of the finite ground plane [105], the gain of the antenna with the finite ground is 3 dB lower than that of the antenna with an infinite ground (4 dB).

It should be noted that in our design, we apply FR4 as the substrate for economical purpose. But the dielectric loss tangent of FR4 is high (0.02), which harms the radiation performance of the proposed antenna. As shown in Table 3.2, the horizontal gains of the proposed antenna fabricated on common FR4 is nearly 3 dB less than those values of the simulated antenna fabricated on lossless FR4. This means that we can employ the low-loss substrate to enhance the proposed antenna's horizontal gain.

Table 3.2: Simulated and measured gains of the omnidirectional antenna in the horizontal plane at 5.9 GHz

Frequency	5.9 GHz
Simulated gain of SCPRA (dB) (Fin)	-2.04 dB
Simulated gain of the omnidirectional antenna(dB) (Fin)	1.06 dB
Simulated gain of the omnidirectional antenna(dB) fabricated on lossless FR4 (Fin)	3.60 dB
Simulated gain of the omnidirectional antenna(dB) (Inf)	4.64 dB
Measured gain of the omnidirectional antenna(dB) (Fin)	0.97 dB

Table 3.3: Simulated and measured gains of the elliptically radiating antenna in the horizontal plane at 5.9 GHz

Frequency (GHz)	5.9
Simulated gain of the elliptically radiating antenna (Fin)	1.70 dB
Simulated gain of the elliptically radiating antenna (Inf)	5.50 dB
Measured gain of the elliptically radiating antenna (Fin)	1.10 dB

3.4.2 Elliptically Radiating Patch Antenna

In this subsection, the desirable radiation patterns for DSRC communication are discussed and an elliptically radiating antenna is designed for the application.

When the vehicle tries to communicate with the vehicles and the roadside units in front of it, the communication path is normally long. Therefore, the gain on the front and back sides of the DSRC antenna is expected to be large. On the contrast, when the telecommunication link is built between the vehicle and other units on the object's two flanks, the communication distance is much shorter; hence, it is normally not necessary to be large for the gain on the two flanks of the DSRC antenna. Therefore, the elliptically radiating antenna is more desirable for the DSRC applications.

Figure 3.13: Simulated and measured results of the proposed elliptically radiating circular patch antenna: (a) reflection coefficients; (b) radiation patterns in vertical plane; (c) radiation patterns in horizontal plane.

To satisfy the requirements of DSRC communications, an elliptically radiating patch antenna is proposed as shown in Fig. 3.11(b). The dimensions of the elliptically radiating patch antenna are the same as those of the omnidirectional antenna except the reduced number of gain enhancement elements in two opposite groups. To verify the performance of the elliptically radiating patch antenna, the fabricated antenna has been tested. The antenna has a bandwidth starting from 5.79 GHz to 5.97 GHz, which is slightly narrower than the simulated results as shown in Fig. 3.13(a).

The radiating characteristics at 5.9 GHz are simulated and measured, and the results are shown in Fig. 3.13. The quasi-elliptical radiation patterns in the horizontal plane are achieved and its gain at 5.9 GHz is 1.70 dB. The measured cross polarization of the proposed antenna is below -15 dB. As shown in Table 3.3, both the simulated and measured gains within the DSRC band are larger than those of the omnidirectional antenna, which is in line with our expectation. Therefore, the elliptically radiating antenna can realize relatively longer communication distance in front and back direction. Furthermore, the simulated gain in the horizontal plane of the elliptically radiating antenna with the infinite ground is also shown in Table 3.3. Its value is 4 dB larger than that of the antenna with the finite ground.

3.4.3 Flush-Mounted Proposed Antennas Integrated into a Metallic Cavity

To further lower the antenna's profile, the proposed antennas with omnidirectional and elliptically-radiating patterns are integrated into metallic cavities.

Figure 3.14: Return loss of the proposed antennas with and without PEC wall.

The integration in both the simulation and the measurement is simplified by surrounding the proposed antenna with PEC walls.

The measured return loss of the proposed antennas with and without PEC walls are compared in Fig. 3.14. It can be seen that the PEC walls have effects, which, however, is quite limited, on the return loss of the proposed antennas. The bandwidths of the proposed antennas with and without PEC walls can cover the DSRC bands. Besides, the measured gains of the antennas are also compared in Table. 3.4. The gains of the antenna with PEC walls is slightly smaller than those of the antennas without PEC wall but still larger than 0 dBi.

Table 3.4: Measured gains of the proposed antennas with and without PEC walls at 5.9 GHz

Frequency (GHz)	5.9
Gain of the omnidirectional antenna without PEC wall	0.97 dB
Gain of the omnidirectional antenna with PEC walls	0.47 dB
Gain of the elliptically radiating antenna without PEC walls	1.10 dB
Gain of the elliptically radiating antenna with PEC walls	0.70 dB

3.5 Summary

In this paper, we employed a curved half-width microstrip line with 2 flanks shorted and 1 edge open to the SCPRA to guide the TM wave along the horizontal surface and increase the gain in the horizontal plane. The gain enhancement element has been fully analyzed both in mathematical analysis and cavity mode. A circuit model of SCPRA has been built, which gives the guidelines to improve the proposed antenna's matching property. Two types of antennas, the omnidirectional and the elliptically radiating patch antennas have been designed, simulated and measured. The latter kind of antenna achieves the elliptical radiation patterns and higher gain up to 1.7 dB in the horizontal plane at 5.9 GHz, making it a good candidate for the DSRC communications. The omnidirectional and the elliptically radiating antennas can both be inserted into the metal cavity for the space-saving purpose with satisfied radiation performance.

Chapter 4

A Wideband Patch Antenna with I-Shaped Structure for Horizontal Gain Enhancement

4.1 Introduction

As introduced in Chapter 3, the CPA with high gain in the horizontal is desired for some wireless communications systems. However, the gain of the CPA in Chapter 3 is only 1 dB which is still relatively low. Besides, the resonant characteristic of the magnetic dipole limits its gain enhancement's bandwidth. Therefore, in this chapter, we aim to develop a wideband CPA with more remarkable gain enhancement in the horizontal plane.

