

# Circularly Polarized Antennas

Steven (Shichang) Gao  
Qi Luo  
Fuguo Zhu



WILEY

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**Steven (Shichang) Gao, Qi Luo and Fuguo Zhu**

*University of Kent, UK*

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# Preface

Due to the features of circular polarization, circularly polarized antennas are very useful for various wireless systems such as satellite communications, global navigation satellite systems, mobile communications, wireless sensors, radio frequency identification, wireless power transmission, wireless local area networks, wireless personal area networks, worldwide interoperability for microwave access and direct broadcasting service television reception systems. Recent decades have seen a lot of research and development activities in CP antennas from industries and institutes worldwide. There are very few books focusing on CP antenna design and principles. Thus, there is a need for a comprehensive book which presents basic principles and up-to-date developments of CP antennas and arrays.

The purpose of this book is to present a comprehensive overview of various types of CP antennas and arrays, including the basic principles, design methods, size reduction techniques, broadband techniques, multi-band techniques, array design techniques and recent developments, as well as their applications and case studies. This book can be used as a reference book for graduate students, academics, researchers and antenna engineers. It includes up-to-date developments from industry and academic research experts worldwide, and numerous CP antenna design examples are illustrated.

The book is organized into six chapters. Chapter 1 introduces the reader to basic concepts of antenna parameters and the principles of typical CP antennas. This will help readers who are not familiar with the antenna basics. Many different types of basic CP antennas including CP patch antennas, crossed dipoles, helix antennas, quadrifilar helix antennas, printed quadrifilar helix antennas, spiral antennas, slot antennas, dielectric resonator antennas, patch arrays and slot arrays are explained.

Chapter 2 discusses various techniques for designing small-size CP antennas. Size reduction of CP antennas is important for practical applications as mobile devices are getting smaller and having more functions. Design techniques of different types of small CP antennas are presented, and many design examples are illustrated.

Chapter 3 introduces various types of broadband CP antennas and their design techniques. Broadband CP antennas are important for applications such as global navigation satellite system receivers, high-speed satellite data downlink, high-speed wireless communication, and so on. Design techniques of different types of broadband CP antennas are reviewed, discussed and compared.

Chapter 4 presents the design techniques of multi-band CP antennas which are important for many applications such as satellite communications, global navigation satellite systems, radio frequency identification, wireless power transmission, wireless local area networks,

and so on. Design techniques of different types of multi-band CP antennas are discussed, and their advantages and disadvantages are explained.

Chapter 5 discusses different types of CP arrays and their design techniques. These include CP patch arrays, the CP dielectric resonator array antenna, CP waveguide slot arrays, CP reflectarrays, and so on. Broadband CP active integrated arrays, beam-switching and electronica beam-steering CP arrays are also discussed.

Chapter 6 presents case studies that illustrate how to design and implement CP antennas in practical scenarios. Five case studies, including a dual-band CP array for GPS remote sensing applications, a small printed quadrifilar helix antenna for satellite communications mobile terminals at the S-band, a broadband CP antenna for a radio frequency identification reader, a CP reflectarray for Ka-band mobile satellite communications and a wideband logarithmic spiral antenna for wideband RF measurement applications, are presented. The design guidelines for each of these CP antennas are given and each design step is explained. Thus the reader is provided with a comprehensive and logical path from CP antenna basics to advanced designs of different types of CP antennas and arrays.

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# Abbreviations and Acronyms

|       |                                         |
|-------|-----------------------------------------|
| ABC   | Absorbing Boundary Condition            |
| AIA   | Active Integrated Antenna               |
| AMC   | Artificial Magnetic Conductor           |
| AR    | Axial Ratio                             |
| ARBW  | Axial Ratio Bandwidth                   |
| CDL   | Coupled Double Loop                     |
| CLL   | Capacitively-Loaded Loop                |
| CMOS  | Complementary Metal-Oxide-Semiconductor |
| CP    | Circularly Polarized                    |
| CPW   | Coplanar Waveguide                      |
| DBS   | Direct Broadcasting Service             |
| DFILA | Double-Folded Inverted-L Antenna        |
| DRA   | Dielectric Resonator Antenna            |
| EBG   | Electromagnetic Bandgap                 |
| ESA   | Electrically Small Antenna              |
| FDTD  | Finite Difference Time-Domain           |
| FEM   | Finite Element Method                   |
| GCPW  | Grounded Coplanar Waveguide             |
| GNSS  | Global Navigation Satellite System      |
| GPS   | Global Positioning System               |
| GSM   | Global System for Mobile Communications |
| HIS   | High Impedance Surface                  |
| HPBW  | Half-Power Beamwidth                    |
| ILA   | Inverted-L Antenna                      |
| IoT   | Internet of Things                      |
| LHCP  | Left-Hand Circular Polarization         |
| LNA   | Low Noise Amplifier                     |
| LTCC  | Low Temperature Co-fired Ceramic        |
| LTE   | Long Term Evolution                     |
| MEMS  | Microelectro-Mechanical System          |
| MIC   | Microwave Integrated Circuit            |
| MIMO  | Multiple Input and Multiple Output      |
| MMIC  | Microwave Monolithic Integrated Circuit |
| MOM   | Method of Moments                       |

|        |                                                                  |
|--------|------------------------------------------------------------------|
| MPQHA  | Meander Printed Quadrifilar Helix Antenna                        |
| MTM    | Meta-material                                                    |
| MVPQHA | Meandered Variable Pitch Angle Printed Quadrifilar Helix Antenna |
| PAE    | Power Added Efficiency                                           |
| PCB    | Printed Circuit Board                                            |
| PEC    | Perfect Electric Conductor                                       |
| PIFA   | Planar Inverted-F Antenna                                        |
| PMC    | Perfect Magnetic Conductor                                       |
| PPW    | Parallel Plate Waveguide                                         |
| PQHA   | Printed Quadrifilar Helix Antenna                                |
| QHA    | Quadrifilar Helix Antenna                                        |
| RFID   | Radio Frequency Identification                                   |
| RHCP   | Right-Hand Circular Polarization                                 |
| RLBW   | Return Loss Bandwidth                                            |
| RLSA   | Radial Line Slot Antenna                                         |
| SIW    | Substrate Integrated Waveguide                                   |
| SLH    | Stub Loaded Helix                                                |
| SMT    | Surface Mount Technology                                         |
| SPDT   | Single-Pole Double-Throw                                         |
| SPMT   | Single-Pole Multi-Throw                                          |
| TTC    | Telemetry, Tracking and Control                                  |
| UHF    | Ultra-High Frequency                                             |
| UMTS   | Universal Mobile Telecommunications System                       |
| UWB    | Ultra-Wideband                                                   |
| VSWR   | Voltage Standing Wave Ratio                                      |
| WHR    | Width-to-Height Ratio                                            |
| WiMAX  | Worldwide Interoperability for Microwave Access                  |
| WLAN   | Wireless Local Area Network                                      |
| WPAN   | Wireless Personal Area Network                                   |

# 1

## Introduction to Circularly Polarized Antennas

### 1.1 Introduction

Circularly polarized (CP) antennas are a type of antenna with circular polarization. Due to the features of circular polarization, CP antennas have several important advantages compared to antennas using linear polarizations, and are becoming a key technology for various wireless systems including satellite communications, mobile communications, global navigation satellite systems (GNSS), wireless sensors, radio frequency identification (RFID), wireless power transmission, wireless local area networks (WLAN), wireless personal area networks (WPAN), Worldwide Interoperability for Microwave Access (WiMAX) and Direct Broadcasting Service (DBS) television reception systems. Lots of progress in research and development has been made during recent years.

The CP antenna is very effective in combating multi-path interferences or fading [1,2]. The reflected radio signal from the ground or other objects will result in a reversal of polarization, that is, right-hand circular polarization (RHCP) reflections show left-hand circular polarization (LHCP). A RHCP antenna will have a rejection of a reflected signal which is LHCP, thus reducing the multi-path interferences from the reflected signals.

The second advantage is that CP antenna is able to reduce the ‘Faraday rotation’ effect due to the ionosphere [3,4]. The Faraday rotation effect causes a significant signal loss (about 3 dB or more) if linearly polarized signals are employed. The CP antenna is immune to this problem, thus the CP antenna is widely used for space telemetry applications of satellites, space probes and ballistic missiles to transmit or receive signals that have undergone Faraday rotation by travelling through the ionosphere.

Another advantage of using CP antennas is that no strict orientation between transmitting and receiving antennas is required. This is different from linearly polarized antennas which are subject to polarization mismatch losses if arbitrary polarization misalignment occurs between transmitting and receiving antennas. This is useful for mobile satellite communications where it is difficult to maintain a constant antenna orientation. With CP, the strength of the received signals is fairly constant regardless of the antenna orientation. These advantages make CP antennas very attractive for many wireless systems.

This chapter serves as a basis for the chapters that follow. It will introduce some basic parameters of antennas. Different types of basic CP antennas such as CP microstrip patch antenna, helix, quadrifilar helix antenna (QHA), printed quadrifilar helix antenna (PQHA), spiral antenna, CP dielectric resonator antenna (DRA), CP slot antennas, CP horns and CP arrays will be described and basic designs illustrated. Typical requirements and challenges in CP antenna designs will be discussed at the end.

## 1.2 Antenna Parameters

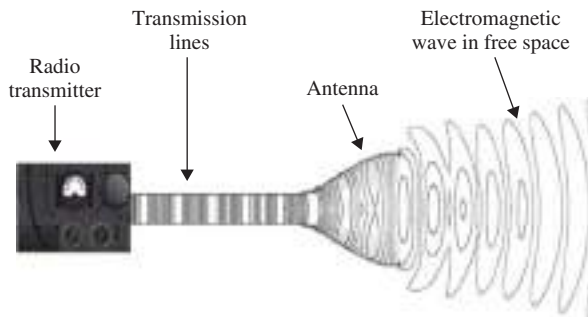
An antenna is a device which can receive or/and transmit radio signals. As a receiving device, it can collect the radio signals from free space and convert them from electromagnetic waves (in the free space) into guided waves in transmission lines; as a transmitting device, it can transmit radio signals to free space by converting the guided waves in transmission lines into the electromagnetic waves in the free space. In some cases, an antenna can serve both functions of receive and transmit.

Figure 1.1 depicts the basic operation of a transmit antenna. As shown, the information (voice, image or data) is processed in a radio transmitter and then the output signal from the transmitter propagates along the transmission lines before finally being radiated by the antenna. The antenna converts the guided-wave signals in the transmission lines into electromagnetic waves in the free space. The operation of a receive antenna follows a reverse process, that is, collecting the radio signals by converting the electromagnetic waves in free space into guided-wave signals in the transmission lines, which are then fed into radio receivers.

### 1.2.1 Input Impedance

The input impedance  $Z_{in}$  is defined as the impedance presented by an antenna at its feed point, or the ratio of the voltage to current at the feed point [5]. The input impedance is usually a complex number which is also frequency dependent. It can be expressed as

$$Z_{in} = R_{in} + jX_{in} \quad (1.1)$$



**Figure 1.1** Basic operations of a transmit antenna

The real part of the impedance,  $R_{in}$ , includes the radiation resistance  $R_r$  of the antenna and the loss resistance  $R_L$ .  $R_r$  relates to the power radiated by the antenna, and  $R_L$  relates to the power dissipated in the antenna due to losses in dielectric materials, antenna conductor losses, and so on.

### 1.2.2 Reflection Coefficient, Return Loss and Voltage Standing Wave Ratio

The antenna input impedance needs to be matched with the characteristic impedance of the transmission line connected to the feed point of the antenna. Usually a 50  $\Omega$  cable is used to feed the antenna. Thus the antenna input impedance needs to be equal to 50  $\Omega$ , otherwise there will be an impedance mismatch at the antenna feed point. In the case of impedance mismatch, there will be signal reflections, that is, some of signals fed to the antenna will be reflected back to the signal sources.

The reflection coefficient  $\Gamma$  denotes the ratio of the reflected wave voltage to the incident wave voltage [5]. The reflection coefficient at the feed point of the antenna can be related to the antenna input impedance by the following equation:

$$\Gamma = \frac{Z_{in} - Z_0}{Z_{in} + Z_0} \quad (1.2)$$

Here,  $Z_{in}$  and  $Z_0$  denote the input impedance of the antenna, and the characteristic impedance of the transmission line connected to the antenna feed point, respectively. As shown in equation (1.2), the reflection coefficient is zero if  $Z_{in}$  is equal to  $Z_0$ .

Return loss (in dB) is defined as:

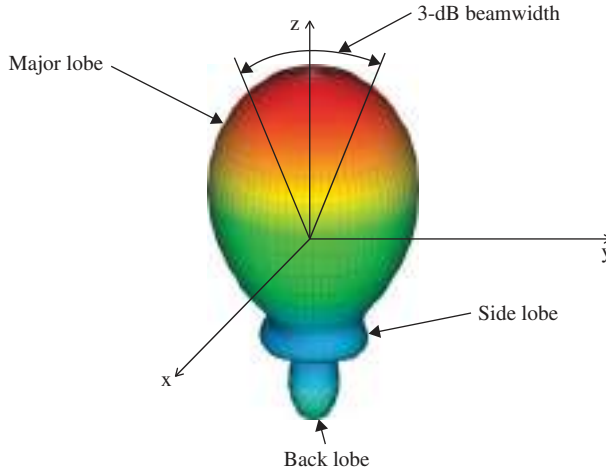
$$RL = -20 \log |\Gamma|$$

For a well-designed antenna, the required return loss should usually be at least 10 dB, though some antennas on small mobile terminals can only achieve about 6 dB. Voltage Standing Wave Ratio (VSWR) is the ratio of the maximum voltage  $V_{max}$  to the minimum voltage  $V_{min}$  on the transmission line. It is defined as:

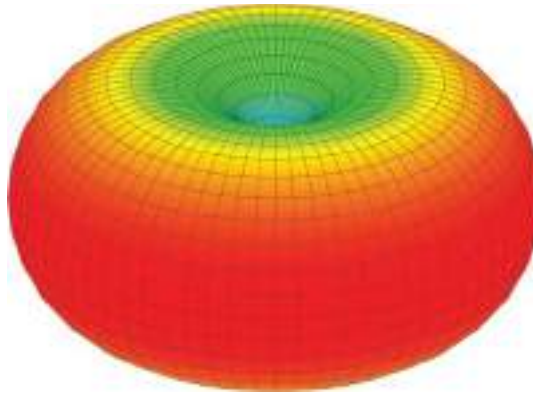
$$VSWR = \frac{|V_{max}|}{|V_{min}|} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (1.3)$$

### 1.2.3 Radiation Patterns

The radiation pattern of the antenna illustrates the distribution of radiated power in the space [6–9]. It can be plotted in a spherical coordinate system as the radiated power versus the elevation angle ( $\theta$ ) or the azimuth angle ( $\varphi$ ). Figure 1.2 shows a radiation pattern plotted as the radiated power versus the elevation angle ( $\theta$ ). As shown, the radiation pattern has a few lobes. The main lobe is the lobe containing the majority of radiated power. The lobe radiating towards the backward direction is the back lobe. Usually there will also be a few other small lobes called the side lobes. The 3-dB beamwidth indicated in the figure refers to the angular range between two points where the radiated power is half the maximum radiated power. Figure 1.2 shows the radiation pattern in the elevation plane. There is also



**Figure 1.2** A directional radiation pattern in the elevation plane



**Figure 1.3** Omni-directional pattern

the radiation pattern in the azimuth plane, which can be plotted as the radiated power versus the azimuth angle ( $\phi$ ). The antenna pattern can be isotropic, directional or omni-directional. An isotropic pattern is uniform in all directions, which does not exist in reality. The pattern in Figure 1.2 is directional. As shown in Figure 1.2, the majority of radiated power is focused at one direction and the maximum radiation is along the  $z$  axis. An omni-directional pattern is donut-shaped, as shown in Figure 1.3.

#### 1.2.4 Directivity, Gain and Efficiency

A practical antenna usually radiates in certain directions. The directivity,  $D(\theta, \phi)$ , is defined as the radiated power per unit solid angle compared to what would be received by an isotropic

radiator [6–9]. It can be calculated by

$$D(\theta, \varphi) = \frac{r^2 \cdot \frac{1}{2} \text{Re}[E \times H^*]}{P_{\text{rad}}/4\pi} = \frac{2\pi r^2 \cdot \text{Re}[E \times H^*]}{P_{\text{rad}}} \quad (1.4)$$

where  $E$ ,  $H$ ,  $r$  and  $P_{\text{rad}}$  denote the peak value of electric field, the peak value of magnetic field, the distance between the source and test point, and the radiated power from the antenna, respectively. It is also assumed that the test point is in the far field region of the antenna, which means the distance  $r > \frac{2D^2}{\lambda}$  [6–9]. Here  $D$  is the maximum dimension of the antenna and  $\lambda$  is the wavelength.

The gain of the antenna is similar as the directivity though it includes the efficiency  $\eta$  of the antenna, since some power will be lost in the antenna.

$$G(\theta, \varphi) = \eta \cdot D(\theta, \varphi) \quad (1.5)$$

Both directivity and gain are normally expressed in dB. It is common practice to write the antenna gain in dBi, which means that it is defined relative to an isotropic radiator.

### 1.2.5 Linear Polarization, Circular Polarization and Axial Ratio

Polarization of an antenna is related to the orientations of electric fields radiated by the antenna. Assuming a half-wavelength dipole is vertically oriented above the Earth, it will produce radiated fields in the far field and the radiated electric fields will be dominated by  $E_\theta(\theta, \varphi)$ . In this case, the polarization of the dipole is called *vertical polarization*. On the other hand, if a half-wavelength dipole is horizontally oriented above the Earth, the radiated electric fields of the antenna will be dominated by  $E_\varphi(\theta, \varphi)$  in the far field. The polarization of the antenna is then called *horizontal polarization*. Both vertical and horizontal polarizations are *linear polarizations*. Linearly polarized antennas are commonly used in terrestrial wireless communications.

To produce circular polarization, two orthogonal components of electric fields in the far field region are required [6–10]. The electrical field radiated by an antenna can be written as

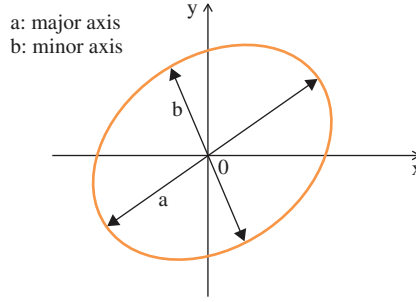
$$\vec{E}(\theta, \varphi) = \vec{\theta} E_\theta(\theta, \varphi) e^{j\phi_1} + \vec{\varphi} E_\varphi(\theta, \varphi) e^{j\phi_2} \quad (1.6)$$

Here  $E_\theta(\theta, \varphi)$  and  $E_\varphi(\theta, \varphi)$  denote the magnitudes of electric field components in the far field of the antenna.  $\phi_1$  and  $\phi_2$  denote the phase shift of each field component.

Circular polarization can be achieved only if the total electric field has two orthogonal components which have the same magnitudes and a  $90^\circ$  phase difference between the two components. That is

$$\begin{aligned} E_\theta(\theta, \varphi) &= E_\varphi(\theta, \varphi) \\ \phi_2 - \phi_1 &= \pm \frac{\pi}{2} \end{aligned} \quad (1.7)$$

For a circularly polarized wave, the electric field vector at a given point in space traced as a function of time is a circle. The sense of rotation can be determined by observing the direction of the field's temporal rotation as the wave is viewed along the direction of wave



**Figure 1.4** Polarization ellipse traced at a certain position as a function of time

propagation: if the field rotation is clockwise, the wave is RHCP; if the field rotation is anti-clockwise, the wave is LHCP.

In reality, it is impossible to achieve a perfect circular polarization, thus the curve traced at a given position as a function of time is usually an ellipse, as shown in Figure 1.4. Lines  $a$  and  $b$  denote the major axis and the minor axis of polarization ellipse, respectively. The ratio of the major axis to the minor axis of the ellipse is termed as the axial ratio (AR) [6–10].

$$AR = \frac{a}{b} \quad (1.8)$$

AR is a key parameter for measuring the circular polarization. Usually AR is required to be below 3 dB for a CP antenna.

### 1.2.6 Bandwidth and Resonant Frequency

Usually an antenna is designed to operate within a specified frequency range. The bandwidth of an antenna is usually determined by the frequency range within which the key parameter of the antenna satisfies a certain requirement, for example, minimum return loss of 10 dB. At the resonant frequency of an antenna, the antenna input impedance is purely resistive. Often the resonant frequency is chosen as the centre of the frequency bandwidth of an antenna. The bandwidth of an antenna can be calculated by using the upper and lower edges of the achieved frequency range:

$$BW = \frac{f_2 - f_1}{f_o} \times 100\% \quad (1.9)$$

where  $f_1$  is the lower edge of the achieved frequency range,

$f_2$  is the upper edge of the achieved frequency range, and  
 $f_o$  is the centre frequency of the range.

Note that this definition is for antennas with a bandwidth below 100%. For antenna bandwidths over 100%, the bandwidth can be calculated using the ratio between the upper and lower edge of frequencies. For a linearly polarized antenna, the input impedance is usually the most sensitive parameter compared to other antenna parameters such as radiation patterns, gain and polarization. Thus the bandwidth of a linearly polarized antenna is often

referred to as the ‘impedance bandwidth’, but it can also be done with other parameters such as radiation patterns, gain and polarization.

When evaluating the bandwidth of CP antennas, one must check both the impedance bandwidth and the bandwidth of AR, that is, the frequency range within which the AR is below 3 dB. A good impedance matching does not necessarily lead to a good gain or a low AR value. The impedance bandwidth of an antenna can be broadened using suitable impedance matching networks, while the AR bandwidth can be broadened by using a broadband phase shifter network [5].

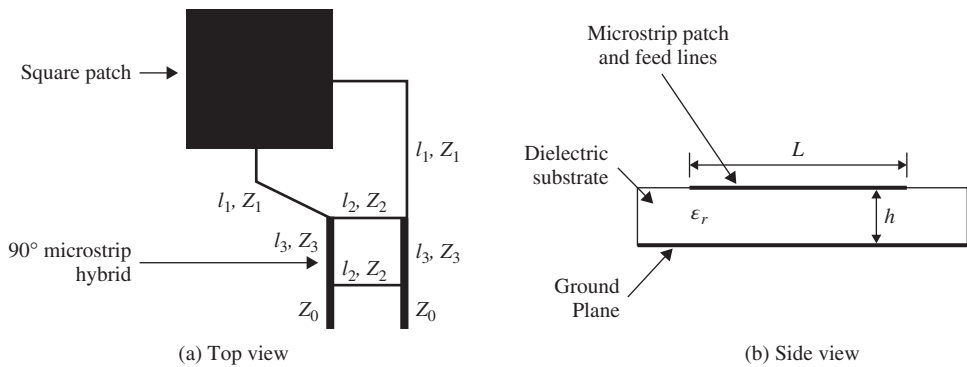
### 1.3 Basic CP Antenna Types

#### 1.3.1 CP Microstrip Patch Antennas

Microstrip patch antenna is one very popular type of antennas, due to its advantages of low profile, easy fabrication, low cost and conformability to curved surfaces. Circular polarization in patch antennas can be achieved using a multi-feed technique or a single-feed technique [6–9,11–13].

Figure 1.5 shows a simple CP microstrip patch antenna using a dual-feed technique. Both the top and side views of the antenna are shown in Figure 1.5. To produce circular polarization, a square microstrip patch is fed by two orthogonal microstrip feedlines as shown in Figure 1.5(a). Two microstrip feed lines excite the patch antenna in  $TM_{01}$  and  $TM_{10}$  modes so that it radiates both a horizontally polarized wave and a vertically polarized wave simultaneously [6–9,11–13]. A microstrip hybrid is employed in Figure 1.5(a) to produce a  $90^\circ$  phase difference between two orthogonally polarized waves. Figure 1.5(b) shows the side view of the antenna. The metallic patch is etched on the top of a dielectric substrate having thickness  $h$  and relative permittivity  $\epsilon_r$ . The dielectric substrate is backed by a metallic ground plane. To design the antenna resonant at a frequency  $f_o$ , the length  $L$  of the patch can be approximately calculated by using the following equation:

$$L \approx \frac{c}{2f_o \sqrt{\frac{\epsilon_r + 1}{2}}} \quad (1.10)$$



**Figure 1.5** A microstrip line-feed patch antenna with a  $90^\circ$  hybrid

where  $c$  is the velocity of light. To achieve accurate antenna designs, full-wave electromagnetic simulators can be employed to do simulations and optimizations of the dimensions of antenna. The results from equation (1.10) can be used as an initial value for the antenna optimization.

As shown in Figure 1.5(a), two microstrip feed lines connect the square patch and two ports of the microstrip 90° hybrid. The microstrip line serves as an impedance transformer between the input of the antenna and the input ports of the hybrid. The length ( $l_1$ ) and characteristic impedance ( $Z_1$ ) of the microstrip feed lines can be calculated by

$$\begin{aligned} l_1 &= \frac{\lambda_g}{4} \\ Z_1 &= \sqrt{Z_{in}Z_0} \end{aligned} \quad (1.11)$$

where  $\lambda_g$  is the guided wavelength of the microstrip line,

$Z_{in}$  is the input impedance of the patch antenna, and

$Z_0$  is the characteristic impedance of the microstrip line at the input of hybrid circuit.