Artificial structure/material is a promising way to increase the low-profile antenna's gain as the artificial structure can be printed on the PCB and control the propagation of electromagnetic waves in the material by changing

effective parameters like refractive index and dielectric coefficient [112, 113]. Some low/zero-index metamaterials (LIM/ZIM) behave as lenses to adjust the magnitude and the phase of the radiated field and are exploited to enhance antenna's radiation performance [114, 115]. However, the antennas loaded with such LIM/ZIM in [114, 115] are bulky. In [116], although a periodic endfire antenna loaded with two rows of low-index split-ring resonator structures (SRR) for gain enhancement is low-profile, the resonant property of SRR makes it difficult to further increase antenna's bandwidth. In contrast to resonant structures, the non-resonant structure is able to lead to a wider bandwidth [117]. In [118], an inhomogeneous anisotropic artificial structure with both broadband epsilon-near-zero (ENZ) and high refractive index characteristics employed to improve Vivaldi antenna's gain in broadband is presented. However, the directions of electric fields of these directional low-profile antennas are parallel to the surface of the 2-D artificial material, which is exactly the plane of the printed circuit board (PCB) [116–118]. On the contrary, the CPA achieving monopole-like radiation patterns is vertically polarized, which means electric fields are perpendicular to the plane of PCB. Therefore, to maintain the low profile of the CPA, the existing low-profile 2-D artificial structures for gain enhancement are not suitable for varying the propagation of electromagnetic waves of CPA.

In this chapter, a type of artificial structure “I-shaped” structure (ISS) with effective refractive index larger than that of the employed substrate over a wide band, is firstly proposed to enhance the radiation performance of the CPA in the horizontal plane. The simulation shows that the ISS can change the propagation direction of the electromagnetic wave which is

parallelly incident to the structure. Hence, the designed ISS behaves as a lens to concentrate the energy and increase the omni-directional gain of the CPA.

The ISS's electromagnetic characteristics are first obtained based on the retrieved S-parameters of the ISS in Section 4.2. Then the gain enhancement principles using ISS for the CPA are analyzed. In Section 4.3, a wideband CPA with one ring (CPA1R) is designed for verifying the performance of gain enhancement. The ISS is then annularly loaded around a CPA with one ring (CPA1R) without increasing CPA1R's thickness. The newly proposed antenna is named CPA with 1 ring and 3 layers of ISSes (CPA1R3I). To give an easier design guideline of the antenna, its parameters are also studied and analyzed in Section 4.4. Finally, both the simulated and the measured results demonstrate that up to 2.1 dB gain enhancement of the CPA can be achieved at a bandwidth from 5.14 GHz to 6.21 GHz in Section 4.5.

4.2 Design of an ISS Artificial Structure

The unit cell of the proposed artificial structure is a compressed "I-shaped" structure (ISS) as shown in Fig. 4.1. The ISS is designed on a 2-layer dielectric substrate with a thickness of 1.52 mm (0.5h as labelled in Fig. 4.1(b)) and a permittivity of 3. The diameter of the shorting pins (*viad*) is 2 mm. To obtain the effective parameters of ISS by using the *S*-parameter approach introduced in [119, 120], one setup is designed in high-frequency structure simulator (HFSS) as shown in Fig. 4.1(c). The dimensions of the box are $dx = 14$ mm, $dy = 6.5$ mm and $dz = 4$ mm. The values of dx and dy depend

(a) Top view of the ISS

(b) Cross section of the ISS.

(c) HFSS model of the ISS.

Figure 4.1: Geometry of the ISS.

on the space between adjacent units in ϕ -direction and x -direction respectively. The front-back faces and the bottom face are set as perfect magnetic conductor (PMC) and perfect electric conductor (PEC) boundary conditions separately. The incident direction is set along the y -axis while the electric polarization is along the z -axis.

The simulated S -parameters and the retrieved real parts of the refractive index versus frequency of the unit cell are presented in Fig. 4.2. It is observed

Figure 4.2: (a) Magnitude and (b) phase of the simulated S parameters of the unit cell ISS. (c) Equivalent retrieved real part of index (n_1).

(a) Top view.

(b) Cross section view.

Figure 4.3: Geometry of the CPA1R.

that its refractive indexes are higher than that of the substrate (1.43) within the frequency band from 5.14 GHz to 6.21 GHz (Fig. 4.2(c)). Besides, the ISS structure can generate a strong resonance at around 6.6 GHz. At this frequency, the return loss of the proposed ISS is very high while its transmitted coefficient is relatively small (Fig. 4.2(a)), which means that low energy can be transmitted through the unit cell and most are reflected back. And it should be noted that due to the asymmetric structure, $S_{11}(\text{ISS})$ and $S_{22}(\text{ISS})$ are not equal. As shown in Fig. 4.2(a) and 4.2(b), although the magnitudes of $S_{11}(\text{ISS})$ and $S_{22}(\text{ISS})$ have similar values, their phase values show the great difference, which is caused by different facing directions of the sides of the ISS to the incident wave.

(a) Top view.

(b) Cross section view.

Figure 4.4: Geometry of the proposed CPA1R3I.

4.3 Design of Circular Patch Antenna with the ISS for Gain Enhancement

4.3.1 Antenna Geometry

A wide-band circular patch antenna with one ring (CPA1R) shown in Fig. 4.3 is first proposed. For enhancing CPA1R's omni-directional gain in the horizontal plane, the CPA1R loaded by the ISS composes a new antenna, CPA with 1 ring and 3 layers of the ISS (CPA1R3I). The proposed antenna is designed on a double-layer substrate, the same as that of the unit cell ISS. A proximity feed is employed to excite the proposed antenna. A ring is added

Table 4.1: Parameters of the proposed omnidirectional antenna

Parameters	Value	Parameters	Value
r_1	1.9 mm	r_2	9.8 mm
r_3	19.6 mm	w	2.1 mm
G_0	0.4 mm	G_1	0.4 mm
s_0	2.8 mm	s_1	2.2 mm
s_2	2.2 mm	$pind$	1.8 mm

to the center-fed patch to introduce additional freedom for improving the matching property. The parameters of the CPA1R3I are shown in Table 4.1.

For reducing the size of the circular patch, six shorting pins are applied in CPA1R to excite TM_{01} mode with monopole-like radiation patterns maintained, which is firstly employed in [64]. The radius of the circular patch, r_3 , is only 12.5mm. On the other hand, the lowest mode for the conventional CPA to achieve monopole-like radiation patterns is TM_{02} mode [17]. The radius of a CPA at the TM_{02} mode can be calculated by (2.9). The calculated result is 20.1mm when the working frequency is 5.2 GHz. Compared to the conventional CPA working at the TM_{02} mode, CPA1R achieves over 38% size reduction. It should be noted that the mode nomenclature of circular patch in this paper is based on [17] and different from that in [11].

4.3.2 Gain Enhancement Principles

The gain enhancement principles can be explained by the refractive law in principles. As shown in Fig. 4.5(a), the electric fields in the ISS and air at the boundary are not parallel to each other. Obviously, the refraction happens when the electromagnetic wave transmits from the ISS to air. Based on the electric fields distribution in Fig. 4.5(a), the sketch map of the elec-

Figure 4.5: (a) Electric field distributions of the ISS (b) Electric field distribution of the substrate; (c) schematic diagram showing how the EM wave transmits from the ISS to air.

tromagnetic wave transmission in the vertical plane, such as xz -plane, is drawn in Fig. 4.5(c). n_0 and n_1 are the refractive indexes of air and the ISS respectively. According to Snell's law, we obtain:

$$\frac{n_1}{n_0} = \frac{\sin\theta_0}{\sin\theta_1} \quad (4.1)$$

When the angle of incidence (θ_1) is fixed, the larger refractive index value of the ISS (n_1) than that of the air will increase the angle of refraction (θ_0),

(a)

(b)

(c)

Figure 4.6: (a) Near electric field distribution of the CPA1R3I at 6.1 GHz; (b) Near electric field distribution of the CPA1R at 6.1 GHz; (c) Far field distribution of the CPA1R3I and the CPA at 6.1 GHz.

which indicates that the electromagnetic wave can be concentrated and propagate closer to the horizontal plane. As calculated in Sec. 4.2, the refractive index of the loaded ISS around the CPA1R is not only over 1 (the refractive index of air) but also higher than that of the substrate (1.43) on which the proposed antenna is printed. Therefore, compared to the substrate, the ISS can more effectively concentrate the energy and aid the electromagnetic wave propagating along the horizontal plane. For further verifying the effects of the ISS, the electric fields of the artificial structure and the substrate without loading are compared in Fig. 4.6(a) and 4.6(b). The electric fields on the boundary between the substrate and the air are obtained from the simulation tool. Obviously, the artificial structure can more markedly vary the propagation of the electromagnetic wave.

For demonstrating how the ISS influences the wave propagation, the near fields and the far fields of the CPA1R3I and CPA1R at 6.1 GHz are shown in Fig. 4.6. In the near-field region, the phenomenon that the beam width of the electric wave of the CPA1R3I is narrower than that of the CPA1R, is caused by the concentration of the ISS. In the far field region, the simulated results show that θ_{d1} (the beam direction of CPA1R3I) is much smaller than θ_{d2} (the beam direction of CPA1R), which illustrates that the loading structures can effectively guide the wave propagating closer to the horizontal plane. Hence, the gains in the horizontal plane of the CPA1R3I at these frequencies are higher than those of the CPA1R.

Besides, as shown in Fig. 4.7, with the increase of layers of the ISS, the horizontal gain can be increased from about 0 dBi to about 2 dBi. However, the effect of gain enhancement is weaker and weaker when adding more layers

Figure 4.7: The gains in the horizontal plane with varying number of the layers of the ISS at 6.1 GHz.

of the ISS. Because less energy can be transferred to the ISS far away from the circular patch. In conclusion, there is a trade-off between the antenna size and the antenna performance.

4.3.3 Wideband Principles

For further verifying the gain enhancement of the ISS on CPA at a wide band, the wide-band CPA1R is proposed, which the ISS can be employed on.

The wide bandwidth of the CPA1R is mainly achieved by utilizing the resonance for the TM_{01} mode and the TM_{02} mode which is firstly proposed in [64]. Besides, the ring and the proximity feed can also control the matching and reduce the antenna's Q value.

As introduced in Sec. 4.3.1, compared to the conventional CPA, the

Figure 4.8: The reflection coefficients and the impedance of CPA1R3I and CPA1R versus frequency.

CPA1R has one additional lower resonant frequency for the TM_{01} mode, where the CPA1R realizes monopole-like radiation patterns with the smaller size. As shown in Fig. 4.8, there are two resonant frequencies of CPA1R, f_1 (5.37 GHz) and f_2 (6.35 GHz), corresponding to the TM_{01} mode and the TM_{02} mode respectively. With the aid of the proximity feed and the ring, the bandwidth of CPA1R is improved to 27%. This is wider than the bandwidth of the CPA (18%) in [64], which is firstly enhanced by applying TM_{01} mode.

When the ISS is applied to the CPA1R, the coupling between the CPA1R and the ISS brings in a new resonant frequency f_r . However, as shown in Fig. 4.8(b), the ISS cause two peaks of resistance from 6.3 GHz to 6.5 GHz, which leads to the mismatch between the two resonant frequencies f_r and f_2 . It is shown in Fig. 4.8(a) that the bandwidth of CPA1R3I, which is not as wide as that of CPA1R, is still over 19% by covering the two reso-

Figure 4.9: The reflection coefficients and the impedance of CPA1R3I and CPA1R versus frequency.

nant frequency f_1 and f_r . This bandwidth is wide enough to prove the gain enhancement of the ISS over a wide bandwidth.

Besides the two resonances of the CPA1R3I, the proximity feed and the ring also contribute to widening the bandwidth. The proximity feed can reduce the antennas value Q . The working principle of the ring can be explained using an equivalent RLC circuit. When the ring is coupled to the circular patch, one coupling capacitance and one extra inductance are introduced. Therefore, more freedom is induced by the parameters like the width of the ring for widening the bandwidth. As shown in Fig. 4.9, with the aid of the proximity and the ring, the matching property of the CPA1R3I is effectively improved. The effects of dimensions of the proximity feed and the ring on the matching property will be discussed in Sec. 4.4.

Table 4.2: Simulated radiation performance of the CPA1R3I and the CPA1R in the horizontal plane at 5.9 GHz

Ground Radius (mm)	40	45	50	55	60
Ripples of CPA1R3I (dB)	0.5	0.8	0.5	0.5	0.6
Ripples of CPA1R (dB)	0.5	0.6	0.5	0.5	0.5
Gains of CPA1R3I (dB)	0.7	1.3	0.8	0.3	0.4
Gains of CPA1R (dB)	-0.4	0	-0.2	-0.7	-0.8
Gain increment (dB)	1.1	1.3	1	1	1.2

4.3.4 The Effects of the Ground on the Gain in the Horizontal Plane

Actually, the horizontal gain of the CPA is closely related to the diffraction on the ground edge and significantly affected by the size of the ground plane. Therefore, we investigated the effects of the ground on the gain enhancement in the horizontal plane.