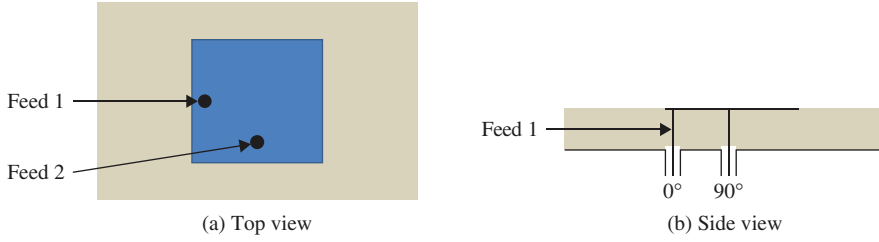
The microstrip 90° hybrid circuit consists of four sections of microstrip lines. The following equations can be employed to determine the length and width of each section of microstrip lines:

$$\begin{aligned} l_2 &= l_3 = \frac{\lambda_g}{4} \\ Z_2 &= Z_0 \\ Z_3 &= \frac{Z_0}{\sqrt{2}} \end{aligned} \quad (1.12)$$

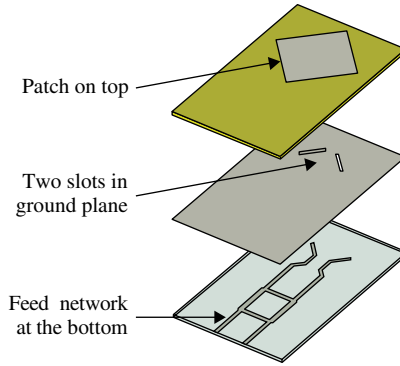
where  $l_2$  and  $l_3$  denote the length of microstrip lines as indicated in Figure 1.5(a),  $Z_2$  and  $Z_3$  denote the characteristic impedance of microstrip line sections indicated in Figure 1.5(a).

The characteristic impedance of the microstrip line at the input of hybrid circuit,  $Z_0$ , is usually chosen to be 50  $\Omega$ . The microstrip hybrid circuit is easy to be fabricated and has been widely used in CP antennas. One drawback of microstrip hybrid circuit is the large size. Many techniques have been developed to reduce the size of microstrip hybrid, for example, by using ‘ $\Pi$  network’ with stub loading or lumped-element loading of transmission lines [5,14,15]. The hybrid circuit can also be implemented by using lumped elements or left-handed transmission lines [14,15].

The CP patch antenna can also use other feeding structures, such as probe feeds, slot-coupled feeds, electromagnetically-coupled feed and coplanar waveguide (CPW) feeds. Figures 1.6 and 1.7 show a CP patch antenna using two probe feeds, and a CP patch with slot-coupled feeds, respectively. Both the top view and side view of the antenna are shown in Figure 1.6. As shown, circular polarization is obtained by feeding a square patch with a 90° phase difference between two feed probes placed symmetrically on the two orthogonal edges of the patch. In this case, an external phase shifter is required for producing the 90° phase difference between two feeds. The antenna in Figure 1.7 employs a microstrip hybrid circuit for producing the 90° phase difference between two feeds. The microstrip hybrid is put at the bottom of the antenna, and coupled to the square patch on the top via



**Figure 1.6** A probe-feed patch antenna with an external 90° phase shifter



**Figure 1.7** Slot-coupled CP patch antenna (dual feed with a hybrid coupler)

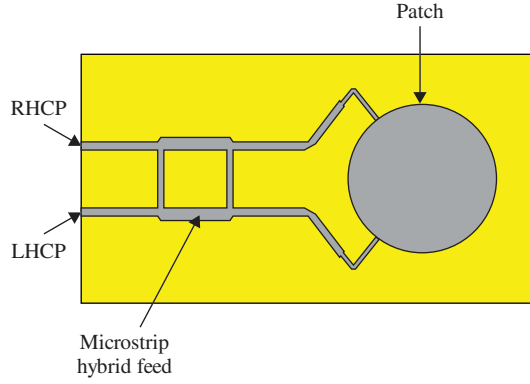
two orthogonal slots cut in the ground plane. The slot-coupled CP antenna has been widely used in wireless systems during recent years, as it has many advantages compared to CP antennas using other feed structures: (1) It allows the patch antenna and the feed circuits to employ different dielectric substrate so that both the patch antenna and the feed circuits can achieve optimized performance; (2) It is easy to integrate active circuits with the feed network; (3) The parasitic radiation of feed network is reduced due to the isolation of the ground plane. The slot-coupled CP patch antenna is a popular choice for radiating elements in phased arrays for satellite communications.

For the dual-feed antenna, the radiating patch can use different shapes, such as square, circular, annular ring, and so on. Figure 1.8 shows a circular patch integrated with a microstrip hybrid circuit. As shown, the antenna can achieve either RHCP or LHCP, depending on the choice of input port.

To design the CP antenna in Figure 1.8, an approximate solution to the radius of patch is given by [6,7]:

$$a \approx \frac{F}{\sqrt{1 + \frac{200h}{\pi\epsilon_r F} \left[ \ln \left( \frac{\beta F}{200h} \right) + 1.7726 \right]}}$$

$$F = \frac{8.791 \times 10^9}{f_o \sqrt{\epsilon_r}} \quad (1.13)$$



**Figure 1.8** Circular patch integrated with a hybrid coupler

where  $a$  is the radius of circular patch,

$f_o$  is the resonant frequency of the CP antenna in  $TM_{11}$  mode  
 $\epsilon_r$  is the relative permittivity of the dielectric substrate, and  
 $h$  is the thickness of the substrate.

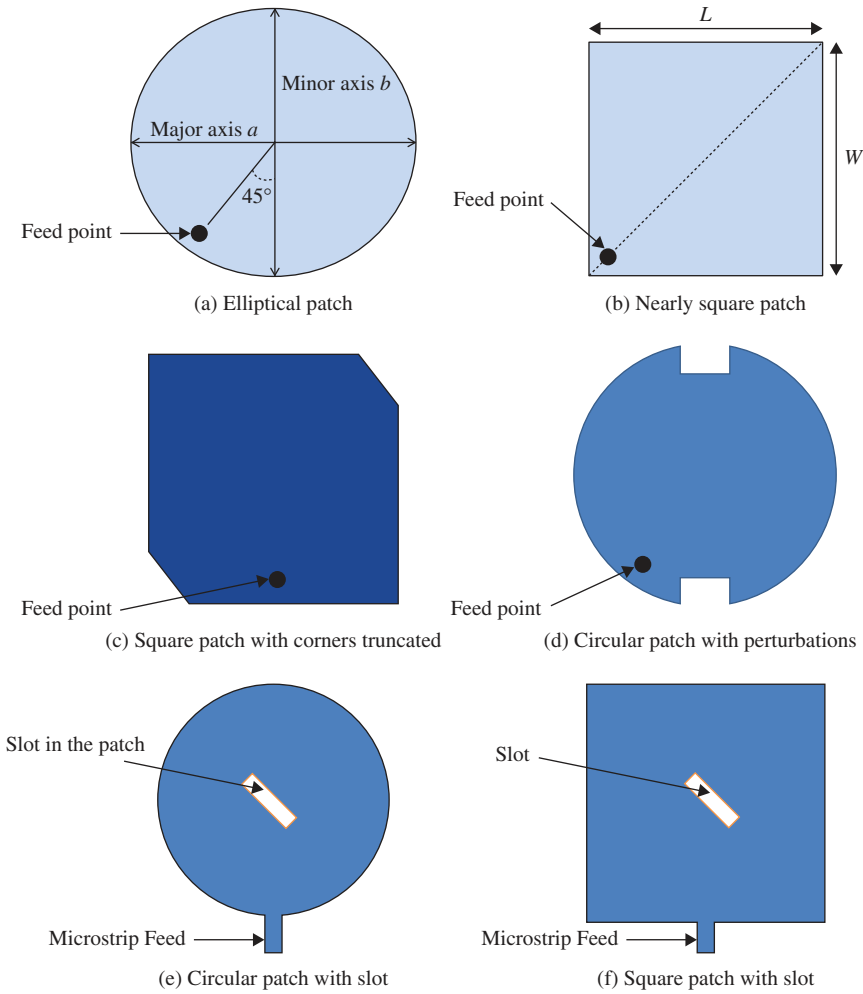
Besides the use of two feeds, it is also possible to excite circular polarization using more than two feeds. For example, circular polarization in the patch antenna can be excited using four feeds orthogonally located at four edges of a square patch and with an appropriate phase difference. Such a multi-feed technique can suppress higher order modes, and provide high polarization purity and a broad bandwidth at the expense of large size and complexity of feed network [13].

To simplify the feed network of CP patch antennas, a single-feed technique has been developed. Figure 1.9 shows six different configurations of single-feed CP microstrip patch antennas [12,13]. Such a technique requires the perturbation of the patch shape. For example, Figure 1.9(a) shows an elliptical patch fed along a line  $45^\circ$  from its major axis. The elliptical patch can be regarded as a circular patch with perturbations. The perturbation of the patch shape is used to excite two orthogonal modes with a  $\pm 90^\circ$  phase difference. To design the CP antenna using a single-feed elliptical patch, the ratio between the major axis and minor axis is given by:

$$\frac{a}{b} = 1 + \frac{1.0887}{Q} \quad (1.14)$$

The value of antenna quality factor  $Q$  can be computed using the cavity model [6,7,12,13]. Alternatively, one can either measure the  $Q$  of the antenna experimentally, or use results from a full-wave electromagnetic analysis to estimate  $Q$ .

$$Q = \frac{f_o}{\Delta f} \frac{VSWR - 1}{\sqrt{VSWR}} \quad (1.15)$$



**Figure 1.9** Single-feed CP patch antenna

where  $f_o$  is the resonant frequency of the antenna, and

$\Delta f$  is the bandwidth of the antenna.

Another option for single-feed CP patch antenna design is to use the configuration shown in Figure 1.9(b), which is a nearly square patch fed at a point along the diagonal line of the patch. The length and width of the patch are  $L$  and  $W$ , respectively. The condition of circular polarization is satisfied when

$$L = W \left( 1 + \frac{1}{Q} \right) \quad (1.16)$$

The value of antenna quality factor  $Q$  can be computed using equation (1.15). The resonant frequencies  $f_1$  and  $f_2$  associated with the length  $L$  and width  $W$  of a rectangular microstrip patch are

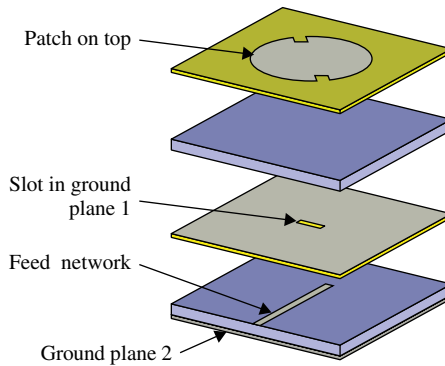
$$f_1 = \frac{f_o}{\sqrt{1 + \frac{1}{Q}}}$$

$$f_2 = f_o \sqrt{1 + \frac{1}{Q}} \quad (1.17)$$

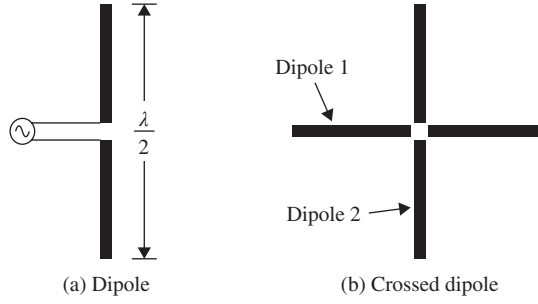
where  $f_o$  is the centre frequency of the bandwidth.

Figure 1.9(c) and (d) show a square patch with two corners truncated, and a circular patch with two notches, respectively. Both antennas employ a single probe feed and the feed position is indicated in the figure. The perturbation can also take the form of a narrow slot cut in the centre of the patch, as shown in Figure 1.9(e) and (f). Other shapes of single-feed patch antennas, using pentagons, annular elliptic patches, and so on have also been reported [12,13]. The single-feed technique does not require a complicated feed network as in the multi-feed CP patch antennas, and is compact in size. The main drawback of this technique is the narrowband AR performance, typically 1–2%.

As in the case of multi-feed CP patch antenna, the single-feed CP antenna can also use different feed structures. Figure 1.10 shows a slot-coupled single-feed patch antenna. As shown, a circular patch with two notches is put on the top of dielectric substrate. The feed network is coupled to the patch via a single slot cut in the ground plane. Due to the use of a single slot, the feed network is much simpler compared to the dual-feed CP antenna in Figure 1.7. The slot can produce bi-directional radiation. To reduce the backward radiation from the slot, another ground plane can be added at the bottom, as shown in Figure 1.10. In this case, the feed network uses the stripline instead of microstrip lines, and usually vias are added between two ground planes for avoiding the excitation of parallel-plate modes in the stripline circuits [13].



**Figure 1.10** Slot-coupled single-feed patch antenna



**Figure 1.11** A dipole and a crossed dipole antenna

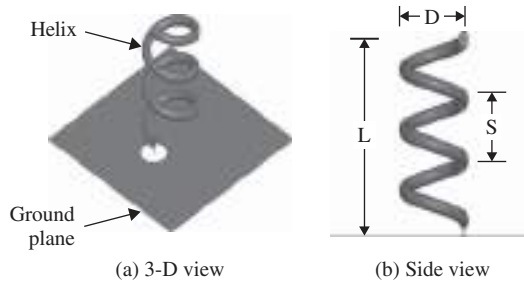
### 1.3.2 CP Wire Antennas

Crossed dipoles using wires have been employed to obtain circular polarization for many years. Figure 1.11 shows a dipole antenna and a crossed dipole antenna. The half-wavelength dipole in Figure 11(a) is vertically polarized and has an omni-directional pattern. For a crossed dipole in Figure 1.11(b), two dipoles are mounted perpendicular to each other and fed with a  $90^\circ$  phase difference between them. The  $90^\circ$  phase network can employ one quarter-wavelength of coaxial cable.

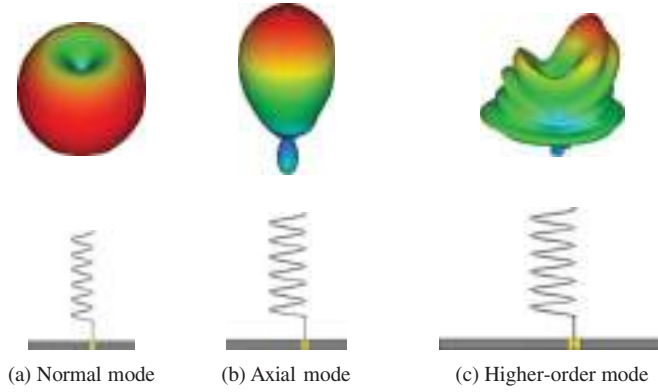
### 1.3.3 Helix Antennas

The helix antenna is one of the most promising antenna types for CP applications [6,9,11]. Figure 1.12 shows a helix antenna. It is basically a conducting wire wound in the form of a screw thread. Key design parameters of a helix antenna include the diameter of one turn ( $D$ ), circumference of one turn ( $C$ ), vertical separation between turns ( $S$ ), the number of turns ( $N$ ) and pitch angle ( $\alpha$ ), which controls how far the helix antenna grows in the axial-direction per turn.

The helix antenna has gained wide application because of its simple structure, wide operation bandwidth and circular polarization. The helix antenna can operate at three different modes [6,9,11]:



**Figure 1.12** Helix antenna



**Figure 1.13** Axial, normal and higher order radiation modes of helix antenna [10,16]

- *Normal mode*, which occurs when the diameter of the helix is relatively small compared to the wavelength. The radiation pattern is omni-directional as shown in Figure 1.13(a).
- *Axial mode*, which occurs when the circumference of the helix is of the order of one wavelength. The maximum radiation is along the axis of the helix, as shown in Figure 1.13(b).
- *Higher-order radiation mode*, which occurs when the dimensions of the helix exceed those required for the axial mode. The major lobe is split-up as shown in Figure 1.13(c).

The axial mode is the one of major interests for CP applications. The normal-mode helix is useful for terminals in terrestrial cellular systems but not CP applications. The following gives the design equations for key design parameters of the helix so that it can achieve optimum performance in the axial mode [6,7,9]:

$$\begin{aligned}
 \frac{3}{4}\lambda < C < \frac{4}{3}\lambda \\
 S &\approx \frac{1}{4}\lambda \\
 12^\circ &\leq \alpha \leq 14^\circ
 \end{aligned}
 \tag{1.18}$$

where  $\lambda$  is the free-space wavelength.

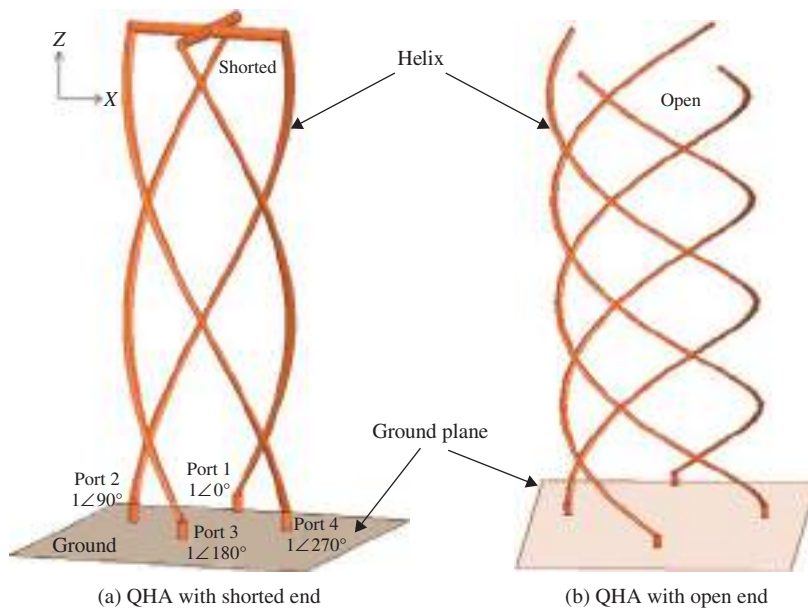
To understand how circular polarization is produced by a helix antenna, the helix can be approximated by  $N$  small loops and  $N$  short dipoles connected together in series. Two orthogonally polarized fields are produced by the loops and the dipoles, respectively. Here the planes of the loops are parallel to each other and perpendicular to the axes of the vertical dipoles. A  $90^\circ$  phase difference between these two orthogonal fields is obtained if the vertical separation between turns  $S$  is chosen to be one quarter-wavelength as given in equation (1.18). The helix antenna has inherently broadband properties, possessing desirable radiation patterns, impedance and polarization characteristics over a relatively wide frequency range. The axial mode pattern exists for a nearly 2 to 1 frequency range because the natural adjustment of phase velocity results in the fields from the different turns adding in phase in the axial direction. The input impedance remains almost constant because of the large

attenuation to the reflected waves from the open end, and the antenna polarization remains circular because the in-phase field condition is a condition for circular polarization too [9].

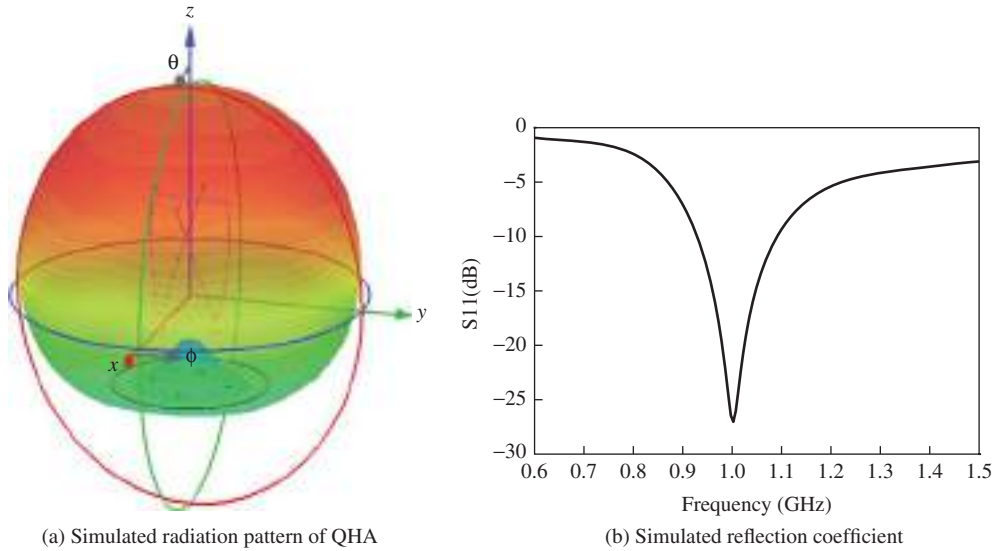
### 1.3.4 Quadrifilar Helix Antennas and Printed Quadrifilar Helix Antennas

The QHA is one of the most commonly used antennas for satellite communications and Global Positioning Systems (GPS) applications [9,16–19]. The QHA can produce a cardioid-shaped radiation pattern with excellent circular polarization over a wide angular range. Such a pattern is suitable for GPS as it allows more satellites to be visible. Figure 1.14 shows two typical configurations of QHA antennas. Basically, a QHA consists of four identical helices interleaved with each other. Four identical helices are fed with a separate phase quadrature network which provides  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  phases to each of four helices, respectively. The helices can be shorted circuited or open-circuited at the end, as shown in Figure 1.14(a) and (b), respectively. QHA has important features of versatility and flexibility, due to the many degrees of freedom; such as the total length of each helical element, the number of turns, radius of helix, pitch angle, axial length and so on. In addition to the cardioid-shaped pattern, optimum choice of antenna parameters in QHA can lead to other types of radiation patterns, gain and bandwidth performance so that it can be used in a variety of applications for satellite communications and terrestrial systems. The QHA is a resonant radiating structure when the total length of each helical element,  $L_{total}$ , is equal to an integer number of quarter wavelengths [9].

$$L_{total} = N \frac{\lambda}{4} \quad (1.19)$$



**Figure 1.14** Configurations of QHA



**Figure 1.15** Simulated results of an end-shortened QHA

Here  $N$  is an integer number. When  $N$  is an even number, the helices should be shorted together, as shown in Figure 1.14(a); while when  $N$  is an odd number, the helices should be open-circuited, as shown in Figure 1.14(b).

Figure 1.15 shows the simulated results of an end-shortened QHA which has helices of  $\lambda/2$  long and  $1/2$  turn. Figure 1.15(a) shows the simulated radiation pattern of a QHA. As shown, QHA can achieve a wide-beam circularly polarized pattern, suitable for GPS signal reception. The simulated reflection coefficient results are shown in Figure 1.15(b). It is noted that, compared to the monofilar helical antenna, the QHA has a narrower bandwidth.

The QHA mentioned previously requires a separate phase quadrature network for providing  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  phases to each of four helices. Such a feed network increases the size and complexity of antennas. To alleviate this problem, self-phased QHA has been developed. In the self-phased approach, two bi-filars with different lengths are employed and the difference in lengths of bi-filars leads to a  $90^\circ$  phase difference between them. As the self-phased QHA does not require a separate phase quadrature network, it reduces the size and complexity of QHA. QHA is a popular choice for handheld mobile terminals in mobile satellite communications due to its small size and hemi-spherical radiation patterns.

The fabrication of QHA will require accurate bending and shaping of wires which is not always easy. Figure 1.16 shows a PQHA, which is basically a printed version of QHA antennas. Four printed helices are wound around a cylinder as shown in Figure 1.16. The four printed helices are to be fed in phase quadrature in order to produce the desired hemi-spherical pattern. Compared to QHA antennas, PQHA is easier to fabricate in mass quantities and is lower cost, as the antennas can be fabricated using standard printed circuit board (PCB) technology. In addition, PQHA allows more flexibility in antenna designs. For example, it is easy to produce a PQHA with meandered-line helices patterns so as to reduce the antenna size [20]. Such a pattern will be, however, difficult to implement in wires



**Figure 1.16** PQHA antenna

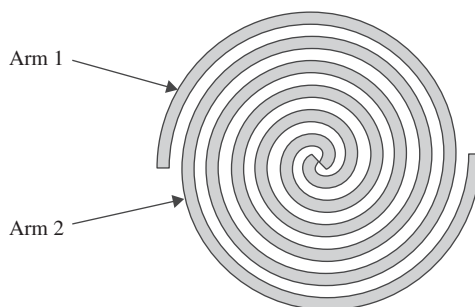
for QHA. The use of printed technology enables accurate fabrication of PQHA with less fabrication tolerance issues as in the case of QHA. Due to the use of PCB technology, it is also possible to integrate PQHA with microwave diodes or devices.

### 1.3.5 Spiral Antennas

Spiral antennas belong to the class of frequency independent antennas which operate over a wide range of frequencies. Radiation pattern, polarization and impedance of such antennas remain unchanged over a wide bandwidth [21]. Figure 1.17 shows one type of spiral antenna called an *Archimedean spiral antenna*. It includes two conductive arms, extending from the centre outwards. The antenna has a planar structure. Each arm of the Archimedean spiral is defined by the equation:

$$r = a\phi \quad (1.20)$$

Equation (1.20) states that the radius  $r$  of the antenna increases linearly with the angle  $\phi$ . The parameter  $a$  is a constant which controls the rate at which the spiral flares out. Arm 2



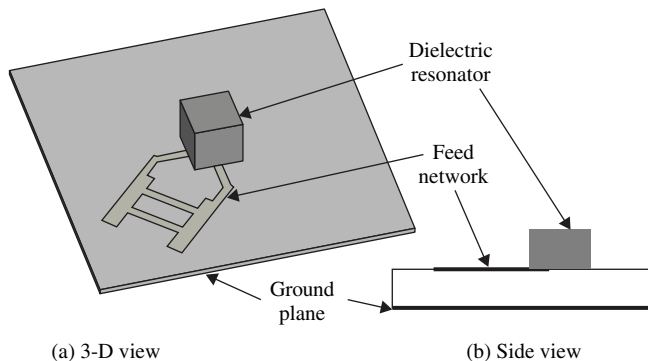
**Figure 1.17** Spiral antenna

of the spiral is the same as the arm 1, but rotated at  $180^\circ$ . The direction of rotation of the spiral defines the direction of antenna polarization. Additional arms may be included to form a multi-spiral antenna. Usually the spiral is cavity backed, and the conductive cavity changes the antenna pattern to a unidirectional pattern. A two-arm spiral antenna shown in Figure 1.17 is excited in a balanced mode (that is, the same amplitude and a  $180^\circ$  phase difference between the two arms). Such a two-arm spiral radiates a CP wave in the antenna-axis direction normal to the antenna plane over a wide frequency range. For exciting this antenna, a coaxial line is used with a wideband balun circuit, which transforms the unbalanced mode of the coaxial line into the balanced mode required for the spiral antenna. It is noted that designing and installing such a wideband balun circuit for the spiral antenna requires considerable effort [21–23].