As shown in Table. 4.2, the simulated ripples of the proposed CPA1R3I at 5.9 GHz are very close to those of the CPA1R. Therefore, we can compare only the peak gain to show the radiation improvement in the horizontal plane. Besides, it is also reasonable that in this paper, when we refer to the gain in the horizontal plane, it means the peak gain in the horizontal plane.

As expected, the gains in the horizontal plane of the CPA1R3I and the CPA1R vary with different ground size. Both the CPA1R3I and the CPA1R achieve the maximum gain when the ground radius is 45 mm. However, the gain increment is quite stable when the ground size increases from 40 mm to 60 mm. Hence, the ISS can effectively improve the gain in the horizontal plane of the CPA with negligible influences from varied ground size.

Figure 4.10: Simulated matching property of CPA1R3I with different r_1 : (a) S_{11} ; (b) input impedance.

4.4 Parametric Studies

In this section, parametric studies are conducted to give guidelines for improving antenna's matching property and increasing its bandwidth. The proximity feed, the slot, the ring and the six shorting pins on the circular patch have significant effects on the matching property of the CPA1R3I. The size of the ISS is fixed for maintaining the antenna's radiation performance. Therefore, r_1 , G_0 , r_2 , $pind$, r_3 , G_1 and w as labeled in Fig. 4.4(a) will be studied in this section to explore their influences on the reflection coefficient.

4.4.1 The Effects of the Proximity Feed on the Matching Property of CPA1R3I

The proximity feed brings in two degrees of freedom r_1 and G_0 to improve the matching property of CPA1R3I. As shown in Fig. 4.10(b), when r_1 increases

Figure 4.11: Simulated matching property of CPA1R3I with different G_0 : (a) S_{11} ; (b) input impedance.

from 1.6 mm to 2.2 mm, the inductive reactance of the proposed antenna increases while its resistance varies slightly within the operating bandwidth from 5.14 GHz to 6.21 GHz. Fig. 4.11(b) shows that with the increase of G_0 , both the resistance and the inductive reactance decrease. By adjusting r_1 and G_0 to 1.9 mm and 0.4 mm respectively, the proposed antenna realizes the widest bandwidth.

4.4.2 The Effects of 6 Shorting Pins on the Matching Property of CPA1R3I

The six shorting pins on the circular patch are very crucial components for improving the antenna's performance.

The position of the shorting pins, r_2 , is a key parameter to improve the antenna's matching property. As shown in Fig. 4.12, within the operating bandwidth from 5.14 GHz to 6.21 GHz, both the resistance and the reactance

Figure 4.12: Simulated matching property of CPA1R3I with different r_2 :(a) S_{11} ; (b) input impedance.

Figure 4.13: Simulated matching property of CPA1R3I with different $pind$:(a) S_{11} ; (b) input impedance.

decrease when r_2 reduces.

Fig. 4.13 shows how the size of shorting pins affects the antenna's matching characteristics. As shown in Fig. 4.13(a), with the diameter of shorting pins $pind$ decreasing, the resonant frequency for the TM_{01} mode shifts to a lower value, when the radius of the circular patch is fixed. Therefore, for CPA1R3I working at the TM_{01} , its size of center-fed circular patch can be reduced by decreasing the radius of shorting pins $pind$. However, it is shown in Fig. 4.13(b) that when the diameter of the shorting pins is small (e.g. 1.6 mm), the resistance is much smaller than 50Ω , which leads to the poor matching property in the band between the two resonant frequencies f_1 and f_r . Hence, CPA1R3I with too thin shorting pins cannot achieve a wide bandwidth. The optimal diameter of the shorting pins in our designed antenna is 1.8 mm.

4.4.3 The Effects of the Slot on the Matching Property of CPA1R3I

The slot also plays a critical role in varying antenna's reflection coefficients. Fig. 4.14 shows how the position of the slot affects the matching property of CPA1R3I. It is found that the optimized value of r_3 in our design is 12.5 mm. The resonant frequency for the TM_{01} mode moves to a lower value with r_3 increasing. However, at the higher frequency, the smaller value of r_3 causes higher inductive reactance and lower resistance.

As shown in Fig. 4.15(d), when G_1 decreases from 0.6 mm to 0.4 mm, the fluctuations of reactance are smaller and the resistance is closer to 50Ω

Figure 4.14: Simulated matching property of CPA1R3I with different r_3 : (a) S_{11} ; (b) input impedance.

Figure 4.15: Simulated matching property of CPA1R3I with different G_1 : (a) S_{11} ; (b) input impedance.

Figure 4.16: Simulated matching property of CPA1R3I with different w : (a) S_{11} ; (b) input impedance.

especially at a lower frequency within the operating band. Therefore, the antenna with $G_1 = 0.4$ mm achieves wider bandwidth. However, when G_1 further reduces to 0.2 mm, the resonance for the TM_{01} mode disappears, which leads to nearly 0 Ω resistance at a lower frequency. This means the CPA1R3I with $G_1 = 0.2$ mm has very poor matching property. Based on the analysis above, the width of the slot G_1 is chosen as 0.4 mm.

4.4.4 The Effects of the Ring on the Matching Property of CPA1R3I

It is shown in Fig. 4.16 how the width of the ring influences CPA1R3I's matching property. With the width increasing from 1.6 mm to 2.1 mm, at higher frequencies, the inductance of the proposed antenna decreases and the

Figure 4.17: Phototype of the fabricated CPA1R3I.

antenna's matching property is improved. When the width further raises to 2.6 mm due to the increase of the electrical size of the antenna, the operating band of CPA1R3I shifts to higher frequencies. In our design, the width of the ring is 2.1 mm.

4.5 Simulation and Measurement Results

As shown in Fig. 4.17, one sample of the proposed CPA1R3I is fabricated to validate its performance. Both the simulated and measured return loss are shown in Fig. 4.18. The simulated and measured -10 dB return loss bandwidths are 5.14 GHz-6.21 GHz and 5.09 GHz-6.36 GHz, respectively. The simulated fractional bandwidth is about 19% while the measured one is slightly wider, which is up to 22%.

The radiation patterns of the CPA1R3I at 5.3 GHz, 5.9 GHz and 6.1 GHz

Figure 4.18: Simulated and measured reflection coefficients of the CPA1R3I.

Table 4.3: Simulated and measured gains in the horizontal plane of CPA1R3I and CPA1R

Frequency (GHz)	5.3	5.5	5.7	5.9	6.1
Simulated gains of CPA1R (dBi)	-0.2	-0.2	-0.2	0	0
Simulated gains of CPA1R3I (dBi)	0.3	0.4	0.7	1.2	1.6
Measured gains of CPA1R (dBi)	-0.5	-0.5	0.1	-0.2	-0.4
Measured gains of CPA1R3I (dBi)	0.3	0.4	0.6	1.8	0.8
Measured enhanced gain (dB)	0.8	0.9	0.5	2	1.2

Figure 4.19: Radiation patterns of the CPA1R3I at 5.3 GHz (a: Vertical Plane, b: Horizontal Plane); 5.9 GHz (c: Vertical Plane, d: Horizontal Plane); 6.1 GHz (e: Vertical Plane, f: Horizontal Plane).

Table 4.4: Measured gains in the horizontal plane of CPA1R3I with and without PEC walls.

Frequency (GHz)	5.3	5.5	5.7	5.9	6.1
Measured gains of CPA1R3I without PEC walls (dBi)	0.3	0.4	0.6	1.8	0.8
Measured gains of CPA1R3I with PEC walls (dBi)	0.1	0.3	0.5	2.1	1.2

are shown in Fig. 4.19. Both the simulation and measurement results show that the proposed antenna can achieve stable monopole-like radiation patterns within the operating bandwidth. Due to the effects of the finite ground, the radiation patterns are tilted away from the horizontal plane.

For exhibiting the gain enhancement effects of ISS, the gains in the horizontal plane of CPA1R3I and CPA1R are compared in Table. 4.3. When no ISS is loaded, the simulated gains in the horizontal plane are not larger than 0.1 dB at 5.14 GHz-5.21 GHz frequency range. On the contrary, the CPA loaded with the ISS achieves -1 dB-1.8 dB measured horizontal gains, which are 0.5 dB-2 dB higher than those of the unloaded one within the operating bandwidth.

To further lower the profile of the proposed antenna, the CPA1R3I is integrated into a metal cavity, which means that the proposed antenna is surrounded by PEC walls. As shown in Table. 4.4, the gains in the horizontal plane of the CPA1R3I with and without PEC walls are close. Therefore, our proposed antenna can be inserted into the top of the vehicle roof.

4.6 Summary

In this paper, a low-profile and wide-band monopolar CPA1R3I is proposed. A kind of “I-shaped” structure (ISS) is applied to enhance the omni-directional gain in the horizontal plane of the proposed antenna. The artificial structure works as a lens to concentrate the wave and enhances antenna’s horizontal radiation performance. Besides, the ISS introduce a new coupling resonance, besides the resonance for TM_{01} mode, to increase CPA1R3I’s bandwidth. One CPA1R3I prototype is fabricated and measured. The results show that the proposed antenna achieves 19% relative bandwidth and 0.5 dB-2.1 dB enhanced horizontal gain. The high gain of the proposed antenna in the horizontal plane makes it a good candidate for the vehicle communication. Besides, it is found that the CPA1R3I can be inserted into the vehicle roof with wide bandwidth and high gain retained for space-saving purpose.

Chapter 5

A Very Wideband Low-Profile Monocone Antenna with Dielectric Loading

5.1 Introduction

The design of ultrawideband (UWB) antennas attracts much attention, as the UWB antennas can cover multi-bands and provide high data speed in modern communication systems [121]. Besides in some systems like V2X communications where antennas will be placed on a large ground, hence vertical polarization and omni-directional patterns are also desired. Under the circumstances, the monocone antenna may be preferred as it can satisfy all the aforementioned requirements.

As reviewed in Chapter 2, since the high thickness of the monocone antenna limits its application, many efforts devoted to decreasing the profile

of the moncone antenna. Top hat, shorting pin, low-cost plastic molding process and additional resonators et al. can effectively reduce the thickness of moncone antenna. However, the moncone antennas applying these techniques cannot achieve an over 140% bandwidth with a thickness below $0.06\lambda_0$ [18, 19]. Therefore, it is still a challenging task to design a low-profile structure with an ultra-wide bandwidth retained. Besides, the employed techniques destroy the principle propagation mode, TEM mode, of the conventional moncone antenna. Hence, the modes of low-profile moncone antennas are supposed to be investigated.

It is worth mentioning that besides the moncone antenna, there are many kinds of low-profile vertically polarized antennas with monopole-like radiation characteristics in [94, 104, 122–125]. Bent diamond-loops [104, 122, 125], bow-tie [123] and cone [94, 124] could successfully increase antenna's bandwidth without markedly increasing the height. But these antennas achieve either relatively narrow bandwidth or large size.

In this paper, the operating modes of the low-profile moncone antenna, the quasi- TM_{01} mode to the quasi- TM_{04} mode, at a wide bandwidth are explored in Section 5.2. These operating modes are similar to those of the circular patch antenna. The LC resonant circuit is employed to explain the excitation of the quasi- TM_{01} mode. Then an UWB low-profile moncone antenna with dielectric loading (MADL) which works at the quasi- TM_{01} mode to quasi- TM_{04} mode is proposed in section 5.3. The dielectric loading (DL) which is first applied to the improve moncone antenna's matching property, is studied. The proposed antenna with a thickness $0.057\lambda_{low}$ is designed, fabricated and measured. It is shown in section 5.4 that the proposed moncone

antenna has the monopole-like radiation patterns and the -10 dB bandwidth from 1.7 GHz to 12.6 GHz. Comparisons are made between our proposed antenna and other low-profile monopolar antennas in term of the bandwidth and the height.