### 1.3.6 CP Dielectric Resonator Antennas

DRA is a resonant antenna fabricated from low-loss microwave dielectric materials. The resonant frequency of a DRA is a function of its size, shape and dielectric permittivity. Compared to other antenna types, DRA offers several attractive features. For example, DRA can achieve high radiation efficiency ( $>95\%$ ), flexible feed arrangement, simple structure, small size and the ability to produce different types of radiation patterns using different modes. In particular, DRA avoids conductor losses in patch antennas and is useful for applications at millimetre-wave frequencies. Various shapes of resonators can be used (rectangular, cylindrical, hemispherical, etc.), and various modes can be excited, producing broadside or conical-shaped radiation patterns for different coverage requirements. A wide range of permittivity values can be used (from about 6 to 100), thus antenna engineers can have control over the antenna size and bandwidth. A wideband DRA can be achieved using low permittivity while a compact size can be achieved with high permittivity.

Figure 1.18 shows a DRA antenna for circular polarization. A square DRA is fed by two orthogonal microstrip lines connected to a  $90^\circ$  microstrip hybrid. A metallic ground plane is at the bottom. This is a typical dual-feed technique as in the case of CP microstrip patch antennas. The required dual-feed network with a  $90^\circ$  microstrip hybrid takes up lots of space and increases the insertion loss (hence decreasing the radiation efficiency). An alternative



**Figure 1.18** CP dielectric resonator antenna

technique is to employ a single-feed technique. As discussed in Section 1.3.1, a single-feed square patch antenna with perturbations can excite two orthogonal modes in the patch and achieve CP operation. Similarly, in the case of DRA, it can achieve CP by using a quasi-square DRA with a single feed. Compared to the single-feed microstrip patch antenna which usually achieves narrow CP bandwidth about 1–2% for 3dB AR, single-feed DRAs can achieve up to 7% CP bandwidth [24–26].

For a rectangular DRA, the resonant frequency can be calculated by

$$f_{MNK} = \frac{1}{2\sqrt{\epsilon\mu}} \sqrt{\left(\frac{M}{l}\right)^2 + \left(\frac{N}{w}\right)^2 + \left(\frac{K}{h}\right)^2} \quad (1.21)$$

where  $\epsilon$  is the permittivity,

$\mu$  is the permeability

M, N and K are integer numbers, and

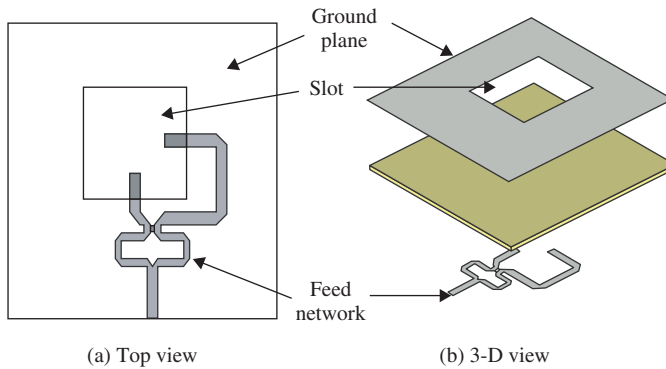
$l$ ,  $w$  and  $h$  are the length, width and height of the rectangular DRA, respectively.

From equation (1.21), it can be seen that the DRA can resonate at various modes, and the resonant frequency is inversely proportional to the square root of the product of material parameters. A high permittivity material will lead to a low resonant frequency of DRA. DRA can be fed using different techniques such as probe feed, aperture slot coupling, microstrip lines and CPW.

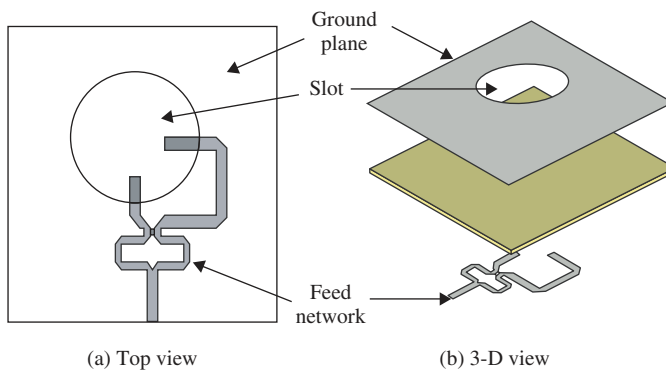
### 1.3.7 CP Slot Antennas

The printed slot antenna is very simple in structure: it consists of a microstrip feed that couples electromagnetic waves through the slot above and the slot radiates them. A microstrip-fed slot antenna is flexible in integration with other active and passive devices in a hybrid microwave integrated circuit (MIC) and microwave monolithic integrated circuit (MMIC) design. They are also easy to make as they can be cut into the surface of the platform they are mounted on. Slot antennas are able to achieve a broader bandwidth compared to microstrip patch antennas. Figure 1.19 shows a CP printed slot antenna. As shown, a square slot is cut in the ground plane, and fed by a feed network at the bottom. The feed network uses a Wilkinson power divider in microstrip lines. The two branches of power divider have lengths with a difference of a quarter-wavelength, leading to a 90° phase difference between two orthogonal modes in the slot excited by two orthogonal feed lines. Thus, circular polarization is produced in the square slot antenna.

To design a slot antenna, the length of slot is usually chosen to be a half-wavelength. The slot can take different shapes, such as a crossed slot, circular slot, annular ring, square ring and so on. Figure 1.20 shows another CP slot antenna which employs a circular slot fed by two orthogonal feed lines at the bottom. The printed slot antenna is easy to fabricate, has a low profile and low cost. However, the use of dual feeds and a feed network at the bottom as shown in Figures 1.19 and 1.20 occupies lots of space below the ground plane. For some applications, it is necessary to simplify the feed network by using a single feed instead. Figure 1.21 shows an example, which is basically a printed square slot antenna fed by a single microstrip line at the bottom. To achieve circular polarization, an L-shaped



**Figure 1.19** CP square slot antenna fed by a microstrip network at bottom



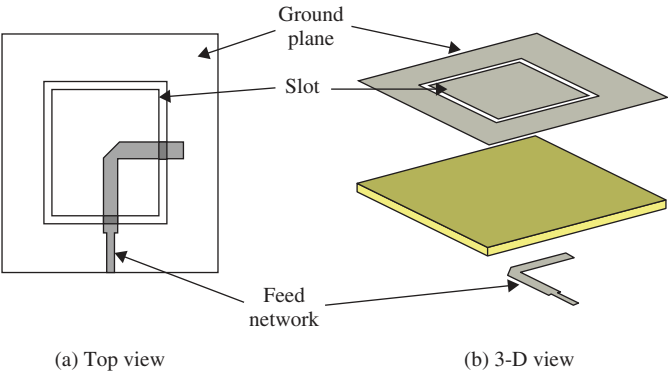
**Figure 1.20** CP circular slot antenna fed by a microstrip network

microstrip line is employed and the line excites two orthogonal modes in the square ring slot. The L-shaped microstrip line has a length of a quarter-wavelength, thus a  $90^\circ$  phase difference between two orthogonal modes in the slot is achieved. Such a CP slot antenna has a very simple feed network and is easy to implement.

In addition to the microstrip line feed shown previously, slot antennas can also be excited using other techniques such as coaxial cable or CPW. One drawback of these CP slot antenna is bi-directional radiation, as the slot will also radiate in the backward direction. Another ground plane can be added at the bottom so that the antenna can achieve broadside radiation only.

### 1.3.8 CP Horn Antennas

Horn antennas belong to the aperture antenna category, and their radiation performance is determined by the field distribution over the horn aperture. Horns are designed to provide a smooth transition between the feed waveguide and a wide aperture which serves to focus the main beam. Horns have found wide applications in satellite communications either as

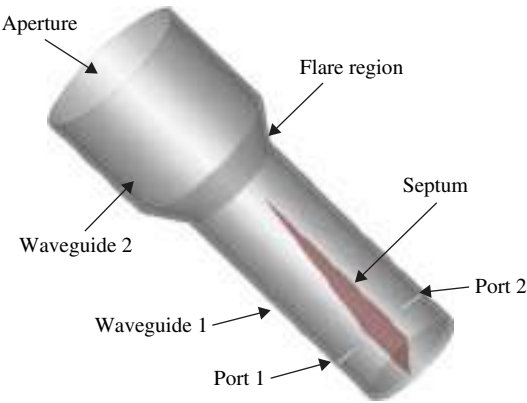


**Figure 1.21** CP ring slot antenna fed by an L-shaped feed

earth coverage antennas or feeds for reflector antennas. Theory and design of horn antennas with linear polarization are well documented in [7–9,11,27].

Most of the horn antennas reported have a single feed and can radiate linear polarization radio waves [27]. CP horn antennas can be realized by using a horn with dual orthogonal feeds and a  $90^\circ$  hybrid. However, the performance of dual linear polarized horn antennas, such as the ridged horn antenna, suffers from manufacturing and assembling tolerances. Also, the use of a  $90^\circ$  hybrid adds more complexity and losses in antennas. During recent years, many new techniques of CP horn antennas have been proposed [28,29]. A typical configuration of CP horn antennas includes three major elements, that is, a wave launcher, a polarizer and a beam shaper, as shown in Figure 1.22 [28,29].

The wave launcher in Figure 1.22 consists of two probes at input ports 1 and 2, which can achieve RHCP and LHCP, respectively. The input probes are located nearly one quarter-wavelength from the short-circuited end of the waveguide section 1, as shown in the figure. Waveguide 1 is a circular waveguide which allows the propagation of waves in TE<sub>11</sub> mode [27–29].



**Figure 1.22** Configuration of a CP horn antenna

The polarizer in Figure 1.22 consists of a circular waveguide divided into two identical semi-circular waveguides by a septum. The septum can use either a dielectric septum or a metallic septum. A metallic septum with sloped height is shown in Figure 1.22. In some cases, a metallic septum with stepped height can be employed to achieve a broadband performance [28,29]. By selecting the appropriate dimensions of the septum's steps, the TE<sub>11</sub> excitation converts into two orthogonal TE<sub>11</sub> modes of equal amplitude and 90° out of phase. The dimensions of the septum steps are very critical if these two conditions are to be met simultaneously over a wide frequency range. In the case that two excitation ports are used for obtaining RHCP and LHCP waves, high isolation between two ports is required.

The beam shaper in Figure 1.22 is a dual-mode circular waveguide (denoted as waveguide 2 in the Figure) which allows the propagation of waves in both TE<sub>11</sub> and TM<sub>11</sub> modes. Note that waveguide 2 has a larger diameter than that of the circular waveguide section 1, and there is a flare region between the two waveguides. The length of waveguide 2 is chosen such that the TE<sub>11</sub> and TM<sub>11</sub> modes' relative phase and amplitude allow the cancellation of the electric field at the aperture boundary, which results in a reduction of undesired sidelobes. A CP horn using a sloped septum can achieve about 10% 3-dB AR bandwidth and 40 dB isolation between two ports, while the use of a stepped septum can increase the bandwidth to 25% at the expense of lower isolation (about 20~25 dB) between two ports.

Other CP horn antennas have also been reported. For example, a horn excited with an L-shaped probe can be used to generate CP with 3 dB AR bandwidth of about 20% [30]. In [31], an oval shape waveguide polarizer is fed into a pyramidal horn antenna and a 3-dB AR bandwidth of 18% is reported. Note that the cross-section dimensions of the polarizer and horn antenna at the transition are different and they must be impedance-matched to minimize the generation of higher order modes.

### 1.3.9 CP Arrays

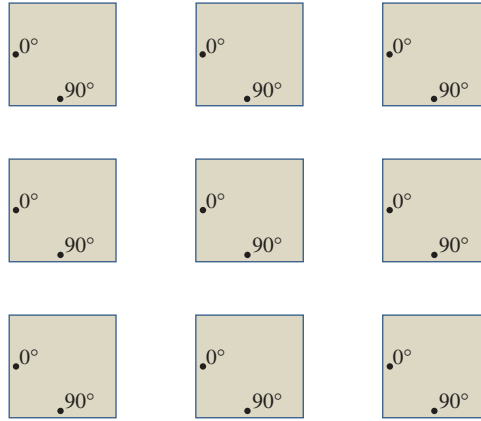
Previous sections have introduced different types of CP antenna elements. An array antenna consists of a number of antenna elements whose radiation is combined together to provide highly directive patterns. CP array antennas are needed in many applications such as satellite communications, inter-satellite communications, transmitting antennas onboard GPS satellites, and so on [11].

The array can have different configurations such as a 1D linear array or a 2D planar array. Figure 1.23 shows a 2D planar CP array with nine identical antenna elements. As shown, each element is a CP antenna with a square patch fed by two probe feeds located at two orthogonal edges of the patch with a 90° phase difference between them. Other types of antenna elements such as helix, DRA and so on, can also be employed to form CP arrays.

Assuming identical elements are employed and regular geometries are adopted in an array, the radiation pattern of the array can be obtained from the multiplication of radiation pattern of a single element,  $E_{element}(\theta, \varphi)$ , and the array factor  $AF(\theta, \varphi)$  [7]:

$$E(\theta, \varphi) = E_{element}(\theta, \varphi)AF(\theta, \varphi) \quad (1.22)$$

The array factor is a function of the array geometry, the inter-element distance, the element excitation in amplitude and phase, the number of elements and the frequency. For a 1D linear array of  $N$  identical elements which are equally spaced along the  $z$ -axis and excited with the



**Figure 1.23** A nine-element CP array

same amplitude and phase, the array factor is

$$AF(\theta, \varphi) = \frac{\sin\left(\frac{N\psi}{2}\right)}{\sin\left(\frac{\psi}{2}\right)} \quad (1.23)$$

where  $\psi = kd \cos \theta + \beta$  and  $k = \frac{2\pi}{\lambda}$ . Note a spherical coordinate system is employed.  $d$  and  $\beta$  denote the distance between elements and the phase shift between successive elements, respectively. For fixed-beam CP arrays,  $\beta$  is set to be a fixed value. The main beam of the CP array can be moved dynamically to point in different directions by changing  $\beta$  electronically. This is called a *phased array* [6,7,11,12]. Design techniques for CP arrays will be discussed in Chapter 5.

## 1.4 Antenna Modelling Techniques

### 1.4.1 Analytical Methods

Lots of analytical methods have been developed to analyse CP microstrip antennas, wire antennas, helical antennas, slots and DRA. For example, transmission-line models and cavity models have been developed for analysis and modelling of microstrip antennas [6–9,12,13]. Analytical methods have also been employed in the analysis of CP arrays [32,33]. These methods are very useful as they can provide a good insight into the physical mechanism of antenna operations. The drawback is the lack of accuracy, as these methods are usually based on some approximate assumptions.

### 1.4.2 Full-Wave Methods

The rapid development of computer technology has driven the development of various numerical methods during the past few decades. Full-wave methods mainly include the method of moments (MOM), the finite element method (FEM), and the finite difference

time-domain (FDTD) method. These methods are accurate as they directly solve Maxwell's equations without making physical approximations [34]. The precision of these methods is mainly dependent on the mesh discretization.

MOM is a versatile numerical method for solving integral equations [35]. One advantage of MOM is the variational nature of its solution: even if the unknown field is modelled to first order accuracy, the solution of MOM is accurate to the second order. MOM involves a significant amount of pre-processing of Maxwell's equations because it needs to use Green's function. MOM has been widely applied in analysis and modelling of all types of CP antenna elements and arrays. One disadvantage of MOM is that it needs to find the Green's function for a specific problem which can be difficult in some cases.

FEM is a numerical method for solving boundary value problems characterized by a partial differential equation and a set of boundary conditions [36]. The main advantage of FEM is that only a very sparse linear system is required to be solved, thus the memory requirement scales linearly with the number of degrees of freedom. In addition, the natural treatment of complex media makes FEM a very popular and powerful method in dealing with inhomogeneous structures. One disadvantage of FEM is the difficulty in dealing with open problems such as antennas. For antennas analysis and modelling, artificial absorbing boundary conditions (ABC)s, such as perfectly matched layers, should be applied around the object to simulate the radiation condition [37–41]. This increases the complexity and instability, unfortunately.

The FDTD method is another popular differential equation-based computational algorithm for antennas modelling [42,43]. Basically it is a time-domain numerical solver. It is simple to program, highly efficient and can be easily adopted. Also it can provide information of antenna radiation in the time domain, which is very useful in many cases such as wideband CP antennas. Fourier transforms are performed to obtain frequency-domain information if needed. Compared to FEM and MOM, the FDTD method can deal with large-scale array antennas simulation by using the parallel computation [44]. Furthermore, it is also a matrix-free technique. In recent years, the application of conformal meshes enables the FDTD method to analyse 3D antennas with curved structures [45,46].

## 1.5 Typical Requirements and Challenges in CP Antenna Designs

The requirements for CP antennas depend on the specific application: During recent years there has been an increasing demand for high data-rate broadband communication links using Ka-band mobile satellite communication systems [47]. The antenna can be put on board airplanes (e.g. for Internet in the sky), high-speed trains or cars. Such applications require CP antennas which have high gain, low profile, low mass, can be easily fabricated and are able to scan the beam electronically within a wide angle range in order to guarantee the service availability to the coverage areas. Future millimetre-wave WLAN and GNSS remote sensing will also require ultra-compact light-weight, low cost, wideband/multi-band high gain CP antennas with fast beam-scanning capability and low cost [48]. For mobile terminals in mobile communications or satellite communications, it is desirable to have small-size smart antennas at low cost [49].

GNSS receivers will need to be more compact, receive more signals and obtain higher accuracy in positioning, thus requiring miniaturized multi-band CP antennas which can cover all frequency bands of GNSS (GPS, Galileo, GLONASS, Compass, etc.), are low

cost, stable phase centres, the capability of multipath mitigation and adaptive patterns (for anti-jamming purposes) and are easy to integrate with RF circuits (filters, low noise amplifiers, mixers, etc.) [50–52].

RFID applications will require miniaturization of CP antennas with broadband, high gain, low cost and high reliability [53–55]. Future multi-band multi-function mobile terminals will require wideband/multi-band/multi-mode antennas which are re-configurable and can cover a variety of services such as (global system for mobile communications (GSM), Universal mobile telecommunications system (UMTS), mobile satellite communications, WLAN, GNSS, RFID, digital video broadcasting-handheld, etc.) while being small and low cost. The application of multiple inputs and multiple outputs (MIMO) technique into land mobile satellite communication will require broadband dual CP antennas. The growing of fields such as Internet of Things (IOT) [56,57], Long Term Evolution (LTE) [58], Cognitive Radio [59,60], body-centric wireless networks [61], satellite communication advanced flexible telecom payloads [11,47], small satellites (micro-sat, nano-sat, cube-sat) and formation flying [62–64], wireless power transmission, millimetre-wave, sub-millimetre-wave and terahertz systems [65] will drive the need for more innovative high-performance CP antennas.

The requirements here lead to lots of interesting challenges for antenna engineers. Typical challenges include the miniaturization of CP antennas, multi-band CP antennas, wideband CP antennas, high gain CP antennas, reconfigurable CP antennas, electronically beam steering CP antennas and so on. Chapters 2–5 will discuss various techniques of advanced CP antenna designs to meet these requirements. Some case studies will be provided in Chapter 6.

## 1.6 Summary

This chapter introduces basic parameters of antennas such as input impedance, reflection coefficient, gain, directivity, radiation pattern and antenna polarization. Different types of CP antennas including microstrip patch CP antennas, crossed dipoles, helix antennas, QHA, PQHA, spiral antenna, DRA, slot antenna, CP horns and CP arrays are also discussed. Basic principles of the operation of each antenna type are explained. Typical requirements and challenges of CP antenna designs are also discussed. Besides the CP antenna types mentioned, there are many other types of CP antennas, which can be realized using waveguide antenna, reflectors, lens antenna, leaky-wave antennas, slot arrays, hybrid combinations of different antennas, and so on [6–13].

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# 2

## Small Circularly Polarized Antenna

### 2.1 Introduction

Small-size CP antennas are of interest for a variety of applications, such as wireless sensors, hand-held GPS terminals, RFID readers, small WLAN access points and satellite communications mobile terminals. This chapter will start with an introduction to the basic theory of small antennas and the relationships amongst antenna size, quality factor and bandwidth are discussed. Then, different techniques of designing small CP antennas are presented and examples of different types of antennas (microstrip antennas, helix, QHA, PQHA, slot antennas, DRA) are illustrated. Finally, a table summarizing the performance of some typical small CP antennas is provided.

### 2.2 Basic Theory of Antenna Size Reduction

A small antenna means that the antenna is electrically small as a function of its overall size or occupied volume relative to the wavelength at its lowest resonant frequency. Usually, a small antenna is one for which [1–4]:

$$ka \leq 0.5 \quad (2.1)$$

where  $k$  is the free space wavenumber, and  $a$  is the radius of an imaginary sphere enclosing the maximum dimensions of the antenna. The lower bound of the antenna quality factor  $Q_{lb}$  is given by [3, 4]:

$$Q_{lb} = \left[ \frac{1}{(ka)^3} + \frac{1}{ka} \right] \eta_r \quad (2.2)$$

where  $\eta_r$  is the radiation efficiency of the antenna. The impedance bandwidth is inversely proportional to the  $Q$  value of an antenna as given by:

$$BW = \frac{VSWR - 1}{Q\sqrt{VSWR}} \quad (2.3)$$

As can be seen from equations (2.2)–(2.3), a small antenna will lead to a high value of  $Q$  and thus results in a narrow impedance bandwidth. For example, the size of a single-layer

microstrip patch antenna can be reduced by using high permittivity ceramic material as the substrate, but the antenna can have an impedance bandwidth of only a few percent, due to its high Q value. It also can be seen from equation (2.2) that an antenna making most use of the full volume of the enclosing sphere could achieve the optimum impedance bandwidth. The small antenna has been a popular research topic for some decades, and many techniques have been developed, such as the use of high-permittivity material, folded antenna, meander-line structure, capacitive loading, shorting PIN loading, and so on [1–4]. For CP antennas, the performance of AR will also deteriorate while the antenna size is decreased, because AR bandwidth is normally narrower than the impedance bandwidth of the antenna.

## 2.3 Small CP Patch Antennas

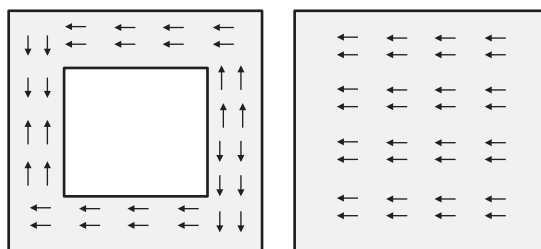
As explained in Chapter 1, CP patch antenna can be realized by either using a single-feed or multi-feed technique. In terms of size reduction, a single-feed CP antenna is usually preferred as a multi-feed CP microstrip antenna requires a phase quadrature feed network which is usually complicated and needs extra space. When a single-layer microstrip patch antenna is employed, the AR bandwidth of a single-feed CP antenna is usually narrower than that of a multi-feed microstrip patch antenna. Thus, there is a trade off amongst the size reduction, bandwidth and complexity during the CP patch antenna designs. In this section, different techniques of small CP patch antenna designs will be discussed.

### 2.3.1 Modified Shapes of Patch

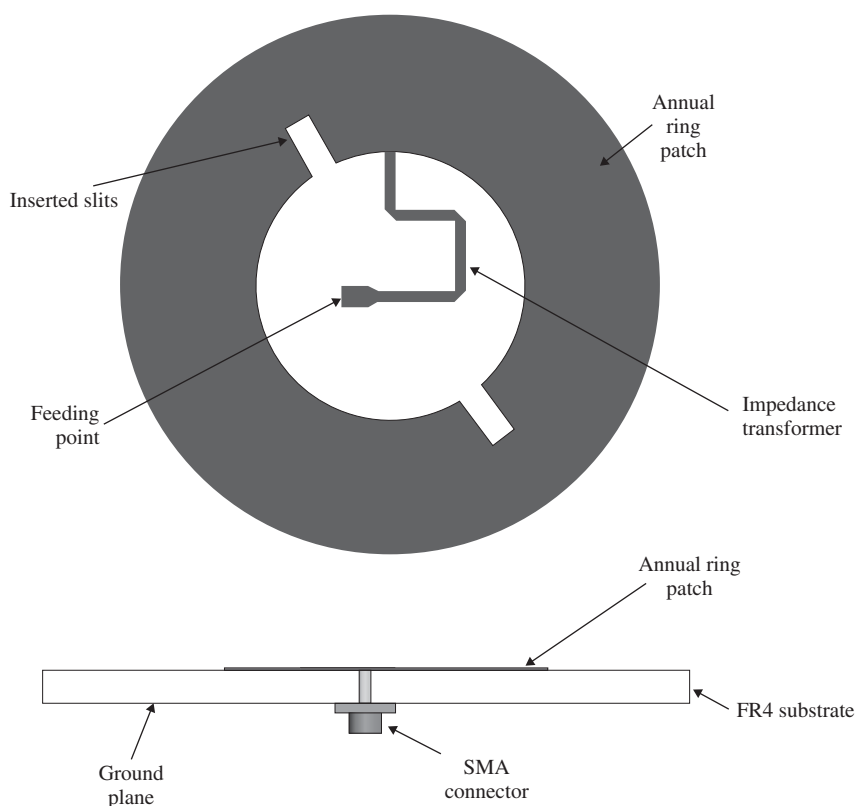
A common technique for antenna size reduction is to employ modified shapes of patch. For a single-feed CP patch antenna, the patch size can be reduced by using a nearly square ring, a corner-truncated square ring, an equilateral triangular patch with perturbation, an equilateral triangular ring with perturbation, a circular ring with perturbation, etc, instead of a nearly square patch or a nearly circular patch [5]. Figure 2.1 shows three examples of single-feed CP patch antennas with different patch shapes, including a corner-truncated square ring, a circular ring with perturbation and a corner-truncated equilateral triangular patch with cross strip. As explained in Chapter 1, small perturbations of regular patch shapes are required for exciting two orthogonal modes with a  $\pm 90^\circ$  phase difference. The square ring, circular ring or triangular patch occupies smaller area compared to that of a square patch or circular patch operating at the same frequency. To illustrate the principle of size reduction, Figure 2.2 compares the electric current distributions on a square ring patch and a square patch. As shown in Figure 2.2, the use of a square ring leads to the lengthening of electrical length of the patch compared to that of the square patch. This is the reason for the size reduction of CP patch antenna when using a ring-shaped patch.