5.2 Mode Analysis of Low-profile Monocone Antenna

A conventional low-profile monocone antenna is designed as shown in Fig. 5.1 for mode analysis at the beginning. Two techniques, top hat and shorting pins, which have been proposed in [91] and widely used, are applied to reduce the lowest working frequency to 1.9 GHz. The thickness of the monocone antenna is $0.063\lambda_{\text{low}}$ (10 mm). As introduced in Section. 5.1, because the side of the truncated cone, 14 mm, is far smaller than a quarter of the wavelength at the minimum operating frequency, the TEM mode doesn't exist at the low frequency. Actually, the modes of the low-profile antenna are mainly the quasi-TM mode, which is slightly different from those of the circular patch antenna. It should be noted that the mode nomenclature in this paper is based on [17] and different from that in [11].

The resistance and the reactance of the low-profile antenna versus frequency are shown in Fig. 5.2. There are four mode resonances at 1.89 GHz, 3.93 GHz, 6.92 GHz and 11.3 GHz which correspond to the quasi-TM₀₁ mode to the quasi-TM₀₄ mode respectively. Although the four mode resonances cannot lead to the continuous -10 dB bandwidth from 1.89 GHz to

Figure 5.1: (a) Top view and (b) cross section view of a conventional low-profile monocone antenna.

11.3 GHz. Here, we mainly aim to explore the characteristics of the modes of the low-profile monocone antenna without considering the matching property.

The electric fields at the mode resonances from 1.5 GHz to 13 GHz are extracted from the simulation tool. The electric fields when the local maximum amplitude propagates to the patch edge within one period are shown in Fig. 5.3. The cross-section views of the electric distributions are along

Figure 5.2: Impedance of the low-profile monocone antenna.

$\phi = 45^\circ$. The electric fields at the TM_{0N} mode have $N-1$ zeros along the ρ direction. Due to the effects of the shorting pins, the electric fields are not circularly symmetric, and there are ripples along the ϕ direction. Therefore, the modes of the low-profile monocone antenna are quasi-TM mode. The variation of the monocone antenna along the z -axis is not explored in this paper. Because the height of the monocone antenna is smaller than $0.5\lambda_{\text{low}}$ even at the quasi- TM_{04} mode. No higher mode except the fundamental mode along the z -axis will be excited.

At the lowest frequency, the antenna height is about $0.06\lambda_{\text{low}}$. The variation of the field along the z -axis can be ignored, and thus the resonant frequency of the low-profile monocone antenna at the lowest frequency can be approximatively expressed as that of the circular patch antenna which is [17]:

$$(f_r)_{mn} = \frac{\chi'_{mn} * c}{\pi a \sqrt{\epsilon_r}} \quad (5.1)$$

where a is the diameter of the patch as marked in Fig. 5.1. The radiation

Figure 5.3: E-field distributions of the low-profile monocone antenna: (a) top view, (b) cross-section view at 1.89 GHz; (c) top view, (d) cross-section view at 3.93 GHz; (e) top view, (f) cross-section view at 6.92 GHz; (g) top view, (h) cross-section view at 11.3 GHz.

Table 5.1: The parameters of the MADL

Parameter	$D_1^{(2)}$	$D_2^{(2)}$	$D_3^{(2)}$	$D_4^{(2)}$	$D_5^{(2)}$	$H^{(2)}$	$h^{(2)}$
Value (mm)	25	31	61	68	160	10	8
Parameter	$d_1^{(2)}$	$d_2^{(2)}$	$c_1^{(2)}$	$c_2^{(2)}$	$c_3^{(2)}$	$c_4^{(2)}$	
Value (mm)	2	18	4	12	28	32	

pattern of the circular patch antenna at the TM_{01} mode is monopole-like, but its corresponding frequency is zero, and thus this mode doesn't work. The excitation of the quasi- TM_{01} of the monocone antenna can be explained with the aid of the LC circuit model. The patch can be regarded as a series inductance (L_R). The patch with a grounded via is a mushroom structure bringing in a shunt conductor (C_R) and a shunt inductance (L_L). The shunt conductor (C_R) and inductance (L_L) can generate a new resonance. Therefore, with the aid of the shorting pins, a new mode, the quasi- TM_{01} , is generated at the low frequency and can be employed to reduce the minimum working frequency of the monocone antenna.

5.3 The Design of the Low-Profile Monocone Antenna with Dielectric Loading

5.3.1 The Geometry of the MADL

The proposed MADL with a height of 10 mm is shown in Fig. 5.4. The circular truncated cone (Fig. 5.4 (c) and (d)) is center-fed by a feeding with a diameter of 1.27 mm. Besides top hat and shorting pins with 2-mm diameter, slot and dielectric loading are employed to improve the bandwidth and

Figure 5.4: Antenna structure: (a) Top view of the antenna; (b) cross section view of the antenna; (c) top view of the cone; (d) cross section view of the cone; (e) top view of the quasi-ring; (f) cross section view of the quasi-ring.

Figure 5.5: Return loss of the monocone antennas with varying cut angle.

decrease the thickness of the monocone antenna. As shown in Fig. 5.4 (e) and (f), the loaded dielectric with a permittivity of 4 is a hollow truncated cone. Other parameters of the MADL are shown in Tab. 5.1.

5.3.2 Analysis of the Circular Truncated Cone

The cone plays an important role in the match of the MADL. The height of the cone ($h^{(2)}$) is limited by the thickness of the MADL. The lower diameter ($d_1^{(2)}$) should be slightly larger than that of the feeding (1.27 mm) for the smooth variance. As the cone is the transition from the feeding to the patch, the cut angle (α) cannot be either too large or too small for avoiding the abrupt change. The cut angle is varied from 30° to 60° in steps of 5° with $h^{(2)}$ and $d_1^{(2)}$ fixed. Several simulated results are selected for demonstration in Fig. 5.5. When the cut angle is larger than 45° , the matching property is good at the lower frequency but poor at the higher frequency. On the contrary,

Figure 5.6: Return loss of the monocone antennas with varying DL or without DL.

when the cut angle is smaller than 45° , the MADL is poorly matched at the lower frequency but achieves below -10 dB return loss at the frequency higher than 12 GHz. Therefore, the cut angle is set at 45° , where $d_2^{(2)}$ is correspondingly equal to 9 mm, to achieve the widest bandwidth.

5.3.3 Analysis of the Dielectric Loading

In this design, the slot and the DL are two effective matching methods applied to the low-profile monocone antenna. Since the influences of the slot on the return loss have been studied in [126], here only the loaded dielectric is explored for its effects.

The loaded dielectric performing as a matching layer between the cone and air is employed to improve the monocone antenna's matching property. For achieving this purpose, the permittivity of the loaded dielectric should be carefully chosen. The reflection coefficients of the MADL with different

Figure 5.7: Effects of the loaded dielectric with different DK and DF on the return loss of the MADL.

Figure 5.8: Impedance of the monocone antennas with or without loading.

kinds of dielectric are compared in Fig. 5.6. When the permittivity is 2 which is close to that of air, compared to the moncone antenna without the DL, the improvement of the matching property of the MADL is quite limited especially from 6.17 GHz to 9.3 GHz for achieving a successful match. However, when the permittivity is 6, although the minimum frequency decreases a little to 1.67 GHz, the matching property is destroyed due to the mismatch between air and the dielectric at the high frequency. Actually, the loaded dielectric with a permittivity of 4 also leads to the mismatch when the frequency is larger than 12.6 GHz compared to the moncone antenna without dielectric loading. However, considering achieving the widest bandwidth, a kind of material with a permittivity of 4 should be chosen to construct the matching layer.

Here, PPS-40GF with a dielectric constant of 4 and a dielectric loss of 0.004 is selected as the loaded material. With the increase of the frequency, both the dielectric constant and the dielectric loss may vary. Therefore, the effects of different dielectric constant (DK) and dielectric loss (DF) on the matching properties are explored in Fig. 5.7. It can be seen that although the dielectric constant increases from 3.5 to 4.5, the -10 dB bandwidths of the three antennas all cover 1.7 GHz to 12.4 GHz. The dielectric loss has a weak influence on the matching property as the reflection coefficients of the MADL with the dielectric loss of 0.004 and 0.01 are almost the same. Hence, the variance of the loaded dielectric needn't to be taken into account for the design of the MADL.

The impedances of the MADL and the moncone antenna without DL are compared to show the effects of the loaded dielectric. Fig. 5.6 shows

Figure 5.9: Prototype of the MADL.

that the monocone antenna without DL has poor matching property from 3.08 GHz to 3.91 GHz and from 6.17 GHz to 9.3 GHz while the MADL is well matched to the source at the bandwidth from 1.71 GHz to 12.5 GHz. It is shown in Fig. 5.8 that with the aid of the dielectric, the ripples of both the resistance and the reactance are markedly reduced. Especially at the band from 6.17 GHz to 9.3 GHz, the resistance of the MADL is reduced to around 50Ω . Therefore, it is easier for the MADL to be matched with a $50\text{-}\Omega$ coaxial line at an ultrawideband.

5.3.4 The effects of key parameters on the impedance matching of MADL

The impact of the key parameters on the lowest frequency (flow) is studied and displayed in Table. 5.2. It is worth mentioning that although some

Figure 5.10: Simulated and measured return loss of the MADL.

Table 5.2: The effects of key parameters on the impedance matching of MADL

Parameters	Varying tendency	f_{low}
Cone height ($h^{(2)}$)	$\uparrow$	$\downarrow$
Radius of top hat ($D_4^{(2)}$)	$\uparrow$	$\downarrow$
Radius of shorting pins (R)	$\uparrow$	$\uparrow$
Permittivity of loaded dielectric	$\uparrow$	$\downarrow$

techniques like increasing the permittivity of the loaded dielectric can reduce the lowest working frequency, the bandwidth is scarified. It is well known that one can trade the profile for wide bandwidth.

5.4 Experimental Results

A prototype of the MADL was fabricated as shown in Fig. 5.9. The top hat with the slot is printed on a 0.8 mm FR4 substrate for easy calibration.

Figure 5.11: Measured and simulated radiation patterns (H-plane¹: horizontal plane; V-plane²: vertical plane).

Figure 5.12: Simulated and measured realized gains of the MADL.

The simulated and the measured return loss of the MADL is shown in Fig. 5.10. The measured bandwidth is from 1.6 GHz to 11.7 GHz which slightly shifts to lower frequency compared to the simulated one. Both the simulated and the measured results show that the MADL achieves over 150% fractional bandwidth.

Fig. 5.11 shows the measured and simulated radiation patterns at 2 GHz, 4 GHz, 7 GHz and 11 GHz. Good agreements are obtained between the simulated and measured results. It can be seen that the monopole-like radiation patterns are achieved at the wide operating band. The measured and the simulated realized gains are shown in Fig. 5.12. The gains at the operating band are all larger than 3 dBi. Besides, the simulated of efficiency of the MADL is over 88% as shown in Fig. 5.13.

The comparisons between our proposed MADL and the current state-

Figure 5.13: Simulated efficiency of the MADL.

of-the-art low-profile UWB antennas with monopole-like radiation characteristics are made in Table. 5.3. It can be seen that our proposed MADL achieves the widest bandwidth than others except the antenna in [104]. However, in [104], the bandwidth is defined as the range of frequency over which the VSWR is below 2.7:1 which is much higher than that in our definition (1.93:1). In addition, only the heights of the antennas in [104] and [122] are lower than that of the MADL, but at the expense of the matching properties. In conclusion, the MADL has satisfactory performance in terms of the matching property and the thickness.

5.5 Summary

A low-profile monocone antenna with dielectric loading is proposed. It is found that the modes of the low-profile monocone antenna are similar to the circular patch antenna. The operating modes at an ultrawide bandwidth are the quasi-TM₀₁ mode to the quasi-TM₀₄ mode which are unlike the TEM

Table 5.3: Comparison between the proposed MADL and the current state-of-art in UWB antennas with monopole-like radiation characteristics

Work (Years)	Dimensions ($H \times L \times W$) ($\lambda_{\min}^3$)	Relative Bandwidth	VSWR
[122] (2013)	$0.033 \times 0.240 \times 0.240$	4:1	3:1
[123] (2014)	$0.085 \times 0.189 \times 0.189$	5.5:1	2.2:1
[124] (2014)	$0.087 \times 0.204 \times 0.204$	3.9:1	1.93:1
[104] (2015)	$0.046 \times 0.260 \times 0.260$	8.5:1	2.7:1
[19] (2016)	$0.068 \times 0.208 \times 0.208$	2.9:1	1.93:1
[125] (2017)	$0.090 \times 0.145 \times 0.145$	4.1:1	2:1
[94] (2018)	$0.064 \times 0.267 \times 0.267$	3.1:1	1.93:1
This work	$0.057 \times 0.363 \times 0.363$	7.3:1	1.93:1

mode of the conventional moncone antenna. The excitation of the quasi- TM_{01} mode is explained by the LC resonant circuit. Besides top hat, shorting pins and slot, the DL is first applied as a matching layer to improve the matching properties by adjusting the impedance of the antenna.