**Figure 2.1** Different geometries of the ring structure for compact CP patch design [6–8]

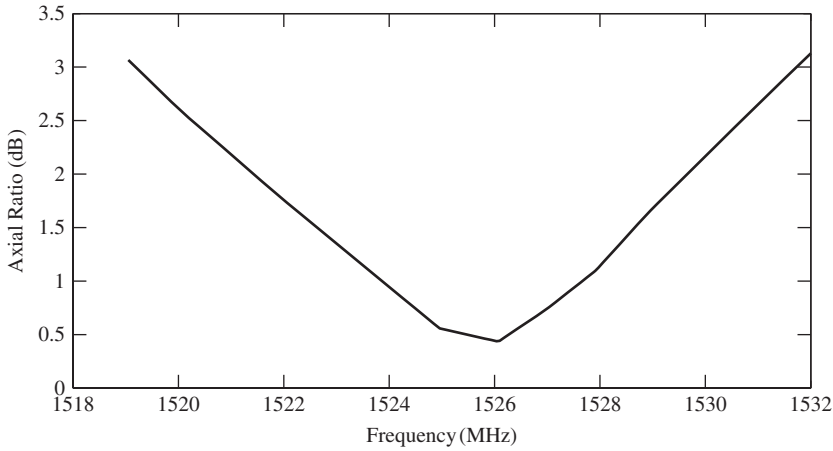


**Figure 2.2** The propagation of the electrical current (indicated by the arrows) on a square ring patch and a square patch of the same size



**Figure 2.3** Top and side view of the CP annular-ring patch antenna [6]. Reproduced with permission of © 1999 IEEE

Figure 2.3 presents one annular-ring patch antenna designed for CP operation at GPS L1 band (1.575 GHz) [6]. The antenna miniaturization is obtained by using the annular-ring patch. Two slits are inserted at the annular-ring patch to split the fundamental  $TM_{11}$  mode into two near-degenerated resonant modes and thus results in CP operation. With the configuration shown in Figure 2.3, the annular-ring patch radiates RHCP waves. Rotating

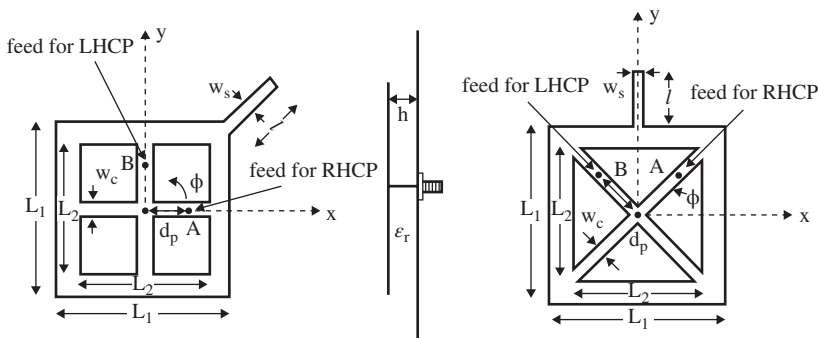


**Figure 2.4** Measured AR of the annual-ring patch antenna [6]. Reproduced with permission of © 1999 IEEE

both slits in a clockwise direction by  $90^\circ$ , LHCP can be obtained. This patch antenna is designed on an FR4 ( $\epsilon_r = 4.4$ ) substrate with thickness of 1.6 mm and the size of antenna is  $40 \times 40 \text{ mm}^2$ , which is  $0.2\lambda_{1.5 \text{ GHz}} \times 0.2\lambda_{1.5 \text{ GHz}}$ .

According to [6], compared to the conventional design that uses a circular patch of the same dimensions, about 24% reduction in the central frequency is achieved. The measurement results show that the 10-dB return loss bandwidth of this antenna is 2.1% and the 3-dB AR bandwidth is 0.8% with central frequency at 1.526 GHz. Figure 2.4 shows the measured axial ratio of this square-ring patch antenna.

The size of the CP square ring patch antenna can be further reduced by introducing a cross strip in the central or diagonal lines of the square ring patch, as shown in Figure 2.5. The cross strip can increase the current path of the fundamental mode and thus contribute to the size reduction of the patch antenna. As presented in Figure 2.5, the feeding positions



**Figure 2.5** Geometries of the square ring antenna with a cross strip in the central or diagonals position [9]. Reproduced with permission of © 1999 IEEE

**Table 2.1** Comparison between CP squaring ring antenna with and without cross strip

|                                                   | Size reduction compared<br>to Conventional CP<br>patch antenna (%) | 3-dB AR<br>bandwidth (%) |
|---------------------------------------------------|--------------------------------------------------------------------|--------------------------|
| Conventional CP patch antenna (Figure 1.9 (c))    | –                                                                  | typically 1–2%           |
| Square ring antenna [8]                           | 26                                                                 | 1.3                      |
| Square ring patch with a central strip [9]        | 33                                                                 | 0.89                     |
| Square ring patch with a diagonal cross strip [9] | 40                                                                 | 0.81                     |

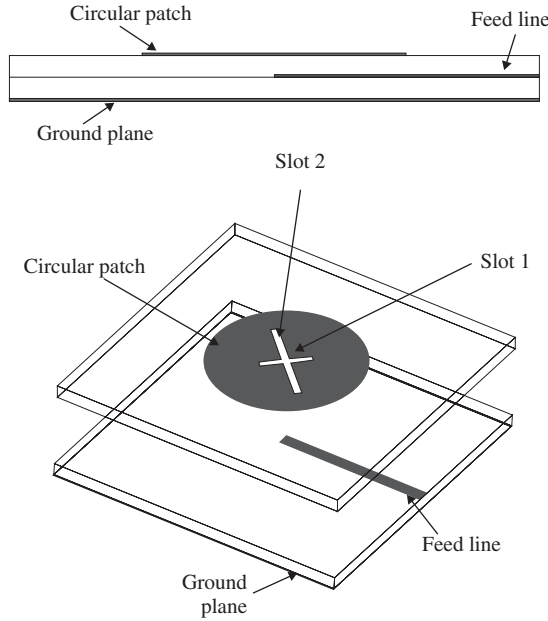
of both antennas are located at the cross strip. In order to generate the CP radiation, a small tuning stub is attached to the patch. It can split the fundamental mode into two near-generated resonant modes with equal amplitude and phase quadrature. Both of these two antennas are designed on a 1.6-mm-thick FR4 substrate. Compared to the conventional single-feed CP patch antenna (e.g. a corner-truncated square patch shown in Figure 1.9(c) of Chapter 1), the square-ring patch with a central line cross strip can reduce the antenna size by 33% while the square-ring patch with a diagonal cross strip achieves a size reduction of about 40%.

Table 2.1 compares the size reduction and the AR bandwidth of the three CP antennas reported in [8, 9] with the conventional single-feed corner-truncated CP patch antenna. As shown, the use of square ring or square ring with a crossed strip can lead to a size reduction but their AR bandwidth is narrower than that of the conventional single-feed CP patch. The square ring antenna with a diagonal cross strip shows the maximum size reduction, but it has a minimum AR bandwidth of only 0.81%.

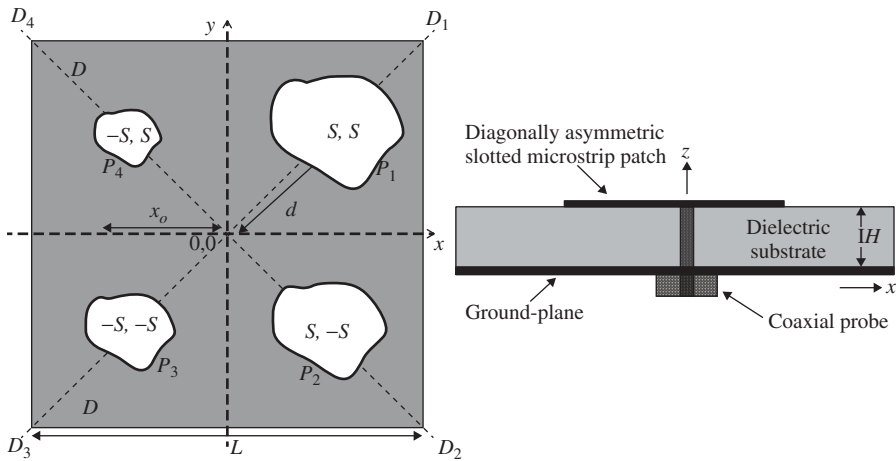
### 2.3.2 Slot Loading

Another technique for designing compact size CP patch antenna is to embed slots on the patch. One example is shown in Figure 2.6 [10]. As shown, crossed slots with unequal lengths are etched on a circular patch. By adjusting the lengths of both slots, the fundamental resonant mode of circular microstrip patch ( $TM_{11}$  mode) can be split into two near-degenerate resonant modes with equal amplitudes and  $90^\circ$  phase difference; thus, CP radiation can be achieved without using a  $90^\circ$  hybrid coupler. Figure 2.6 shows both the side view and exploded view of the cross-slot loaded circular patch antenna proposed in [10]. Both the circular patch and the microstrip feed line are printed on 1.6-mm-thick substrate with a relative dielectric constant of 2.6. By making the length of slot 1 ( $L_{s1}$ ) longer than the length of slot 2 ( $L_{s2}$ ), RHCP radiation can be obtained; if the  $L_{s1}$  is shorter than the  $L_{s2}$ , the antenna radiates LHCP waves. Compared to a conventional circular patch, about 36% size reduction is reached. The measurement results show that this antenna has a 2:1 VSWR bandwidth of 3.4% and 3-dB AR bandwidth of about 0.8% with central frequency of 1.525 GHz.

Instead of introducing cross slots, CP microstrip patch antenna can also be designed by introducing multiple asymmetric slots on the patch [11]. Figure 2.7 shows the top and side view of a square patch with four arbitrarily shaped slots. The centres of the slots are located along the diagonal line of the square patch with coordinates ( $S$ ,  $S$ ), ( $-S$ ,  $S$ ), ( $S$ ,  $-S$ ) and



**Figure 2.6** The configuration of the CP circular patch antenna with a cross slot [10]



**Figure 2.7** Square patch antenna with arbitrarily-shaped slots [11]. Reproduced with permission of © 2012 John Wiley & Sons, Inc.

$(-S, -S)$ , respectively. The distance from the centre of the patch to the centre of the slots is represented by  $d$  and the perimeter of each slot is represented by  $P_i$ . In order to generate two orthogonal modes with  $90^\circ$  phase difference, the location of these arbitrary-shaped slots must satisfy:

$$S \approx \frac{L}{4} \quad (2.4)$$

**Table 2.2** Relation between slot perimeters of the patch with arbitrary-shaped slots [11]. Reproduced with permission of © 2012 John Wiley & Sons, Inc.

| Slot(s) | Relation between slot parameters for CP                   |
|---------|-----------------------------------------------------------|
| One     | $P_1 > 0$ or $P_2 > 0$ or $P_3 > 0$ or $P_4 > 0$          |
| Two     | $P_1 < P_3$ or $P_1 > P_3$ or $P_2 > P_4$ or $P_2 < P_4$  |
| Three   | $P_1 > P_2 > P_3$ or $P_1 < P_2 < P_3$ or ...             |
| Four    | $P_1 > P_2 > P_3 > P_4$ or $P_1 < P_2 < P_3 < P_4$ or ... |

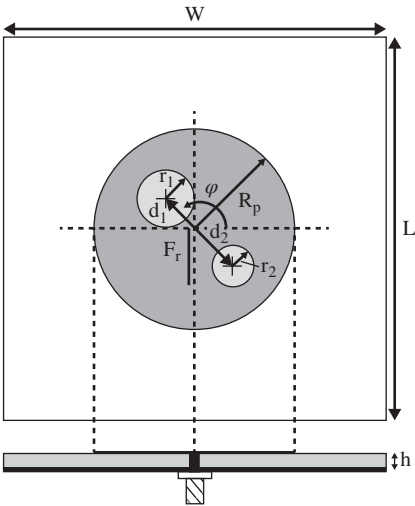
or

$$d \approx \frac{L}{2\sqrt{2}} \tag{2.5}$$

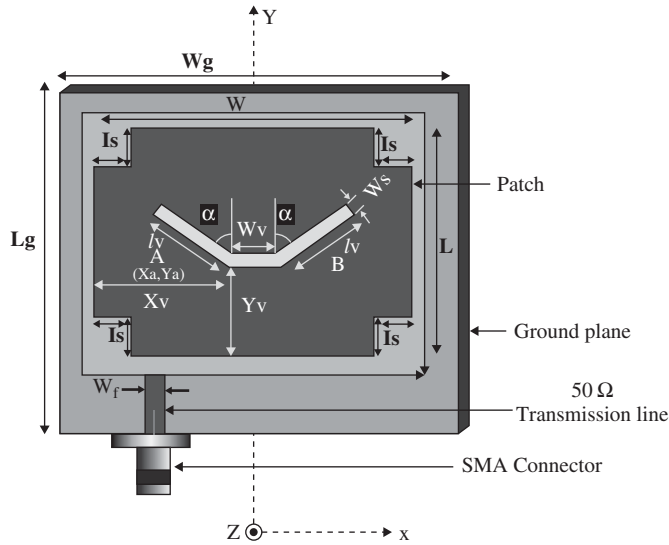
where  $L$  is the side length of the square patch.

An empirical table regarding the relationship between the slot’s perimeters and positions is summarized in [11] and given in Table 2.2. This table provides the guidelines for the CP patch antenna design with varied number (from one to four) of slots. To obtain the CP radiation, the perimeter of the slots must satisfy certain relationship and the exact value needs to be optimized for the desired resonant frequency by performing some numerical simulations.

One single feed CP circular patch with two embedded circular slots is reported in [12]. Figure 2.8 shows the geometry of this antenna. As can be seen from Figure 2.8, the two slots are introduced at the diagonal position of the circular patch with different size, as suggested in [11]. This circular patch is printed on an FR4 substrate with thickness of 1.6 mm. The radius of the circular patch is 15.5 mm and compared to the conventional



**Figure 2.8** Configuration of the CP circular patch with two circular slots [12]. Reproduced with permission of © 2008 John Wiley & Sons, Inc.

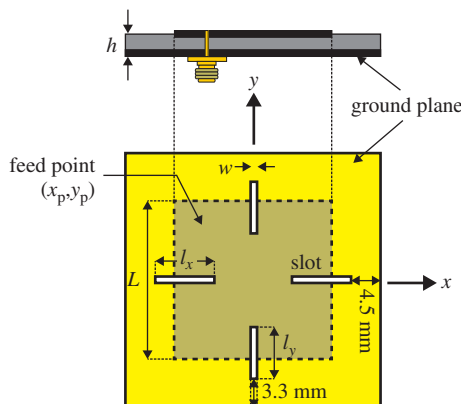


**Figure 2.9** Geometry of the cross shaped patch with slot [13]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.

design, there is a 12% size reduction. Measurement results show that its 10-dB return loss of bandwidth is 5.6% (from 2.397–2.512 GHz), and its 3-dB AR bandwidth is 1.3% with central frequency of 2.45 GHz.

There are other examples that employ embedded slots to design a CP patch. For instance, one compact GPS (L1 band) patch antenna is achieved by embedding a slot on a cross shaped patch, as shown in Figure 2.9. The use of cross-shaped patch can contribute to the size reduction of the antenna. The CP radiation is produced by letting the slot have some tilted angle and placing it symmetrically along the YZ-plane. For better impedance matching, proximity feeding is employed. It is also found that the impedance bandwidth of the antenna can be improved with the insertion of the slots. RHCP and LHCP radiation can be achieved by feeding the antenna at the left side and right side of the cross patch, respectively. The cross patch has a size of  $49 \times 35 \text{ mm}^2$  ( $0.245\lambda_{1.5 \text{ GHz}} \times 0.175\lambda_{1.5 \text{ GHz}}$ ) and is printed on an FR4 substrate with thickness of 1.6 mm. The reported 3-dB axial ratio bandwidth is 5% with central frequency of 1.57 GHz. The impedance bandwidth of this patch is 5.1% (1.53–1.61 GHz), within which it has average gain of 5 dBi [13].

The CP patch antennas discussed so far employ slots embedded on the radiating patch. It is also possible to achieve CP operation by cutting slots in the ground plane instead. It is demonstrated in [14] that this technique can lead to a small CP antenna with higher gain, broader bandwidth and higher radiation efficiency compared to the traditional CP antenna as shown in Figure 1.9(c) in Chapter 1. Figure 2.10 presents the configuration of the antenna proposed in [14]. The antenna is designed to operate at GPS L1 band and is etched on an FR4 substrate with thickness of 1.6 mm. The antenna element is a square patch of  $43 \times 43 \text{ mm}^2$  ( $0.21\lambda_{1.5 \text{ GHz}} \times 0.21\lambda_{1.5 \text{ GHz}}$ ) and the size of the ground plane is  $60 \times 60 \text{ mm}^2$ . The CP operation of this antenna is realized by introducing two pairs of narrow slots, which have different lengths on the antenna ground plane to excite two orthogonal near-degenerate



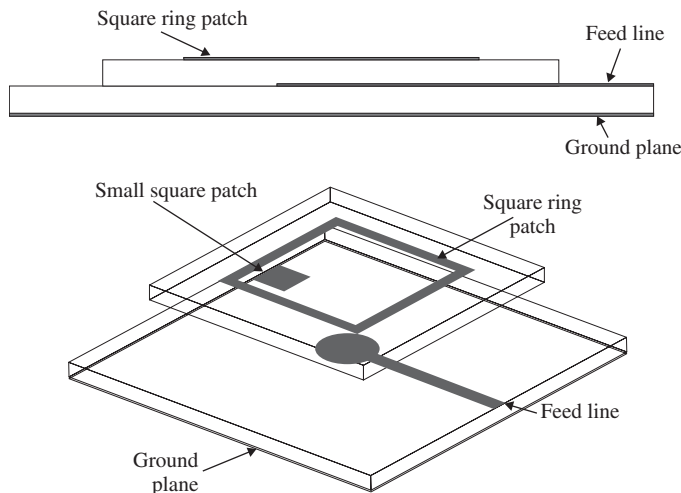
**Figure 2.10** Configuration of the CP antenna with a slotted ground plane [14]. Reproduced with permission of © 2008 John Wiley & Sons, Inc.

resonant modes. The patch antenna is fed at its diagonal position and with the feed at the position as shown in Figure 2.10, the patch can radiate RHCP waves. If the feed position is at the right side, this antenna can radiate LHCP waves. The measured 10-dB impedance and 3-dB AR bandwidth are 4.4% and 1.4% with central frequency at 1.5 GHz, respectively. Within the GPS L1 frequency band, the peak gain of the antenna is around 2.5 dBi.

### 2.3.3 High Permittivity Dielectric Material

One well-known technique of reducing the size of microstrip antenna is to employ a dielectric substrate with high permittivity. As this technique will decrease the bandwidth and radiation efficiency of the antenna, it is necessary to take the system requirements into the consideration before using this method. For certain applications, such as GPS antenna for hand-held device, small CP antennas are required to be equipped on portable devices such as mobile devices. In these scenarios it is more important to have a small antenna and the low gain performance can be compensated for by using high-performance receivers.

As introduced in Section 2.3.1, the size reduction of a CP antenna can be achieved by using a square ring patch. Therefore, combining the techniques of using a square ring antenna geometry and a high-permittivity dielectric substrate, CP patch antenna with significant size reduction can be reached. Figure 2.11 shows the structure of one single microstrip-fed square-ring patch antenna for GPS L1 band applications. Size reduction of the antenna is achieved by using a square-ring patch and a high permittivity ceramic ( $\epsilon_r = 30$ ) as the substrate. A small square patch is added at the same plane of the square ring for the purpose of making the square-ring patch antenna operate at its  $TM_{11}$  mode and enhancing the impedance matching. Attaching this square patch near the upper-left corner can make this antenna radiate RHCP waves. LHCP radiation can be obtained if this square patch is moved to right side. The overall volume of the antenna is  $15 \times 15 \times 3 \text{ mm}^3$ , which is only  $0.075\lambda_{1.5 \text{ GHz}} \times 0.075\lambda_{1.5 \text{ GHz}} \times 0.015\lambda_{1.5 \text{ GHz}}$ ; however, the bandwidth of this antenna is quite narrow (around 1.1%), the gain of this antenna is below  $-5 \text{ dBi}$  and the radiation efficiency is only about 11%, which is mainly due to the use of high-permittivity dielectric substrate.



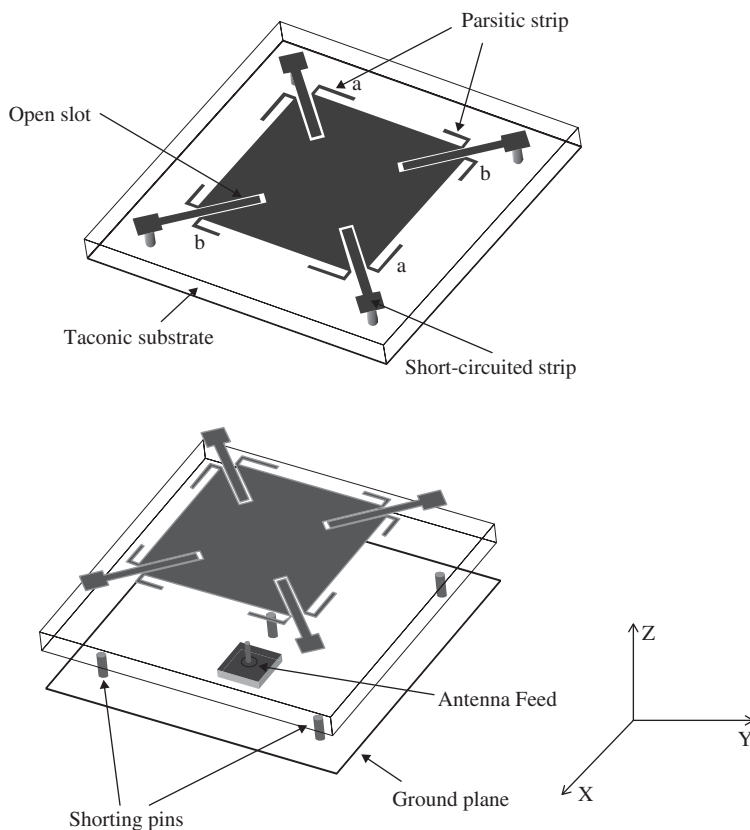
**Figure 2.11** Side and exploded view of the microstrip-fed CP square-ring patch antenna [15]

Although a significant antenna size reduction can be achieved by using high-permittivity dielectric substrates such as ceramic, the antenna has a narrow bandwidth, poor radiation efficiency and low gain. This technique often leads to poor radiation efficiency due to the excitation of surface waves on the substrate when a high-permittivity substrate is employed. Despite these drawbacks, this technique is very effective in reducing the antenna size and has been widely employed in many applications. The so-called ‘chip antennas’, commercially available nowadays, are based on the principle of using high-permittivity dielectric materials. As will be shown in the following sections, this technique can be also applied to other types of antennas such as PQHA.

### 2.3.4 Shorting Pins

Employing shorting pins or shorting walls is one of the methods widely used in the design of small patch antennas. For example, the planar inverted-F antenna (PIFA) is always shorted to the ground with a shorting pin for the purpose of size reduction. Although that shorting the patch antenna to the ground can decrease the length of the conventional patch from half-wavelength to quarter-wavelength, the cross polarization of the patch antenna normally exhibits a higher level due to the asymmetric structure.

Compact CP patch antenna can also be designed by introducing shorting pins. In order not to deteriorate the radiation pattern of the patch, the shorting pins or walls need to be placed symmetrically. Moreover, to have the CP radiation, it is necessary to make the patch have two orthogonal modes with  $90^\circ$  phase differences. One shorted CP patch antenna is reported in [16] and Figure 2.12 presents the configuration of this design. In the aspect of antenna size reduction, four open slots are loaded at the four corners of the patch. Then, four parasitic short-circuited strips are introduced inside these open slots, which can provide an LC loading to the patch antenna and thus increase the effective resonant length of the patch.

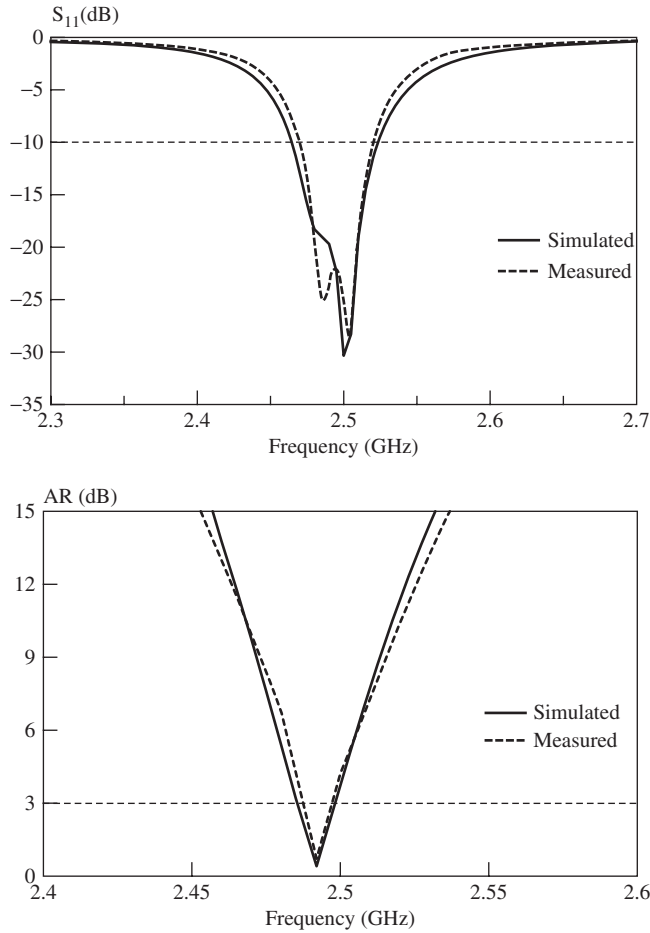


**Figure 2.12** Configuration of the compact circularly polarized patch antenna with shorting pins [16]

The use of the parasitic strips can suppress the unwanted radiation from the shorting pins: the strip can be considered as a capacitive loading whilst the shorting pin can be regarded as an inductive loading. To generate the CP radiation, two open stubs are attached at each corner of the patch. By adjusting the length of each pair of the stub (stub  $a$  and stub  $b$ ), two orthogonal modes with phase quadrature can be generated for CP radiation.

The CP patch antenna presented in Figure 2.12 is designed on a Taconic substrate with relative dielectric constant of 10 and thickness of 3.18 mm. The side length of the patch is 11.36 mm. With the parasitic shorting strips, it can resonate at 2.46 GHz with 3.25% impedance bandwidth and 0.682% 3-dB AR bandwidth. These two measurement results are given in Figure 2.13. According to [16], 82% size reduction has been achieved by using this technique compared to a conventional half-wavelength patch antenna.

Figure 2.14 shows the measured radiation patterns of the compact CP patch antenna at the XZ and YZ plane (the coordinate system is given in Figure 2.12). The antenna is shown to achieve a good broadside radiation with low level of cross polarization, because of the use of symmetry shorting pins. This proves the effect of using a parasitic strip to suppress the radiation from the vertical shorting pins.

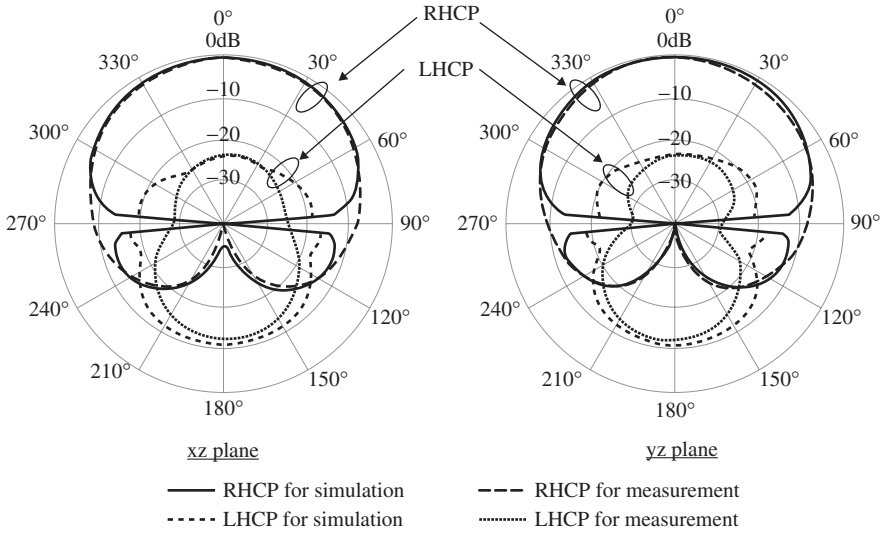


**Figure 2.13** Measured return loss and AR bandwidth of the compact CP patch antenna with symmetrical shorting pins [16]. Reproduced with permission of © 2010 IEEE

### 2.3.5 Metamaterial

As explained in Section 2.3.3, the use of a high permittivity substrate usually leads to poor radiation efficiency and narrow bandwidth because of surface wave propagation. To overcome this problem, the Electromagnetic Bandgap (EBG) structure is used as the high impedance surface (HIS) substrate for the CP patch antenna. With the HIS substrate, a gain improvement of 1.5 dB and improved axial-ratio bandwidth is reported [17]. An integrated solution, in which the antenna and artificial magnetic conductor (AMC) are fabricated by using multilayer PCB technology, is presented in [18]. However, it is observed that although these antennas have a low profile, the overall size of the antenna actually increases due to the use of the HIS as the substrate.

Using the so-called metamaterial (MTM) inspired antenna design technique, compact antennas can be constructed without increasing the overall volume of the antenna. Note that

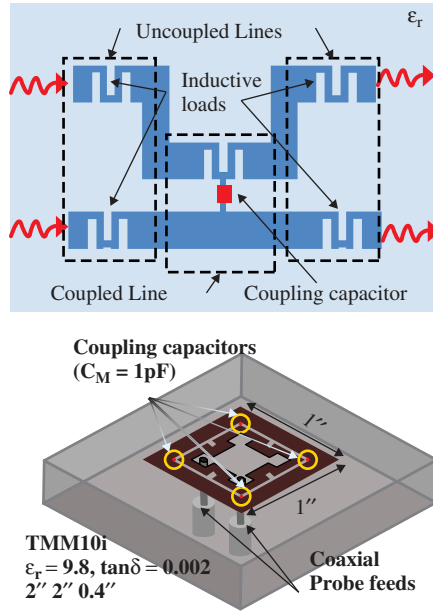


**Figure 2.14** Measured radiation patterns of the compact CP patch antenna with symmetrical shorting pins [16]. Reproduced with permission of © 2010 IEEE

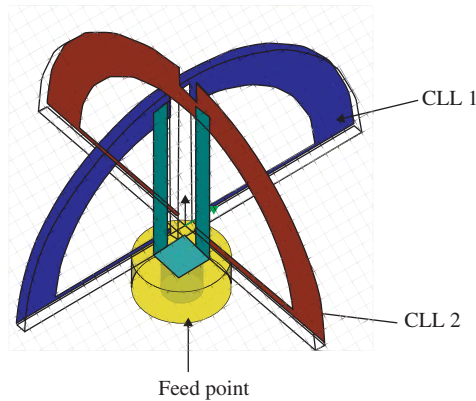
there are few designs on MTM inspired CP antenna and some of the proposed MTM-based antennas for the GPS application are only linearly polarized [19, 20]. Recently, works have been carried out to design CP GPS antennas by using MTM techniques. One miniature CP patch antenna is designed by using coupled double loop (CDL), which is an anisotropic material that can split the conventional  $K - \omega$  curves into two branches. The mode coupling between these branches can create novel slow group velocity modes [21, 22]. Figure 2.15 presents the concept of emulating anisotropy on a uniform substrate and one small antenna designed based on such CDL structure. The circular polarization is achieved by making the layout of the antenna be rotationally symmetric and feed it by two capacitively coupled coaxial probes with  $90^\circ$  phase differences. Controlling the input phase at each ports, the antenna can be made to either radiate RHCP or LHCP waves. This antenna is printed on a 10.1-mm-thick Roger TMM 10i ( $\epsilon_r = 9.8$ ) and the size of the antenna is reduced to  $25.4 \times 25.4 \text{ mm}^2$  ( $0.1\lambda_{1.29 \text{ GHz}} \times 0.1\lambda_{1.29 \text{ GHz}}$ ) as a result of using the slow wave anisotropic material. This antenna can operate at GPS L1 and L2 band. The simulation results show that it has 10-dB return loss bandwidth of 8 MHz and 14 MHz at 1.223 GHz and 1.575 GHz, respectively. The RHCP gain is 3.8 dBi at GPS L2 band and 3.3 dBi at GPS L1 band.

Another metamaterial inspired electrically small antenna (ESA) for CP GPS application is recently presented in [23]. This antenna is designed by employing the capacitively-loaded loop (CLL), which can be made to resonate at specific frequency by adjusting the shape of its gap to control its corresponding capacitance. RHCP or LHCP radiation can be obtained by using two orthogonally oriented capacitively-loaded loops with slightly length differences (to introduce the  $\pi/2$  phase difference). This antenna has a compact volume of only  $30 \times 30 \times 15 \text{ mm}^3$  ( $0.15\lambda_{1.5 \text{ GHz}} \times 0.15\lambda_{1.5 \text{ GHz}} \times 0.075\lambda_{1.5 \text{ GHz}}$ ) with a central frequency at around 1.57 GHz. The structure of this GPS antenna is shown in Figure 2.16.

Figure 2.17 shows the simulated return loss and axial ratio of this electrically small GPS antenna using CLL. It has 10-dB impedance bandwidth of 29.4 MHz and 3-dB

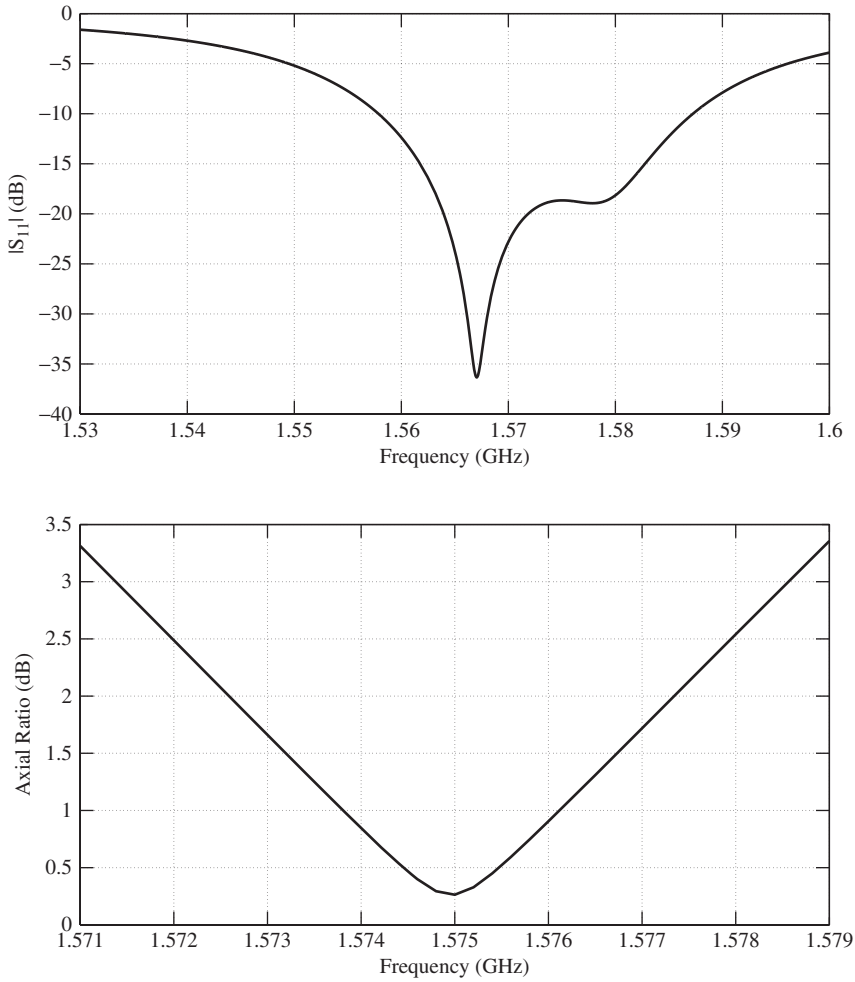


**Figure 2.15** The concept of the emulating anisotropy on uniform substrate and the compact CP antenna designed based on CDL structure for GPS application [21]. Reproduced with permission of © 2010 IEEE



**Figure 2.16** Structure of the electrically small GPS antenna designed by using CLL [23]. Reproduced with permission of © 2011 IEEE

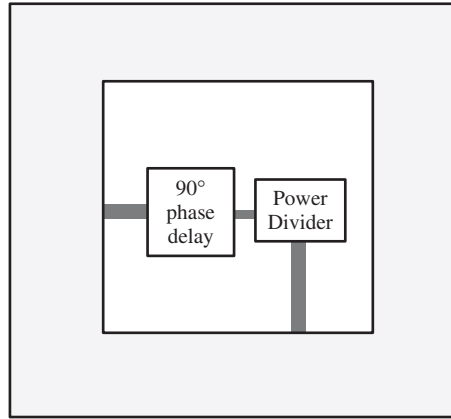
AR bandwidth of 7.4 MHz with central frequency of 1.57 GHz. This corresponds to an impedance bandwidth of 1.9% and AR bandwidth of 0.5%. Compared to the small CP patch antennas presented in last section, the MTM-inspired CP antenna shows great reduction in size but the bandwidth is quite narrow.



**Figure 2.17** Simulated return loss and axial ratio of this electrically small GPS antenna using CLL [23]. Reproduced with permission of © 2011 IEEE

### 2.3.6 Multi-Feed

Using a multi-feed technique to design a CP patch antenna normally results in an antenna of a larger size compared to the single-feed CP patch antenna, as it needs extra space for allocating the feeding lines and power dividers. In some applications, such as for the RFID hand-held reader, it is important to have one antenna of compact size while exhibiting wide bandwidth, moderate gain and high radiation efficiency at the desired frequency band. However, single-feed CP patch antennas have the inherent disadvantages of narrow impedance and AR bandwidth. Therefore, there is a need to design a compact CP patch antenna with relatively wide operation bandwidth and high radiation efficiency.



**Figure 2.18** The concept of using the multi-feed technique to design a CP square-ring antenna [24]

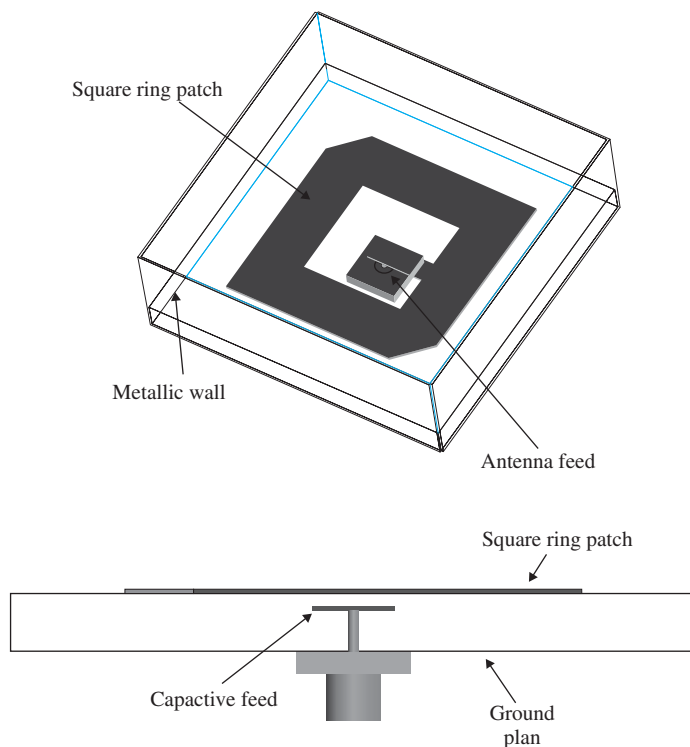
One possible solution is to introduce a multi-feed technique to a CP patch with ring geometry, such as square ring or annual ring. Figure 2.18 presents the concept of using multi-feed technique to design a CP square-ring patch. The power divider can be a Wilkinson power divider and the required phase shift can be created by using a quarter-wavelength microstrip line. Alternatively, a branch line coupler or rat-race coupler can also be used to feed the square ring patch antenna. Since the feed lines and the power divider are printed inside the ring patch, the size of the antenna is kept small; thus, compact size and good radiation performance can be obtained.

One such design is presented in [24], where a square-ring antenna is designed for portable RFID reader at Ultra-high frequency (UHF) band (860–960 MHz). This ring patch is fed by a Wilkinson power divider with two output microstrip lines that have length differences of a quarter-wavelength to produce a  $90^\circ$  phase shift, thus inducing the CP radiation of the antenna. The antenna is printed on an FR4 substrate with thickness of 1.6 mm and the size of the radiating square ring is  $95 \times 95 \text{ mm}^2$  ( $0.29\lambda_{915\text{MHz}} \times 0.29\lambda_{915\text{MHz}}$ ) with a strip width of 9.5 mm. The measurement results show that it can operate from 725–1149 MHz with a 3-dB AR bandwidth of 8.7% at central frequency (915 MHz) and the average gain is around 2 dBi within the UHF band. The measurement results show that with the multi-feed, the square-ring antenna exhibits much wider bandwidth than most of the single feed CP patch with moderate size reduction. Because the feed network is located inside the ring structure, it is important to make sure that the feed network has little influence on the radiation of the patch.

### 2.3.7 Other Small CP Patch Antennas

There are other techniques that can be applied to design small CP patch antennas. Some of these methods are variations of conventional techniques whilst some of the methods are combinations of several antenna miniaturization techniques.

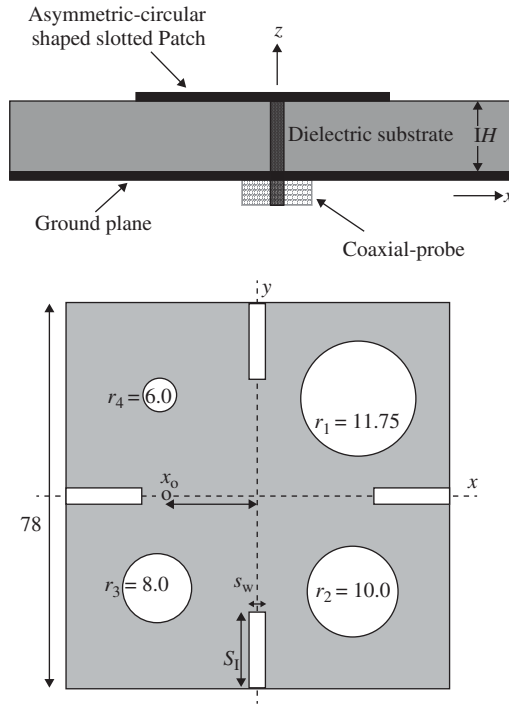
One compact RHCP square-ring CP antenna for RFID application at 920–925 MHz band is presented in [25]. This antenna has a similar structure to the one presented in [8] except



**Figure 2.19** Structure of the capacitively feed RFID CP antenna [25]

that the antenna is fed by a capacitively-coupled feed and vertical grounds are employed to surround the four sides of the antenna, for the purpose of enhancing the antenna gain and reducing antenna size. This antenna structure is shown in Figure 2.19. Employing the capacitively-coupled feed instead of using the direct coaxial probe feed, the feed position can be located more close to the centre of the patch to reach a better impedance match. According to [25], with the vertical walls the resonant frequency of the antenna can be further decreased by around 10% and the size of the antenna is 33% smaller than conventional one as depicted in Figure 1.9(c). Moreover, the gain of the patch is also improved. This antenna is designed on a 12 mm thick foam ( $\epsilon_r = 1.08$ ) and the dimension of the square-ring is  $0.34\lambda_{920\text{ MHz}} \times 0.34\lambda_{920\text{ MHz}}$ . The measurement results show that this antenna has return loss higher than 20 dB over 920–925 MHz and has a 3-dB AR bandwidth of 6 MHz (0.65%). Meanwhile, the gain is about 6.9 dBi within the desired frequency band.

Employing a patch with protruded structures has been used to design compact CP antennas. These designs include embedding asymmetric slits or symmetrical T-slits on the rectangular patch [26, 27], introducing triangular cuts and slits on the circular patch [28] and using a cross-shape slotted ground for the circular patch [29]. As an example, one asymmetric circularly shaped slotted patch antenna with slits for RFID handheld readers for operation at UHF band is proposed by [30] and its configuration is presented in Figure 2.20. CP radiation of this antenna is achieved by embedding four asymmetric circular slots along the



**Figure 2.20** Side view and top view of the asymmetric-circular shaped slotted patch antenna for RFID application [30]. Reproduced with permission of © 2010 IEEE

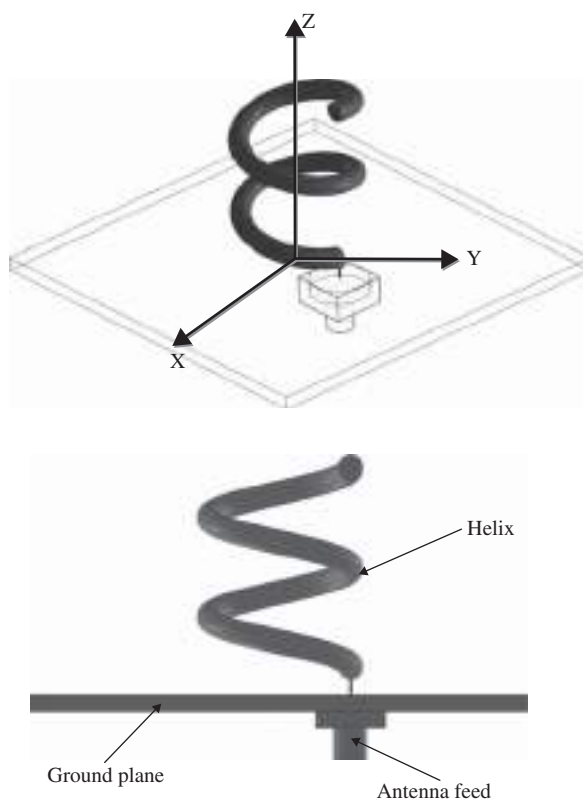
orthogonal direction with different radius. Four additional symmetric-slits are added along the orthogonal direction of the patch antenna for the purpose of antenna size reduction and antenna resonant frequency tuning. This antenna is printed on a 4.572-mm thick Roger 4003 ( $\epsilon_r = 3.38$ ) substrate and has a dimension of  $0.27\lambda_{900\text{ MHz}} \times 0.27\lambda_{900\text{ MHz}} \times 0.0137\lambda_{900\text{ MHz}}$ . The measured 10-dB return loss of this antenna is 4.01% and the 3-dB AR bandwidth is 1.02% with central frequency of 915 MHz. The minimum gain at the desired frequency band is 3.8 dBi. This design is a good example of combining different antenna miniaturization techniques to design a small CP patch.

## 2.4 Small Helix, QHAs and PQHAs

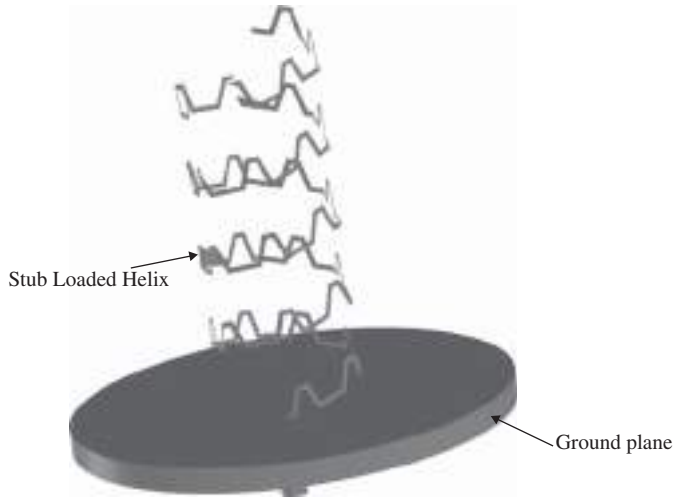
As discussed in Chapter 1, helix antenna, QHA and PQHA have been widely employed for CP applications. Basic operation principles of helix antenna, QHA and PQHA have been explained in Chapter 1. A helix can operate in axial mode and achieve broadband CP performance when the circumference of the helix is about one wavelength. For QHA and PQHA, the total length of each helical element needs to be equal to an integer number of quarter wavelengths. In the following, different techniques of reducing the size of helix, QHA and PQHA will be discussed.

### 2.4.1 Small Helix Antennas

As stated before, in order to have a helix antenna radiate CP waves, it has to operate at its axial mode, which means that the circumference of the helix is approximately one wavelength. Meanwhile, the number of the helical turns need to be more than seven and with pitch angle about  $11^\circ$  to  $15^\circ$  [31]. Since it is required to have the circumference of the helix approximately one wavelength, to make the helix antenna more compact, there are two approaches that can be employed. The first approach is to reduce the height of the helix antenna to make it have a low profile. It is reported in [32] that design a CP helix antenna with very low profile can be achieved by using small pitch angle and a small number of turns. It is shown that when the pitch angle is small, a good axial ratio can be achieved by reducing the number of turns. Figure 2.21 shows the structure of a low profile helix antenna above the ground plane. This antenna is designed to resonate at 12 GHz. The axial ratio of the helix antenna against the turns of the helix when the pitch angle varies from  $4^\circ$  to  $7^\circ$  is calculated. It is found that good AR can be obtained when the pitch angle is  $4^\circ$  with two turns of the helix (0.19 wavelength). In this case, the helix antenna shows 12% 3-dB AR bandwidth and



**Figure 2.21** The configuration of a helix antenna [32]



**Figure 2.22** The configuration of the Stub Loaded Helix [33]

9 dBi gain, which is comparable to a conventional long helix with an axial length of 0.9 wavelengths with a pitch angle of  $12.5^\circ$ . The calculated results in [32] also show that the half-power beamwidth of the radiation pattern is about  $70^\circ$  in both principal planes.