Both simulation and measurement show that the proposed antenna with a thickness of about $0.055\lambda_{\text{low}}$ has an over 150% fractional bandwidth. The monopole-like radiation characteristics are achieved over the wide bandwidth. Comparisons between our proposed moncone antenna and the current state-of-art in vertically-polarized UWB omni-directional antennas show that the MADL has satisfactory performance in terms of the profile and the bandwidth.

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

This thesis devotes to designing low-profile vertically-polarized omnidirectional antennas for V2X communications. Based on the literature review, it is found that the circular patch antenna and the monocone antenna are two promising candidates for their vertical polarization and omnidirectional radiation characteristics. However, the two types of antennas have their respective drawbacks that may limit their applications on V2X communications. The circular patch antenna with monopole-like radiation patterns has low gain in the horizontal plane and relatively narrow bandwidth. Besides, it is still a challenging task to design a low-profile monocone antenna with an ultra-wide bandwidth retained.

In Chapter 3, it is found that a half-width microstrip line working as a

magnetic dipole can successfully increase the gain in the horizontal plane of the circular patch antenna. The resonator can guide the electromagnetic wave along the antenna surface and hence increase the gain in the horizontal plane. The gain enhancement element has been thoroughly analysed both in mathematical analysis and cavity mode. A circuit model is built to give the physical insight of the antenna. Two types of antenna with omnidirectional and elliptically radiating patterns are designed, simulated, fabricated and measured. Results show that the enhanced gain in the horizontal plane can be increased to 1.7 dBi in the DSRC band. The two proposed antennas have the ability of being inserted into the roof with an enhanced gain in the DSRC band.

Since the bandwidth of the proposed antennas in Chapter 3 is relatively narrow, a wideband circular patch antenna with the I-shaped structure for horizontal gain enhancement is proposed. It is found that the proposed I-shaped structure works as a lens to concentrate the wave and enhances antenna's horizontal radiation performance due to its large refractive index in a wide band. In addition, the ISS introduces a new coupling resonance, besides the resonance for the TM_{01} mode, to increase CPA1R3I's bandwidth. Both the simulated and the measured results show that the proposed antenna can achieve up to 0.5 dBi to 2 dBi gain increase in the horizontal plane over 1 GHz bandwidth. Besides, inserting the antenna into the roof has slight effects on the radiation and the matching properties of the antenna.

Next, a low-profile monocone antenna with dielectric loading is proposed. The propagation modes of the low-profile monocone antenna are studied and it is found that these modes are quasi- TM_{0N} modes which are similar to

those of the circular patch antenna. Besides, LC resonant circuit is applied to explain the excitation of the first mode. The proposed technique, dielectric loading, can effectively improve the matching properties of the monocone antenna by providing additional freedom of adjusting the impedance. Both the simulated and the measured results show that the proposed antenna can achieve an over 153% fractional bandwidth with a thickness of only $0.05\lambda_{\text{low}}$.

In conclusion, three kinds of low-profile vertically-polarized omnidirectional antennas are proposed. These antennas have either enhanced omnidirectional gain or ultrawideband which are beneficial for V2X communication.

6.2 Recommendation for Future Research

Although this thesis has presented several promising low-profile vertically-polarized omnidirectional antennas for V2X communication, there are still some opportunities for us to improve low-profile omnidirectional antennas' performance further. The recommendations are listed below.

To increase the horizontal gains of circular patch antennas, the surface wave can be excited and guided by some special structures, for example, ridged ground, along with the surface of the antenna. The tapered ground can be applied to insert the proposed antenna into the roof of vehicles for low-profile purpose. The newly proposed antenna may be promising to achieve the higher gain in the horizontal plane with a very low profile.

Inspired by the low-profile monocone antenna, we can apply a proximity feeding, a cone, to excite the patch antenna. The proximity feeding may be able to greatly improve the matching properties of CPA. Shorting

pins should also be applied to excite the TM_{01} modes. Then a wideband vertically-polarized CPA with monopole-like radiation patterns is firstly proposed. Based on wideband principles, a wideband circularly-polarized circular patch antenna may be achieved. Some additional carefully-designed structure can introduce ϕ component and the centre-fed circular patch can generate θ component. By tuning the dimensions of the additional structures, the magnitudes of the two orthogonal component can be equal and therefore good circularly polarized property can be obtained.

Although the loaded dielectric can effectively increase the matching properties of the low-profile monocone antenna, due to the mismatching between the dielectric and the air, the proposed antenna has poor matching properties at high frequencies. Double matching layers with diminishing permittivity may be applied for smooth transformation from the dielectric loading to the air. In this case, a wider bandwidth may be achieved and cover higher order modes.

Author's Publications

Journal List

- J1. A.K. Liu, Y.L. Lu and L. Huang, "Low-profile patch antennas with enhanced omnidirectional gain in the horizontal plane for DSRC applications," *IET Microwaves, Antennas & Propag.*, vol. 12, Iss. 2, pp. 246-253, Feb. 2018.
- J2. A.K. Liu, L. Huang and Y.L. Lu, "A wideband circular patch antenna with I-shaped structure for horizontal omnidirectional gain enhancement," to appear in *IET Microwaves, Antennas & Propag.*, 2019.
- J3. A.K. Liu and Y.L. Lu, "A very-wide low-profile monocone antenna with dielectric loading," submitted to *IEEE Transaction on Antennas and Propagation*, (minor).

Conference List

- C1. A.K. Liu and Y.L. Lu, "A 6.8GHz low-profile H-plane ridged SIW horn antenna," *2015 IEEE 4th Asia-Pacific Conference on Antennas and*

Propagation (APCAP), Bali Island, Indonesia, 30 Jun.-3 Jul. 2015.

C2. A.K. Liu and Y.L. Lu, "A low profile stepped wideband monocone antenna," *2017 IEEE International Symposium on Antennas and Propagation*, San Diego, California, 9-17 Jul. 2017.

C3. A.K. Liu and Y.L. Lu, "An ultra-wideband monocone antenna with ceramic loading," Accepted by *2018 IEEE International Symposium on Antennas and Propagation*, Boston, Massachusetts, 8-13 Jul. 2018.

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