The second approach that can be used to design compact helix antenna is to introduce stubs around the circumference of each turn, and thus reduce both the size and height of the helix antenna. This technique is called Stub Loaded Helix (SLH) in literature [33]. Figure 2.22 presents the configuration of the SLH. As can be seen from this figure, the helical winding has loading stubs periodically placed around the circumference of each turn, which are oriented along a radial path from the central axis to the helical winding. With the existence of these stubs, the path around the helix is electrically longer than its physical circumference, thus resulting in a more compact structure. According to [33], the size reduction is about 25% compared to the conventional helix. Because the radiation from the adjacent stub wires effectively cancels out, there is little effect on the radiation pattern. Experimental results show that this antenna has a 3-dB AR bandwidth of 25%, which is narrower than the conventional one (50%). The smallest circumference that can be used without serious performance degradation is approximately  $0.6\lambda$  and the optimum pitch angle is between  $7^\circ$  and  $9^\circ$ . The maximum gain can be reached when the pitch angle is  $8^\circ$ . It is also noted that the SLH has a complex input impedance that varies with frequency, while for a conventional axial mode helix antenna it has real input impedance across its operational bandwidth.

#### 2.4.2 Small QHAs and PQHAs

Similar to the helix antenna, small QHA can be designed by modifying the geometry of the helices. By inserting a gap at the midpoint of each helix, where the current null is located, and extending the two half loops by adding a stub around the cylindrical surface means the size of the QHA can be decreased by around 15% [34]. By varying the length of the stub, it

is shown that the resonant frequency of the QHA can be tuned from 1200 MHz to 901 MHz with bandwidth better than 2% and size reduction of up to 25% can be achieved. According to the experimental results provided in [34], the gain of the antenna is always better than 5 dBi within its operation bandwidth. Since this QHA is air-loaded, although it has a larger size compared to the dielectric-loaded design, it has a higher radiation efficiency and better gain, both of which make it a good candidate for applications such as an RFID reader or WLAN receiver.

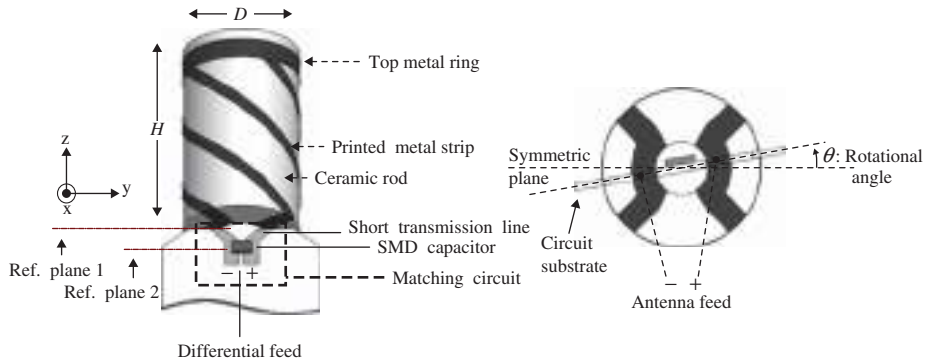
Another approach is to meander the wires of the QHA in order to increase its electrical length within the same volume, similar to the Stub Loaded Helix introduced in last section. By combining modified fractal curves and sinusoidal profiles, a very compact QHA is reported in [35] and the reported size reduction is 70%. However, it is important to point out that for such structures, the fabrication complexity is high and the fabrication accuracy is low; therefore, PQHA is usually preferred as it can resort to PCB technology for easy, low cost and high accuracy fabrication.

To design a small PQHA, which has gained its application in handset devices, such as for GPS reception, it is important to have a PQHA with a small volume. As the PQHA consists of two bifilar helices, exciting these parts with two slightly offset resonant frequencies  $f$  can produce a phase angle of  $+45^\circ$  (when  $f$  is below the operation frequency) and a phase angle of  $-45^\circ$  (when  $f$  is higher than the operation frequency) at the desired operation frequency, then a total of  $90^\circ$  phase difference can be reached [36]. This technique is named the self-phasing technique and can be realized by letting one of the bifilar helices be slightly longer than the other one.

With such a technique, PQHA with dielectric loading is proposed for the GPS application. The parametrical study in [37] shows that when a dielectric rod with a thickness of  $t$  is placed into the helix, either increasing the thickness  $t$  or increasing the relative permittivity of the dielectric can decrease the resonant frequency of the PQHA. In [37], the size of the PQHA is greatly reduced as the result of using high permittivity dielectric material with  $\epsilon_r = 36$  as the core. This dielectrically-loaded PQHA can resonate at 1575 MHz with a compact volume of  $17.75 \times 10 \times 10 \text{ mm}^3$  ( $0.09\lambda_{1.5 \text{ GHz}} \times 0.05\lambda_{1.5 \text{ GHz}} \times 0.05\lambda_{1.5 \text{ GHz}}$ ); however, its radiation efficiency is only 25% because of the use of high  $\epsilon_r$  dielectric material. This result is similar to the CP patch using high permittivity dielectric material presented in Section 2.3.3, where a very low radiation efficiency is also observed.

To alleviate the influence of high permittivity dielectric material, instead of using a solid dielectric rod, a hollow ceramic rod can be used as the dielectric load of the PQHA [38]. In this way, the effective permittivity of the ceramic rod can be reduced and thus, the PQHA can radiate more efficiently. The structure of the QHA is shown in Figure 2.23. This PQHA is fed by a  $100 \Omega$  differential line and a matching circuit is required.

The ceramic rod used in this design has a relative permittivity of 40 with loss tangent lower than 0.0001 at the frequency required for the GPS application (1.575 GHz). The volume of this PQHA is only  $14.8 \times 9 \times 9 \text{ mm}^3$  ( $0.07\lambda_{1.5 \text{ GHz}} \times 0.045\lambda_{1.5 \text{ GHz}} \times 0.045\lambda_{1.5 \text{ GHz}}$ ) with total radiation efficiency of 22%. According to [38], this size is only 2.7% of an air-loaded PQHA. To achieve the CP operation, a self-phasing technique is used. As shown in Figure 2.23, feed lines are connected to the circuit substrate that is rotated by a small angle  $\theta$ . With this rotation, the lengths of the two bifilar helices become different and in this way, a phase quadrature can be realized. RHCP can be obtained by having a positive  $\theta$  while



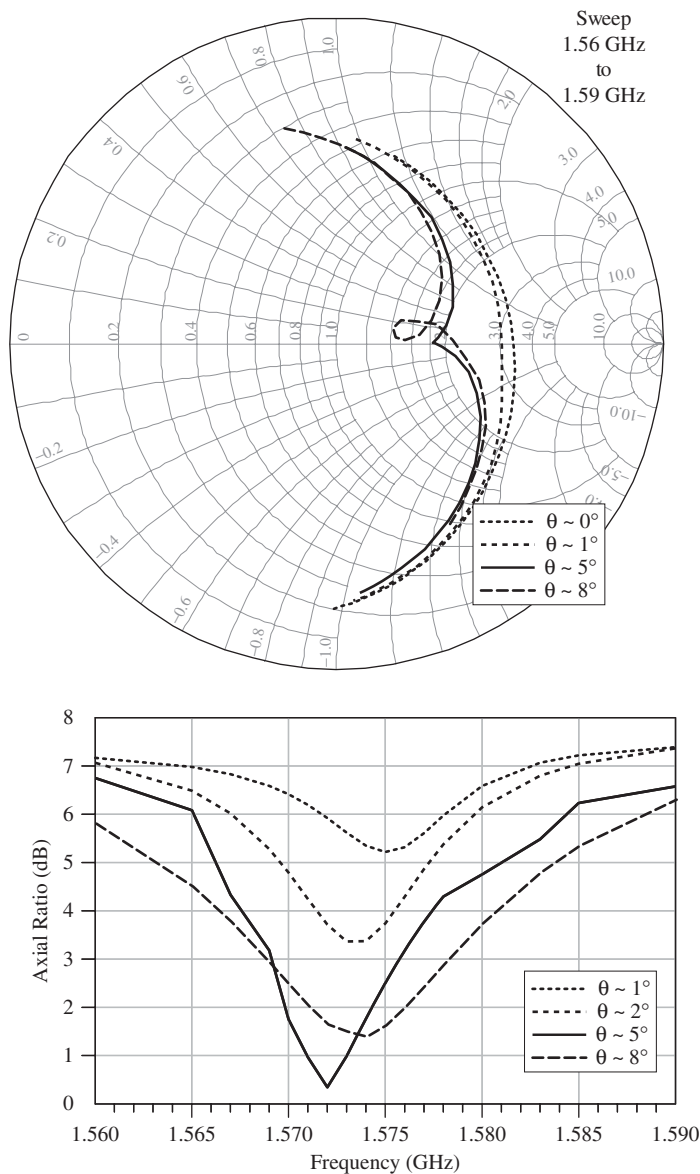
**Figure 2.23** Structure of the miniaturized PQHA using a hollow ceramic rod and its feeding structure [38]. Reproduced with permission of © 2009 IEEE

LHCP can be obtained by having a negative  $\theta$ . Both the feeding line and matching structure are printed on an FR4 substrate with thickness of 0.8 mm. Figure 2.24 shows the simulated input impedance and axial ratio of the PQHA with different value of  $\theta$ . It is clear that after optimizing this parameter, good circular polarization can be reached.

Figure 2.25 shows a commercially available small PQHA. This antenna is made by Sarantel Limited and is designed for GPS application at 1.5 GHz. A dielectric loading is employed to make the PQHA be more compact. Since this PQHA is designed to be fed by a coaxial cable, a quarter-wavelength balun is employed to convert the coax to 50  $\Omega$  balanced output.

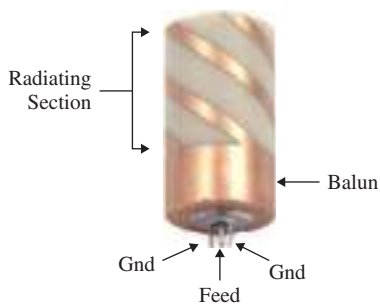
Similar to the small helix antenna design, the size of the PQHA can also be decreased by changing the configuration of each turn. Therefore, size reduction of a conventional PQHA antenna can be reached by using the meander line to replace the straight microstrip line normally employed on a classic PQHA, as presented in Figure 2.26. Using this concept, a meander printed quadrifilar helix antenna (MPQHA) is designed to resonate at 2 GHz [40]. The PQHA is printed on a flexible dielectric film and rolled into a cylinder. This PQHA can be tuned to resonate at a different frequency by simply resizing the diameter of the helix. The four ports of the antenna are fed by using one additional feed circuit, which provides excitation with equal amplitude and required phase differences at each port. The height of the antenna is 38.9 mm and the radius of the cylinder is 7 mm. According to [40], there is a 53% size reduction of this design compared to the conventional PQHA. Measurement results show that this antenna has a 10-dB return loss bandwidth of 190 MHz and 3-dB AR bandwidth between 1.9–2.1 GHz.

A slightly different version of the meander line loaded PQHA is reported in [41]. Figure 2.27 show the configuration of this PQHA. This PQHA is also printed on a flexible dielectric film. The size reduction of this design is 33%, which is 20% smaller than the one presented in [40]. Another variation of the MPQHA with similar structure is proposed by [42], in which the pitch angle of the PQHA is varied to achieve wider bandwidth. Alternatively, giving the helix a sinusoidal profile can also contribute to antenna miniaturization [43]. According to [43], height reduction from 48 to 72% compared to the conventional PQHA can be obtained by using various combinations of sinusoidal profiles. The experimental results indicate that although some decreases in the antenna

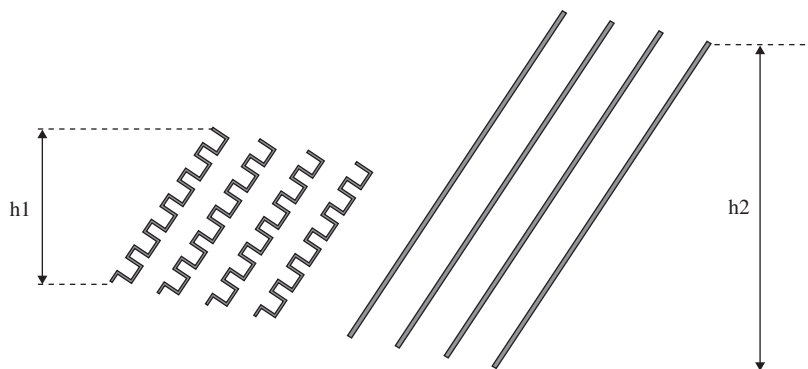


**Figure 2.24** Simulated input impedance and AR of the PQHA shown in Figure 2.23 with different rotation angle  $\theta$  [38]. Reproduced with permission of © 2009 IEEE

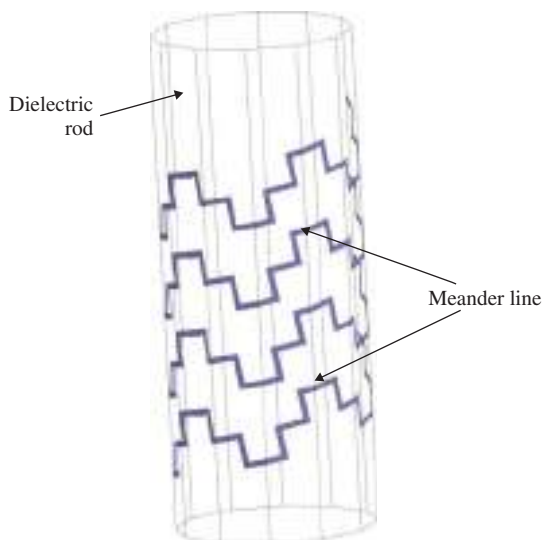
radiation efficiency and gain are observed (mainly caused by the antenna miniaturization) there is no significant degradation in terms of radiation pattern and polarization purity. It is important to point out that the meander line PQHAs presented in this section need to have an additional feed network to provide the desired phase and ideally equal amplitude to reach the expected CP radiation. Thus, the size of the feed network should be taken into account when calculating the overall volume of the PQHA.



**Figure 2.25** A compact PQHA for GPS application [39]. Reproduced with permission of © 2012 Sarantel



**Figure 2.26** The planar representation of the meander printed quadrifilar helix antenna and printed quadrifilar helix antenna [40]. Reproduced with permission of © 2002 IEEE



**Figure 2.27** The layout of the PQHA using different kind of meander line [41]

## 2.5 Small CP Slot Antennas

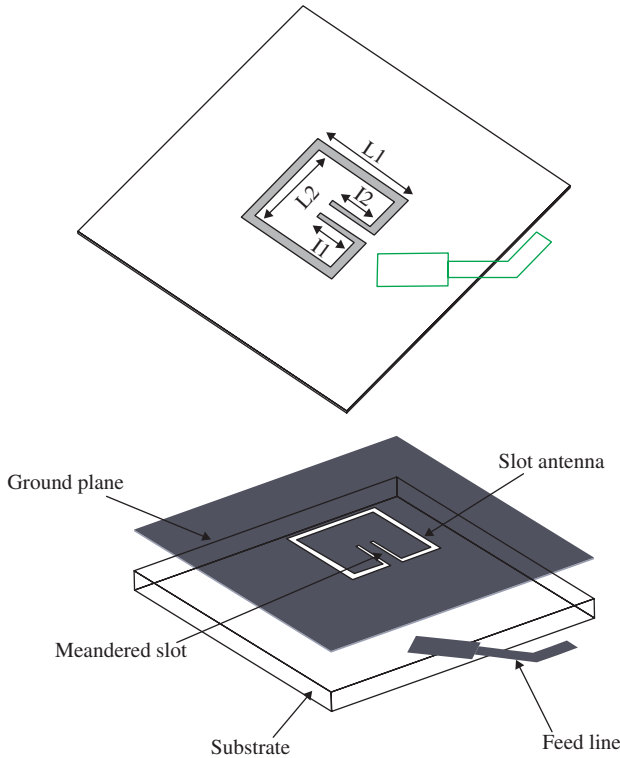
Generally speaking, printed slot CP antennas can achieve a broader bandwidth than that of printed microstrip patch antennas [44, 45]. To achieve circular polarization, it is necessary to excite two orthogonal modes with equal magnitude and a  $90^\circ$  phase difference between them. Chapter 1 has shown some examples of classic designs of CP slot antennas, such as a square or circular slot fed by a  $90^\circ$  microstrip feed network, and a square ring slot fed by an L-shaped microstrip feed line. For the conventional design, the length of the slot is about half a wavelength long, which is too large for portable devices (e.g. GPS devices). This section will discuss different techniques of designing small CP slot antennas.

### 2.5.1 Shapes of Slot

One approach to reduce the size of the conventional CP slot antenna is to introduce proper asymmetry in the ring slot, as presented in Figure 2.28.

For a conventional ring slot antenna to reach the resonant condition, it is required that the circumference of the ring is approximately one wavelength at the desired resonant frequency. After introducing the meandered slot section, the resonant condition becomes [45]:

$$f_0 \approx \frac{c}{2(L_1 + L_2) + l_1 + l_2} \times \left(\frac{1 + \epsilon_r}{2\epsilon_r}\right)^{1/2} \quad (2.6)$$



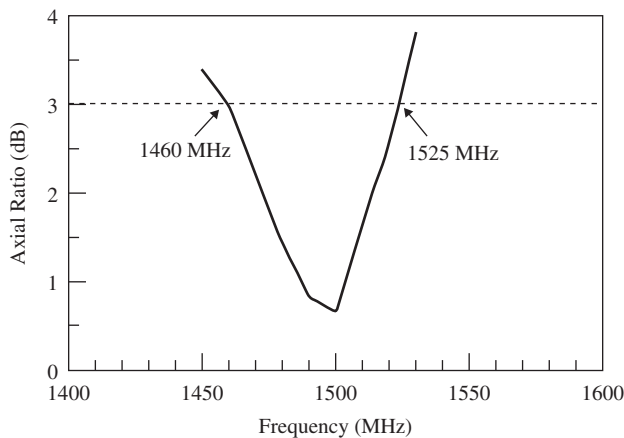
**Figure 2.28** The top and exploded view of the printed CP square ring slot antenna with asymmetry slot [45]

where the  $L_1, L_2$  represents the dimension of the slot and  $l$  means the length of the introduced asymmetry slot;  $c$  is the speed of light in free space and  $\epsilon_r$  is the relative dielectric constant of the substrate. The ring antenna is fed along the diagonal direction and a widened tuning sub is added at the end of the feed line for the purpose of enhancing the coupling between the microstrip feed line and the ring slot, which can result in better impedance matching. By feeding the antenna at the position as indicated in Figure 2.28, this slot ring antenna radiates a RHCP wave; the LHCP radiation can be obtained by changing the feed line to the other diagonal direction. This slot antenna is printed on a 1.6 mm thick FR4 substrate with  $\epsilon_r = 4.4$  and the size of the square ring is  $40 \times 40 \text{ mm}^2$ . Experimental result shows that this antenna has center frequency at 1.5 GHz with 4.3% 3-dB axial-ratio CP bandwidth (Figure 2.29), which covers the required frequency band for GPS application. Compared to the conventional design that only has a square-ring slot, there is about a 12% frequency reduction after the meandered slot is introduced.

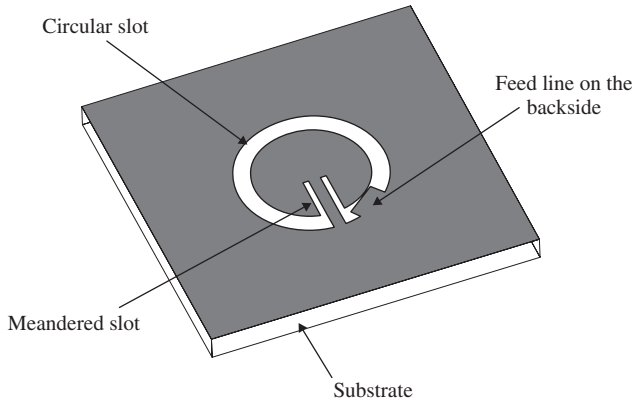
Instead of using a square ring slot, an annular ring slot can also be employed to design a small CP slot antenna, as the one presented in Figure 2.30. It is found that both antennas (shown in Figure 2.28 and Figure 2.30) have a similar radiation performance but with the same size (in this case, the diameter of the annual ring is chosen to be the same as the side length of the square ring), the square-ring slot antenna has a resonant frequency lower than the annular ring structure. Figure 2.31 shows the measured AR bandwidth of the printed CP annular ring slot antenna with an asymmetry slot. As shown, with an annular ring slot, the resonant frequency is about 13% higher than using a square-ring slot.

### 2.5.2 Monopole Slot Antennas

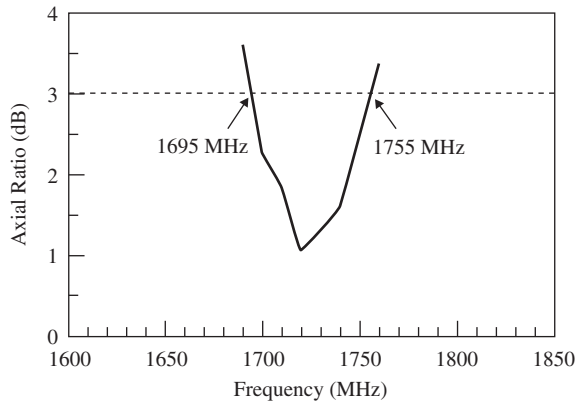
Another approach to reduce the size of the slot antenna is to design the antenna as a monopole slot antenna, which only needs to be a quarter-wavelength long to reach the resonant condition. In this way, it occupies less space compared to the conventional slot antenna. To design a CP monopole slot antenna, an L-shaped slot can be employed [46]. Figure 2.32 shows the



**Figure 2.29** Measured axial ratio of the antenna shown in Figure 2.28 [45]. Reproduced with permission of © 2002 IEEE

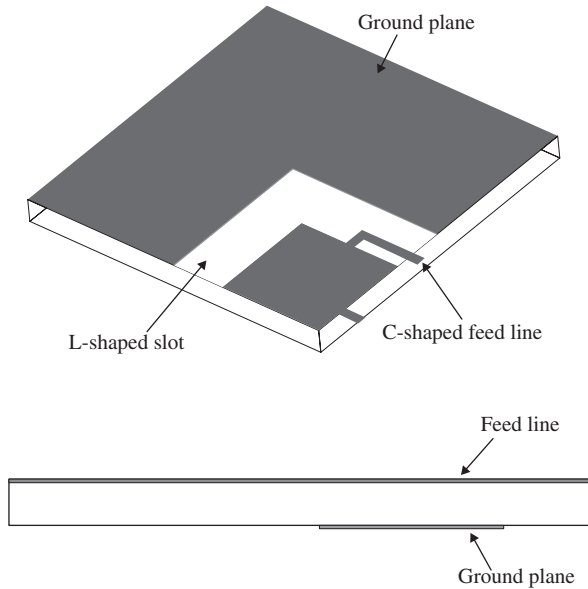


**Figure 2.30** The configuration of the printed CP annular ring slot antenna with an asymmetry slot [45]



**Figure 2.31** Measured axial ratio of the antenna shown in Figure 2.30 [45]. Reproduced with permission of © 2002 IEEE

structure of the L-shaped CP monopole slot antenna. The L-shaped slot, which has equal lengths at each arm, is cut at the left corner of the PCB board and a C-shaped feed line is used to feed the slot. By optimizing the width of the slot, the length of the horizontal and vertical portion of the feeding line, a 90° phase difference between the orthogonal electric field can be created. Using such monopole-like structure, besides being compact, wide-band operation can be achieved. Figure 2.33 presents both the simulated and measured return loss and axial ratio as well as gain of this antenna. These experimental results show that this antenna has a 10-dB return loss bandwidth of more than 25% and 3-dB AR bandwidth of 23% with central frequency of 1.68 GHz, which is several times larger than the ring slot antennas presented previously.



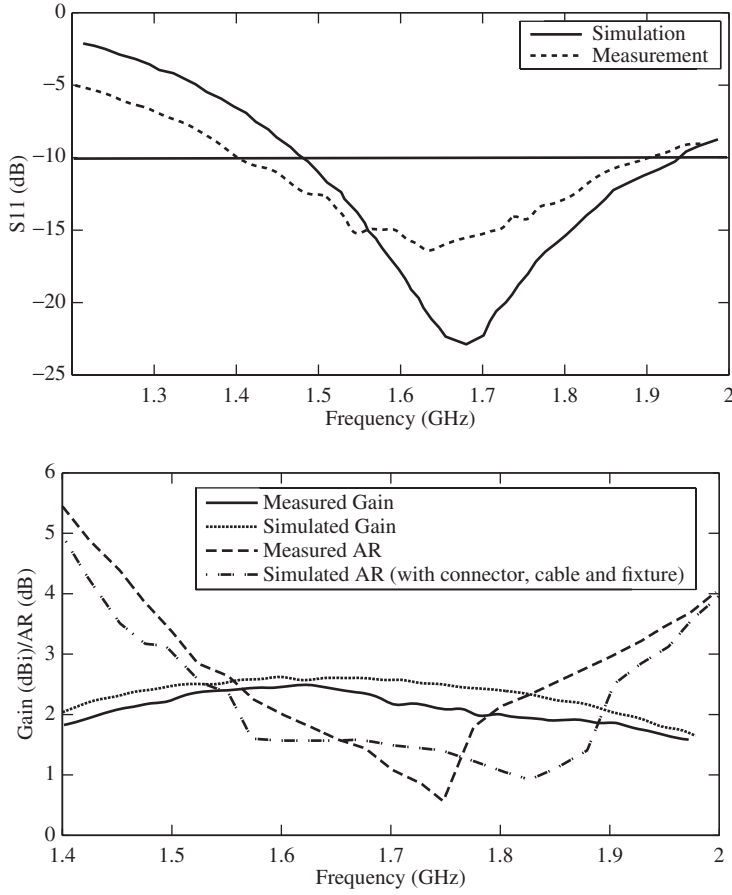
**Figure 2.32** Geometry of the L-shaped CP monopole slot antenna [46]

### 2.5.3 CPW Feed

The CPW-fed CP slot antenna has attracted much research interest because of its advantages of light weight, low cost and wider CP bandwidth than the microstrip patch antenna. There are several studies that have used CPW-fed structure to design wide-band CP slot antennas for different applications including WLAN and Ultra-wideband (UWB) [47–49]. However, in the aspect of size, CPW-fed CP antenna is normally larger than the microstrip antenna. Therefore, studies have been carried out to investigate how to miniaturize the CPW-fed CP slot antenna.

One technique that can be employed to design a small CPW-fed CP slot antenna is to introduce a halberd-shaped strip to the signal line, such as the one presented in [50]. Figure 2.34 shows the geometry of the CP printed square slot antenna. This antenna is printed on a 1.6-mm-thick FR4 substrate and the side length of the slot is chosen as 70 mm. The size of the gap size ( $g$ ) and the width of the signal strip ( $w_f$ ) are chosen to let the CPW have an impedance of  $50\ \Omega$ . The CP radiation of the antenna is induced by the use of the halberd-shaped strip. By properly adjusting the vertical ( $t_1$ ) and horizontal length ( $t_2$ ) of the halberd-shaped strip, the AR can be optimized. For this particular antenna, the optimum AR of the antenna can be reached when the value of  $\frac{t_1}{t_2}$  is around 1.08. A grounded asymmetric inverted-T strip is added at the upper left corner of the slot for the purpose of improving the impedance matching.

This CP slot antenna is designed to operate at 1575 MHz and the measurement results show that it has VSWR 2:1 impedance bandwidth of 5.2% and 3-dB AR bandwidth of 3.81%, which is obviously wider than most printed CP microstrip antennas. The overall size of the



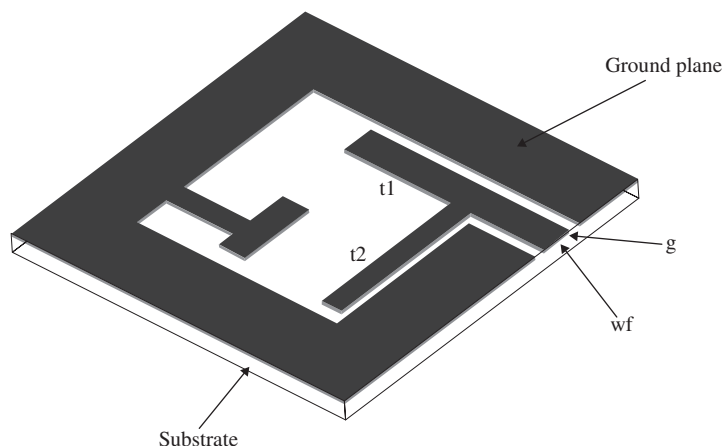
**Figure 2.33** Simulated and measured  $S_{11}$ , gain and axial ratio of the L-shaped CP monopole slot antenna presented in Figure 2.32 [46]. Reproduced with permission of © 2010 IEEE

slot antenna is  $0.36\lambda_{1.575 \text{ GHz}} \times 0.36\lambda_{1.575 \text{ GHz}}$ ) and this is reasonably compact compared to the conventional CPW-fed CP antenna.

More size reductions can be achieved through introducing corrugation to the ground plane [51]. The resonant frequency of the CPW antenna can be estimated by [52]:

$$f_0 \approx \frac{c}{2\sqrt{\epsilon_{eff}}} \times \sqrt{(1/L)^2} \quad (2.7)$$

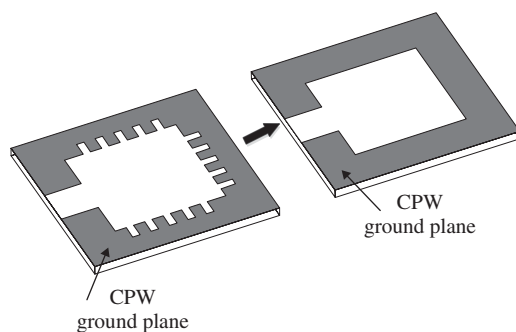
where  $c$  is the speed of the light,  $\epsilon_{eff}$  is the effective dielectric constant of the substrate and  $L$  is the side length of the square slot. Therefore, with the corrugation on the square slot, the side length of the slot ( $L$ ) can be increased, which results in the decrease of the antenna resonant frequency. Figure 2.35 shows the concept of this technique.



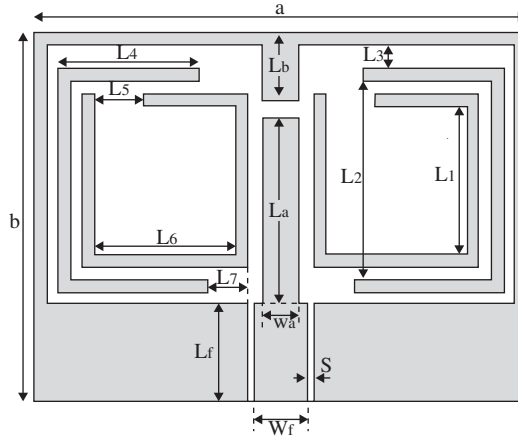
**Figure 2.34** Geometry of the CPW-fed CP printed square slot antenna [50]

By using this technique, a CPW-fed CP antenna is designed to have the central frequency at 1.42 GHz in [51]. The measured results show that compared to the design where no corrugations are introduced on the square slot [52], 44% size reduction has been achieved. The measured return loss bandwidth is about 17.1% and the 3-dB axial ratio bandwidth is 12.4%. It is also noticed that the complexity of this design is high and in order to generate the CP radiation, an additional meander line is required on the backside of the CPW ground plane. This means there are many parameters that need to be optimized in order to obtain the optimum performance of the antenna.

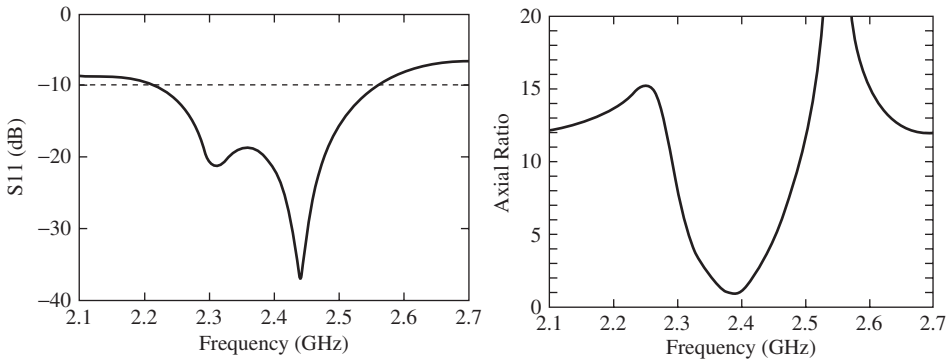
Employing multi-slot structures can also contribute to the size reduction of a CPW-fed slot antenna, as the design presented in Figure 2.36. In this design, by using two interleaved parasitic resonant strips (one with the C-shape and one with open ring structure) within the slot, two near orthogonal currents with different phases can be generated. After optimizing the lengths of these two parasitic resonant strips, with the configuration as presented in Figure 2.36, LHCP can be obtained. Changing the orientation of the resonant strips, this slot antenna can be made to resonate RHCP waves.



**Figure 2.35** The concept of using a corrugated ground plane to design a miniaturized CPW-fed CP slot antenna



**Figure 2.36** Geometry of the compact CPW-fed CP slot antenna [47]. Reproduced with permission of © 2011 IEEE

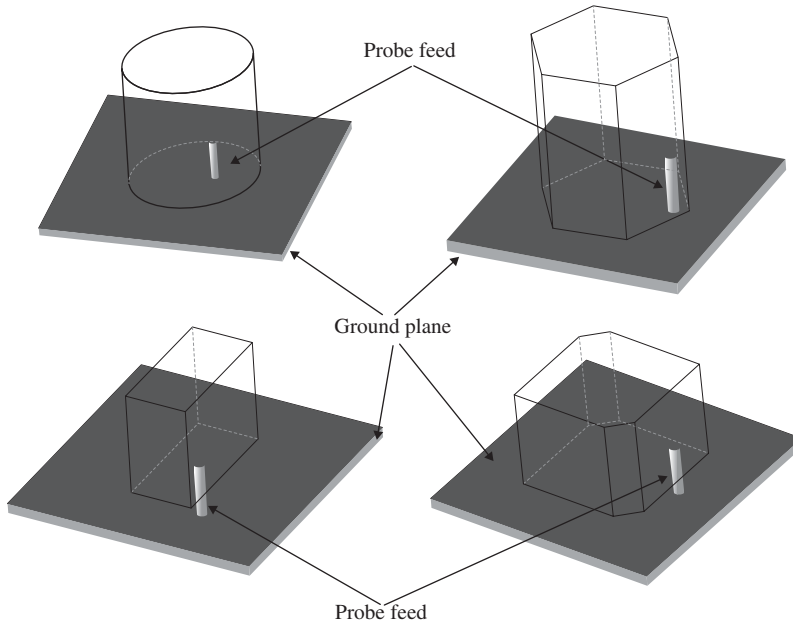


**Figure 2.37** Measured S11 and axial ratio of the antenna shown in Figure 2.36 [47]. Reproduced with permission of © 2011 IEEE

This antenna is designed on an FR4 substrate with thickness of 1.6 mm and the overall size of this antenna is  $30 \times 40 \times 1.6 \text{ mm}^3$  ( $0.24\lambda_{2.4 \text{ GHz}} \times 0.32\lambda_{2.4 \text{ GHz}} \times 0.128\lambda_{2.4 \text{ GHz}}$ ). The experimental results presented in [47] show that it exhibits 10-dB return loss bandwidth of 13.39% with central frequency at 2.44 GHz and 3-dB AR bandwidth of 3.77% with central frequency at 2.38 GHz, which covers the required frequency band for WLAN 2.4 GHz application.

## 2.6 Small CP DRAs

DRA has attracted much research interest due to the fact that it has less conductor loss and does not suffer from surface-wave losses compared to the microstrip antenna, especially for millimetre-wave applications [53]. It has been reported that, at up to millimetre wave frequencies, the DRA can radiate more efficiently than the microstrip antenna [54, 55].



**Figure 2.38** Some examples of conventional CP DRA [56–58]

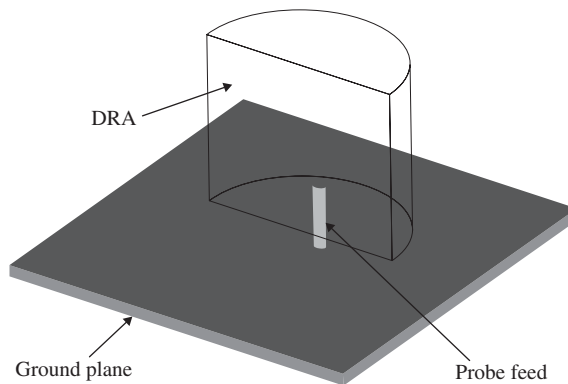
To design a CP DRA, like the microstrip antenna, it is possible to employ either a single-feed or multi-feed structure. Using the multi-feed structure can attain wider bandwidth; however, it is more complicated to implement and is usually larger compared with the single-feed DRA. The single-feed DRA is more compact but normally has a narrow AR bandwidth. Since the objective of this chapter is to address the small CP DRA, therefore, this section will focus on the single-feed CP DAR structure.

Common methods to design a single feed CP DRA include employing elliptic dielectric, changing the shape of the dielectric, using rectangular dielectric with feeding at its diagonal position and truncating the two corners of the square dielectric [56–58], as depicted in Figure 2.38. These techniques are quite similar to the ones used to design CP microstrip antennas, which all aim to excite two orthogonal modes with phase quadrature.

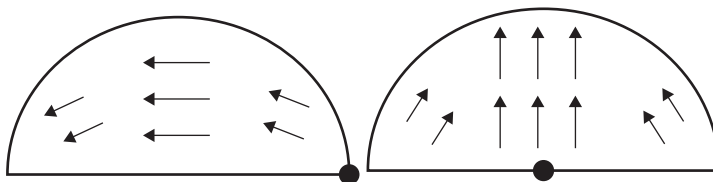
### 2.6.1 Different Shapes

Using a circular sector instead of cylinder-shaped dielectric, as shown in Figure 2.39, the size of the DRA can be reduced by 50%. The CP operation of this DRA is realized by adjusting the position of the feed to excite the fundamental resonant mode  $TM_{1,1,\sigma}$  and the next lowest resonant mode  $TM_{2,1,\sigma}$  of the DRA, perpendicular to each other as presented in Figure 2.40. Through driving the two modes smaller or higher than its resonant frequency by selecting an appropriate position of the feed along the radial edge of the DRA, a  $90^\circ$  phase difference can be obtained at desired frequency when:

$$f_1 + \frac{\Delta f_1}{2} = f_2 + \frac{\Delta f_2}{2}$$



**Figure 2.39** The geometry of the circular sector DRA [59]. Reproduced with permission of © 2000 IEEE



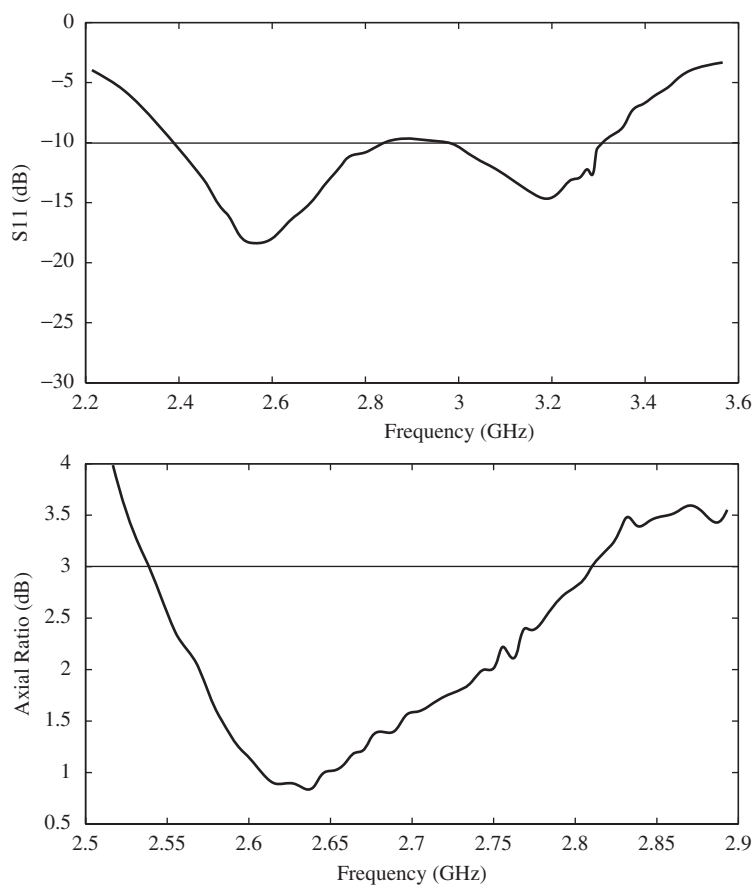
**Figure 2.40** The horizontally polarized  $TM_{1,1,\sigma}$  mode (left) and vertically polarized  $TM_{2,1,\sigma}$  mode (right) of the sector DRA [59]. Reproduced with permission of © 2000 IEEE

where  $f_1$  and  $f_2$  are the excited resonant frequencies of the orthogonal modes.  $\Delta f_1$  and  $\Delta f_2$  are the 3-dB bandwidth of  $|S_{11}|$  for the two modes.

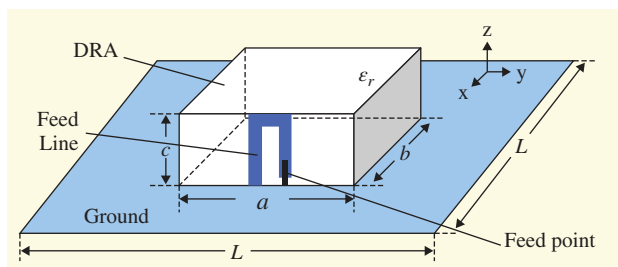
The volume of the DRA presented in Figure 2.39 is  $7.63 \text{ cm}^3$  and the relative permittivity of the dielectric is 12. Figure 2.41 shows the measured return loss and axial ratio of this sector DRA. Measurement results show that it has a 10-dB return loss bandwidth from 2.4–3.3 GHz and 3-dB AR bandwidth is from 2.54–2.81 GHz. Both bandwidths are wider than the conventional cylinder-shaped DRA while the sector DRA has a more compact structure. By simply resizing the antenna, it can be made to resonate at 2.4 GHz for WLAN application.

## 2.6.2 High Permittivity Dielectric Material

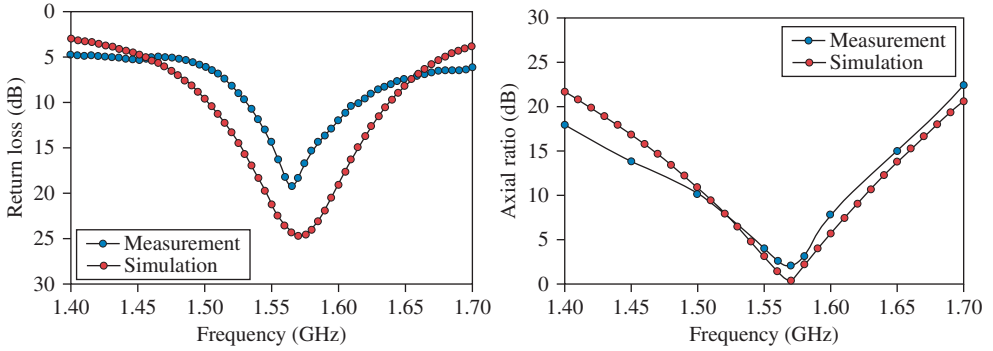
Another method to decrease the size of the DRA is to use high permittivity dielectric material. The consequence is that the bandwidth of the DRA will be narrow; however, for some narrow band applications such as GPS, this technique provides a simple and effective solution. One small CP DRA antenna using high permittivity dielectric material is presented in Figure 2.42. In this design, the DRA is a square ceramic with relative dielectric constant of 30 and it is fed by a loop-like microstrip line. By using this feeding technique, two different resonant modes,  $TE_{\delta 11}^x$  and  $TE_{1\delta 1}^y$ , can be excited with  $90^\circ$  phase difference.



**Figure 2.41** Measured return loss and axial ratio of this sector DRA [59]. Reproduced with permission of © 2000 IEEE



**Figure 2.42** Geometry of the CP DRA with single loop feed [60]. Reproduced with permission of © 2009 ETRI



**Figure 2.43** Simulated and measured return loss and AR bandwidth of the CP DRA presented in Figure 2.42 [60]. Reproduced with permission of © 2009 ETRI

With the feed position shown in Figure 2.42, the DRA radiates RHCP waves. LHCP radiation can be obtained if the feed point is moved to the left side of the loop. The central frequency of this DRA is 1.57 GHz and its volume is  $34 \times 34 \times 12$  mm, which is only  $0.17\lambda_{1.5 \text{ GHz}} \times 0.17\lambda_{1.5 \text{ GHz}} \times 0.06\lambda_{1.5 \text{ GHz}}$ . The measured 10-dB return loss and 3-dB AR bandwidth are 5.4% and 1.95%, respectively, as presented in Figure 2.43. Within the GPS band, the gain of this antenna is found to be 2.5 dBi.

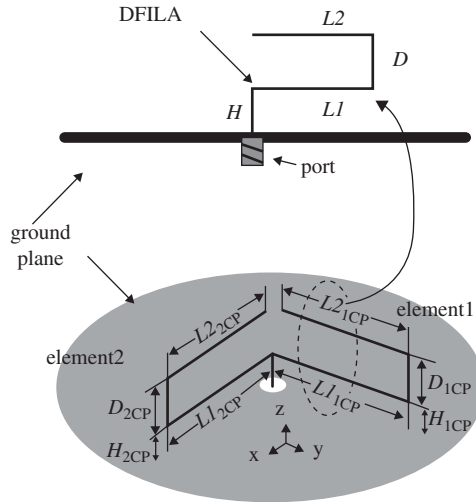
## 2.7 Other Small CP Antennas

### 2.7.1 CP Double-Folded ILAs

Besides the CP antennas introduced in the previous sections, there are other types of small CP antennas. For example, the Inverted-L antenna (ILA) is a kind of antenna widely used in the portable wireless devices. To design a compact CP ILA, double-folded structures can be employed [61]. Figure 2.44 shows the configuration of this low profile CP double-folded ILA (DFILA). As can be seen from this figure, the ILA has two arms, each of which is double-folded for the purpose of reducing the profile of the antenna.

The CP operation is obtained by letting the two arms have slightly different lengths, thus creating the desired  $90^\circ$  phase difference at the central frequency for CP radiation (the same technique that is applied to the small PQHA design and is presented in Section 2.4.2). This ILA is designed on a circular shape ground plane and the height of the ILA is only  $\frac{\lambda_0}{7}$ . The maximum lateral size of the ILA is about  $0.3\lambda_0$ , which represents a 64% size reduction compared to the normal crossed-dipole CP antenna as presented in [62], where the dipole only consists of a pair of straight wires.

Figure 2.45 shows the measured return loss and AR bandwidth of the DFILA. The measured results show that this small CP ILA has a 10-dB return loss bandwidth of 10.4% and 3-dB AR bandwidth of 2.92% with central frequency at 2.35 GHz. The return loss of this ILA with only one arm (element 1 or element 2 as indicated in Figure 2.44) is also presented in



**Figure 2.44** Configuration of the CP double-folded ILA [61]. Reproduced with permission of © 2010 IEEE

this figure. It can be seen that with only element 1, the ILA resonates at 2.42 GHz while with only element 2, the ILA has central frequency at 2.3 GHz. The length differences between these two elements are critical in determining the CP radiation of the ILA and when the ILA has both elements, it can resonate at 2.35 GHz with good AR.

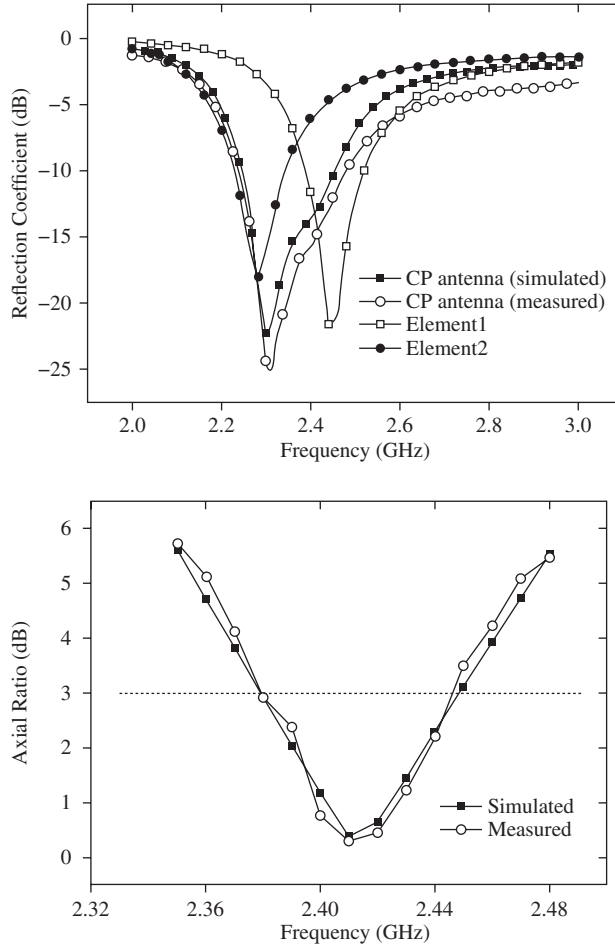
### 2.7.2 Compact Printed Dipole Antennas

A printed dipole antenna can also be used to design small CP antennas. Recently, a compact printed CP dipole is proposed by [63], in which the antenna is designed by using two dipoles connected parallel to a coaxial cable, as presented in Figure 2.46. The two dipoles are arranged orthogonally and the length differences between these two dipoles, which is determined by the ending arrows, generates a  $90^\circ$  phase difference. Making one dipole longer or shorter in length than the other one can control the LHCP or RHCP radiation of the antenna.

This antenna is designed for the RFID reader application at 924 MHz and is printed on a 1.6 mm thick FR4 substrate. The size of this antenna is only  $50 \times 50$  mm ( $0.15\lambda_{924 \text{ MHz}} \times 0.15\lambda_{924 \text{ MHz}}$ ), which is largely due to the use of meander lines. Measurement results show that this antenna exhibits 10-dB return loss from 900–965 MHz, 3-dB AR bandwidth of 1% with central frequency at 924 MHz. Moreover, it has peak gain of 0.8 dBi within the UHF band. Although this design has a narrow AR bandwidth (only 1%), the bandwidth is enough for RFID application at the UHF band.

### 2.7.3 Small Multi-Feed CP DRA Antennas

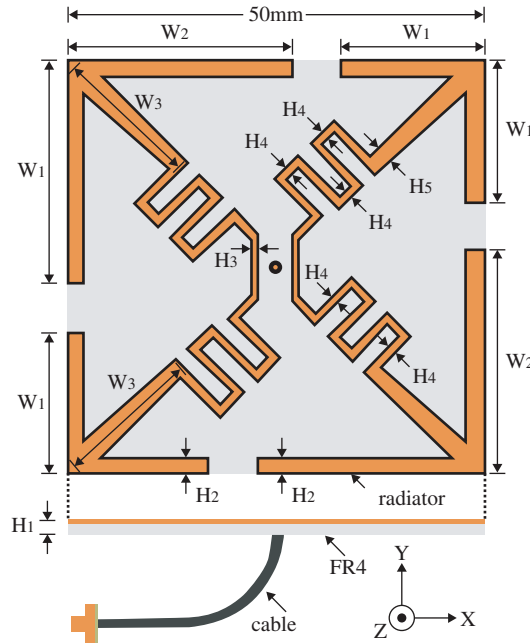
In the field of compact CP DRA antenna design, similar to other types of antennas, using single-feed DRA can result in a more compact structure but the bandwidth of it is always narrow. Using multi-feed structures can effectively increase the bandwidth of the CP DRA;



**Figure 2.45** Measured S11 and axial ratio of the CP double-folded ILA [61]. Reproduced with permission of © 2010 IEEE

however, it requires extra space for the feeding lines, which makes it difficult to keep the size of the antenna compact. A novel solution, which makes use of the dual-feed to design a wide band CP DRA while not needing an increase in the footprint of the DRA, is presented in [64]. The DRA is chosen to be a hollow rectangular dielectric. A quadrature coupler is located beneath the DRA and is placed entirely inside the hollow region of the DAR, as shown in Figure 2.47. With this configuration, the overall size of the antenna can be kept small. This technique is similar to the one presented in Section 2.3.6, where the feed network is located inside the square-ring patch.

The hollow DRA has a square cross section and the relative dielectric constant of the material is 10. The size of the DRA (not including the grounded substrate) is  $31.8 \times 31.8 \times 20.5 \text{ mm}^3$ , which is around  $0.25\lambda_{2.4 \text{ GHz}} \times 0.25\lambda_{2.4 \text{ GHz}} \times 0.16\lambda_{2.4 \text{ GHz}}$ . To realize CP radiation, a pair of adhesive conducting strips are stuck on two adjacent side walls of the DRA for

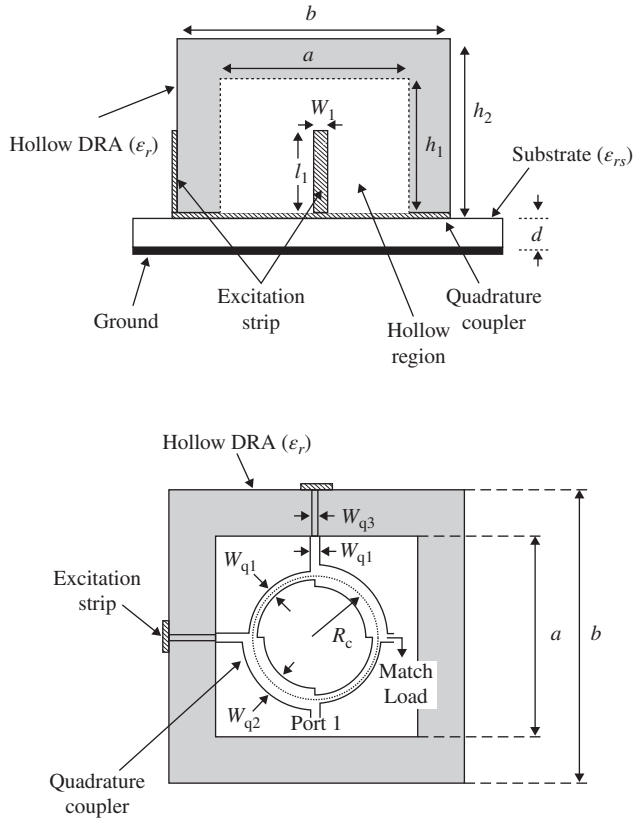


**Figure 2.46** Structure of the compact RFID reader antenna using unequal length dipoles [63]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.

the purpose of exciting two orthogonal modes. This DRA is designed to operate at 2.45 GHz and the experimental results (Figure 2.48) show that it has a 10-dB return loss of 24.95% and 3-dB AR bandwidth from 2.02–2.8 GHz with central frequency at 2.4 GHz, which is obviously several times wider than single feed DRAs.

## 2.8 Summary

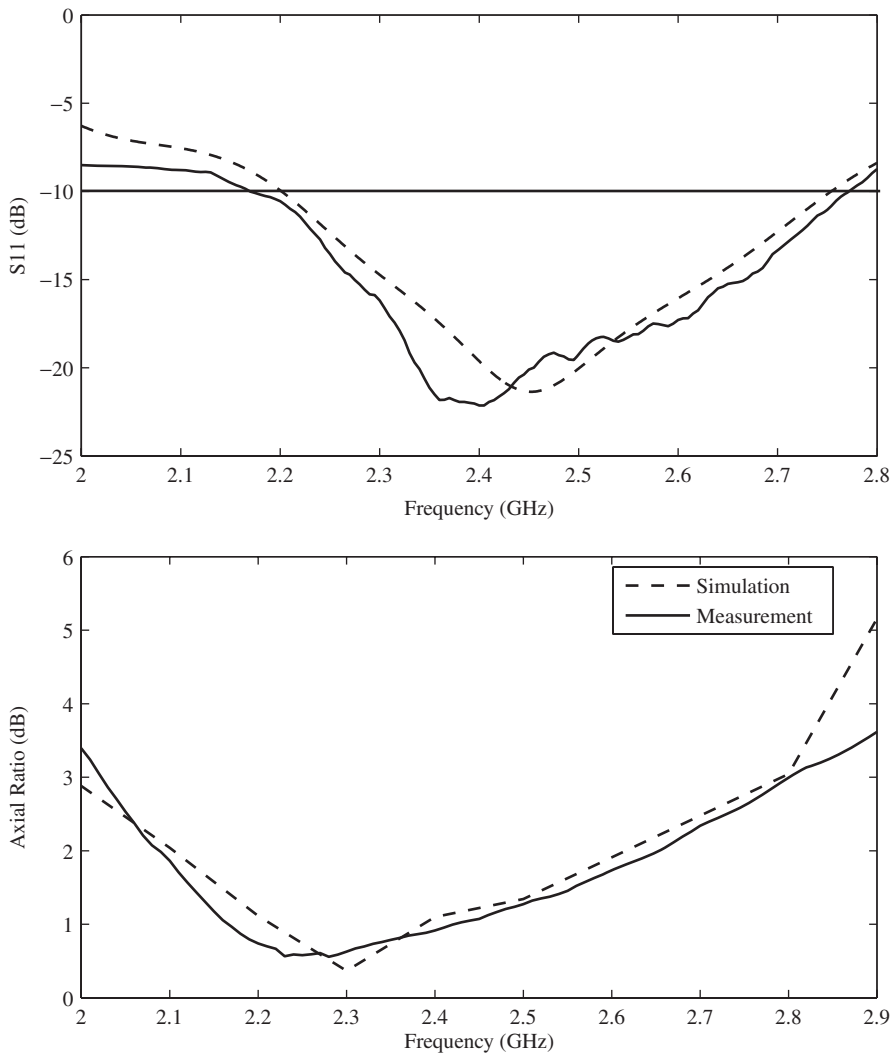
Many techniques that can be employed to design small CP antennas have been presented in this chapter. These methods include the use of modified shapes of radiators (patch, DRA, meander-lined helix), high permittivity dielectric materials, meta-materials, slot loading on the radiator or ground plane, shorting pins or shorting walls, and so on. To explain each technique, lots of examples of small CP antenna designs (patch, slot, helix, QHA, PQHA, slot, DRA) are illustrated and their results are discussed. It is shown that the antenna size reduction is often achieved at the expense of degraded radiation performance. Therefore, the suitable miniaturization method needs to be chosen based on the requirement of the application and it is necessary to optimize the radiation performance, especially the axial ratio of



**Figure 2.47** The side view and top view of the DRA with an under laid quadrature coupler [64]. Reproduced with permission of © 2011 IEEE

the antenna at the desired operation frequency band. Besides the methods presented in this chapter, there are other techniques that can also be implemented to small CP antenna design and in some cases, it is also possible to combine several different antenna miniaturization techniques to design one small CP antenna.

To provide a general idea of small CP antennas, Table 2.3 summarizes the performance of some small CP antennas. Despite the narrow bandwidth, the high-permittivity material loaded QHA in [38] seems to be one of the smallest CP antennas amongst antennas listed in Table 2.3. The meta-material loaded patch antenna in [21] is also quite small compared to other CP antennas in the table. The DRA antenna in [59] is small while achieving broadband performance (10% 3-dB AR bandwidth). Due to space limitation, only a limited number of small CP antennas are included in this table.



**Figure 2.48** The simulated and measured S11 and AR of the antenna presented in Figure 2.47 [64]. Reproduced with permission of © 2011 IEEE

**Table 2.3** Comparison of the main antenna parameters of the small CP antennas

| Ref. | Ant. Type            | $f_0$ (GHz) | size                                                 | substrate $\epsilon_r$ | 10-dB Return Loss Bandwidth (%) | 3-dB AR Bandwidth (%) |
|------|----------------------|-------------|------------------------------------------------------|------------------------|---------------------------------|-----------------------|
| [6]  | Microstrip patch     | 1.5         | $0.2\lambda \times 0.2\lambda \times 0.008\lambda$   | 4.4                    | 2.1                             | 0.8                   |
| [65] | Microstrip patch     | 1.6         | $0.17\lambda \times 0.17\lambda \times 0.008\lambda$ | 4.4                    | n.a.                            | 0.81                  |
| [13] | Microstrip patch     | 1.5         | $0.24\lambda \times 0.24\lambda \times 0.008\lambda$ | 4.4                    | 5                               | 5                     |
| [10] | Microstrip patch     | 1.5         | $0.33\lambda \times 0.33\lambda \times 0.008\lambda$ | 2.6                    | 3.4                             | 0.8                   |
| [14] | Microstrip patch     | 1.5         | $0.21\lambda \times 0.21\lambda \times 0.008\lambda$ | 4.4                    | 4.44                            | 1.4                   |
| [21] | MTM-patch            | 1.29        | $0.1\lambda \times 0.1\lambda \times 0.043\lambda$   | 9.8                    | 0.6                             | –                     |
| [24] | Multi-feed patch     | 0.915       | $0.29 \times 0.29\lambda \times 0.005\lambda$        | 4.4                    | 46                              | 8.7                   |
| [32] | Helix                | 12          | $0.32\lambda \times 0.32\lambda \times 0.19\lambda$  | 1                      | –                               | 12                    |
| [38] | QHA                  | 1.5         | $0.07\lambda \times 0.04\lambda \times 0.04\lambda$  | 40                     | 0.8                             | 0.5                   |
| [40] | PQHA                 | 2           | $0.09\lambda \times 0.09\lambda \times 0.26\lambda$  | 1.0                    | 10                              | 10                    |
| [45] | Slot antenna         | 1.5         | $0.2\lambda \times 0.2\lambda \times 0.008\lambda$   | 4.4                    | 12.4                            | 4.3                   |
| [51] | CPW-fed slot antenna | 1.42        | $0.33\lambda \times 0.33\lambda \times 0.008\lambda$ | 4.4                    | 17.1                            | 12.4                  |
| [59] | DRA                  | 2.7         | $0.16\lambda \times 0.08\lambda \times 0.14\lambda$  | 12                     | 33                              | 10                    |
| [66] | DRA                  | 1.5         | $0.17\lambda \times 0.17\lambda \times 0.06\lambda$  | 30                     | 5.4                             | 1.95                  |

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# 3

## Broadband Circularly Polarized Antennas

### 3.1 Introduction

The increasing demands for more capacity and higher data rate in wireless systems have led to the development of broadband CP antennas. During recent decades, a variety of broadband CP antennas have been proposed for applications in mobile satellite communications, WLAN, DBS, RFID, GNSS, space communications and wireless power transmission systems.

This chapter reviews different techniques of designing broadband CP antennas, including broadband CP microstrip patch antennas, broadband CP helix antennas, broadband QHAs and PQHAs, planar spiral antennas, broadband CP slot antennas, broadband CP DRAs, broadband CP loop antennas, and so on. Many antenna design examples are explained and their performances are discussed. Advantages and disadvantages of each antenna design are illustrated. A table summarizing the performance of some typical broadband CP antennas is presented at the end.

### 3.2 Broadband CP Microstrip Patch Antennas

#### 3.2.1 Broadband Single-Feed CP Patch Antennas

##### 3.2.1.1 Thick Air Substrate

As discussed in Chapter 1, a CP patch antenna can be realized by using a single- or multi-feed technique, and single-feed CP patch antennas have the advantages of simple structure and compact size. It is well-known that the traditional CP patch antenna has a narrow bandwidth due to the high quality factor  $Q$ . The closed-form expressions for the axial-ratio bandwidth (ARBW) and return-loss bandwidth (RLBW) of a single-feed CP patch antenna have been

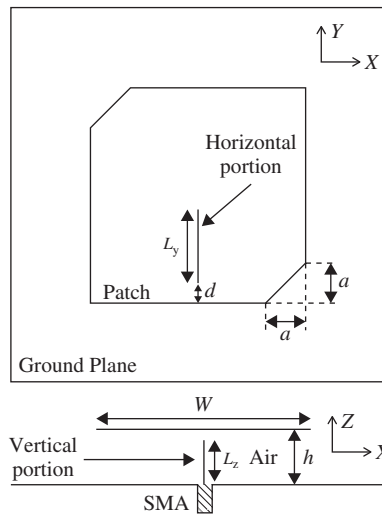
derived in [1] and given by

$$ARBW = \frac{AR - 1}{\sqrt{ARQ}} \quad (3.1)$$

$$RLBW = \frac{\sqrt{2(VSWR - 1)}}{Q} \quad (3.2)$$

The equations (3.1–3.2) show that the AR and return loss bandwidths are inversely proportional to the  $Q$ -factor of the antenna. It is known that the  $Q$ -factor is directly related to the dielectric constant and thickness of the substrate. A low  $Q$ -factor and broad bandwidth of microstrip patch antennas can be obtained by utilizing a thick air substrate. However, in this case, the long probe of probe-fed CP patch antennas may introduce a large inductance which deteriorates the impedance matching of the antenna. In order to solve this problem, several modifications to the feeds of broadband CP patch antennas have been proposed, such as an L-probe feed [2–4], a meandering strip feed [5], using a deformed ground plane [6,7] or an L-shaped patch [8].

Figure 3.1 presents the configuration of an L-probe fed patch antenna [2]. As shown, two corners are removed from a square patch and the patch is fed by an L-shaped probe at the middle of one edge of the patch [2]. Circular polarization is realized due to the excitation of two orthogonal modes with a  $90^\circ$  phase shift as a result of perturbations of the square patch. The L-shaped probe leads to a capacitance due to the electromagnetic coupling between the horizontal portion of the L-probe and the patch. This capacitance compensates for the inductance caused by the long probe, thus the impedance matching of the antenna is improved. In addition, the CP patch antenna with an L-shaped probe feed has more parameters such as the height and length of the probe, which make it easier to obtain the optimal performance of the antenna. Therefore, better impedance matching can be achieved by the L-probe fed

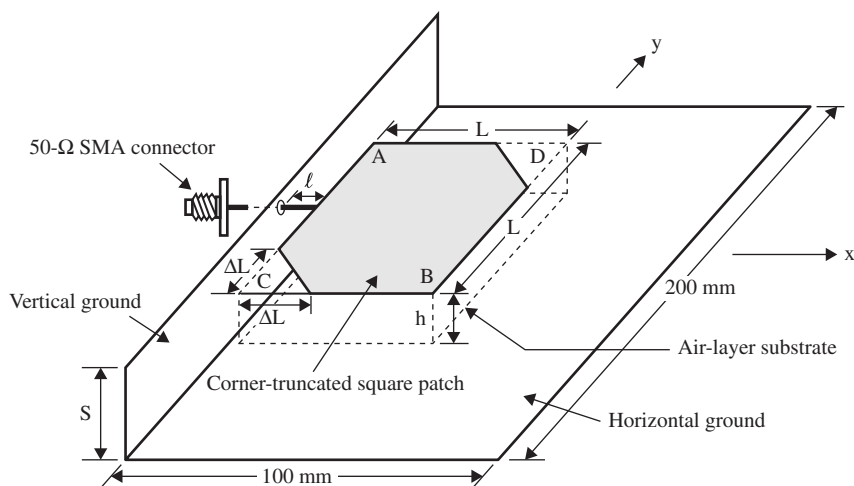


**Figure 3.1** Geometry of the L-probe fed CP patch antenna [2]. Reproduced with permission of © 2008 EMW

CP patch antenna compared to a traditional single-feed CP patch antenna as introduced in Chapter 1. The corner-truncated square patch has a side length of  $W = 28.6$  mm and is supported by an air substrate with a thickness of  $h = 15$  mm. The ground plane has a size of  $150 \times 150$  mm<sup>2</sup>. The L-shaped probe has a horizontal length of  $L_y = 7.5$  mm and a vertical length of  $L_z = 10.5$  mm. The optimized values of other parameters for achieving maximum bandwidth are:  $d = 3$  mm and  $a = 14$  mm. The measured impedance bandwidth ( $|S_{11}| \leq -10$  dB) is 25.3% (3.62–4.67 GHz) and 3-dB AR bandwidth is 16.6% (4.15–4.9 GHz).

In [4], the horizontal arm of the probe is realized by means of a printed monopole which is diagonally coupled to a small H-shaped copper plate for achieving circular polarization. The CP antenna in [4] can obtain a wide bandwidth of 19.4%. More recently, a 3D meandering strip feed is proposed in [5] and the antenna achieves an impedance bandwidth ( $VSWR \leq 2$ ) of 25.2% (2.18–2.81 GHz) and a 3-dB AR bandwidth of 22.4% (2.18–2.73 GHz). The antenna in [5] is composed of a square patch, a 3D meandering strip, a simple network for impedance matching and a square ground plane. Two orthogonal modes of the patch are excited by the 3D meandering strip, one end of which is connected to the microstrip line. The microstrip line consists of two parts: the wide part is fed by a 50  $\Omega$  SMA connector and the narrow part is connected to the ground plane through a shorting via.

The use of an L-shaped probe feed is effective in tackling the problem of impedance mismatching due to a large inductance caused by the long probe. An alternative technique is to use an L-shaped ground plane. As shown in Figure 3.2, the antenna consists of a corner-truncated square patch, an L-shaped ground plane and a probe feed. The square radiating patch has a side length of  $L$ , and a pair of opposite corners with a side length of  $\Delta L$  cut from the patch for achieving circular polarization. The corner-truncated square patch is placed above an L-shaped ground plane which is formed by a horizontal ground ( $200 \times 100$  mm) and a vertical ground ( $200$  mm  $\times$   $s$ ). Through a via hole in the vertical ground, the antenna is excited by a probe feed oriented in the same plane as the patch. In this case, the length of

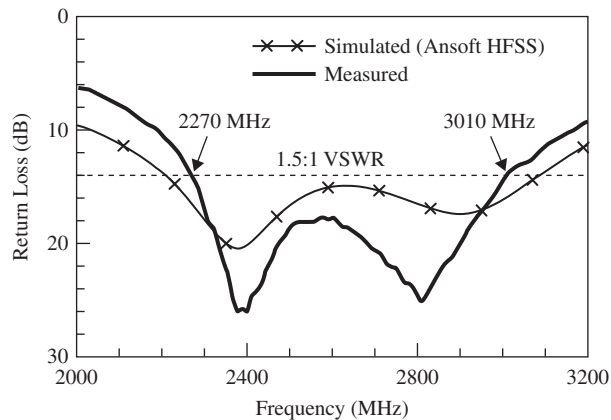


**Figure 3.2** Configuration of a CP patch antenna with an L-shaped ground plane [6]. Reproduced with permission of © 2003 IEEE

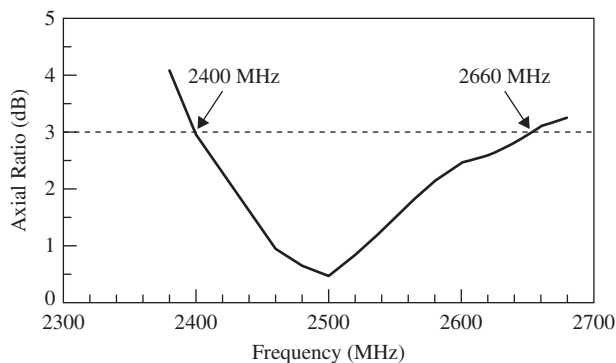
the probe-pin ( $l$ ) can be reduced, thus the problem of impedance mismatching caused by a large probe inductance is avoided. A broad impedance bandwidth can be obtained by using a thick air substrate and the L-shaped ground plane. It is shown in [6] that, in order to achieve good CP performance over a broadband bandwidth, the width of the vertical ground should be larger than two times the side length of the square patch. The dimensions for achieving a wide bandwidth are:  $L = 43$  mm,  $s = 23$  mm,  $\Delta L = 3.1$  mm,  $l = 3.5$  mm and  $h = 18$  mm.

The simulated and measured return loss results for the CP antenna are illustrated in Figure 3.3. Good agreement between the simulation and measurement can be observed. The measured impedance bandwidth (VSWR  $\leq 1.5$ ) is found to be 30% (2.27–3.01 GHz). The obtained 3-dB AR bandwidth in the broadside direction reaches 10.4% (2.4–2.66 GHz), as shown in Figure 3.4.

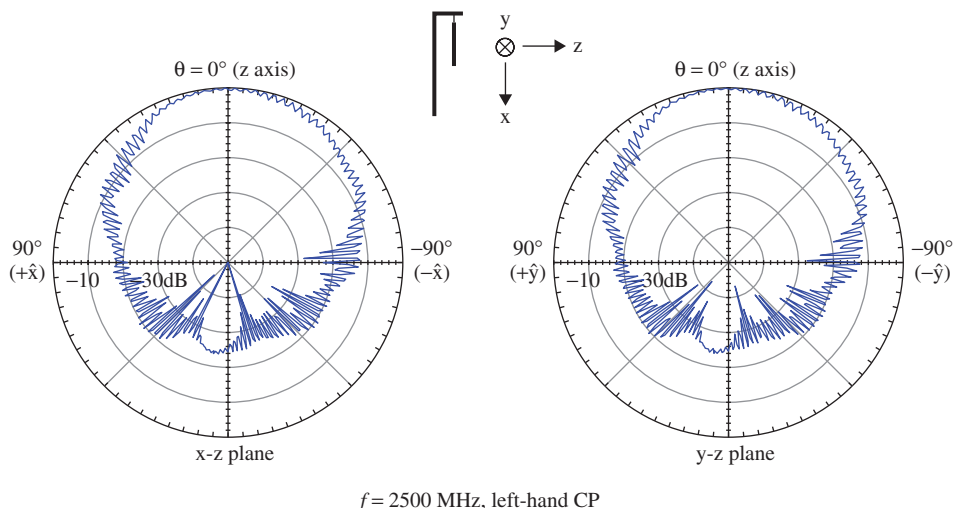
Measured spinning linear radiation patterns in two principal planes at 2500 MHz are shown in Figure 3.5. Slight asymmetry of the radiation patterns may be caused by the presence of the vertical ground plane. The backward radiation is less than  $-20$  dB compared to the main lobe in the forward radiation. It is also worthwhile mentioning that the larger



**Figure 3.3** Simulated and measured return loss for the CP antenna [6]. Reproduced with permission of © 2003 IEEE



**Figure 3.4** Measured AR for the CP antenna [6]. Reproduced with permission of © 2003 IEEE



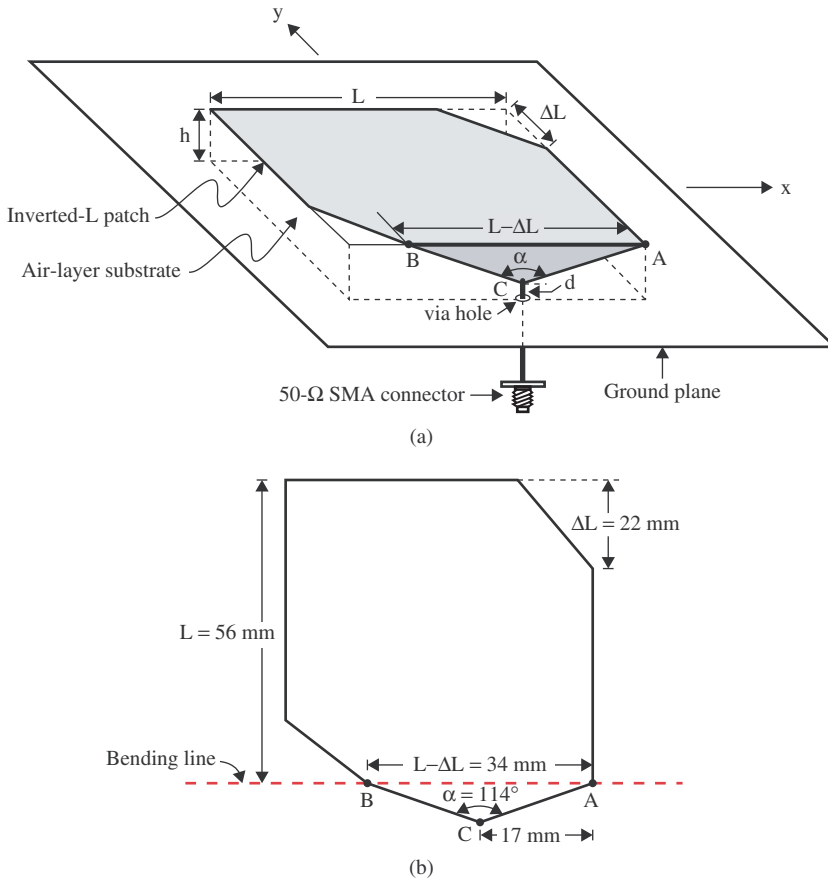
**Figure 3.5** Measured spinning linear radiation patterns at 2500 MHz in two principal planes for the CP antenna [6]. Reproduced with permission of © 2003 IEEE

the ripples in the radiation pattern, the larger the cross polarization levels. Hence, the cross polarization is minimum in the  $+z$  direction and increases away from the  $z$  axis. The measured peak gain is found to be about 9.2 dBi and the average gain level is around 8.5 dBi. The radiation efficiency is estimated to be around 80% by comparing the measured antenna gain and directivity.

Another technique for designing a broadband CP patch antenna with a thick air substrate is to use an L-shaped patch, as proposed in [8]. Figure 3.6 presents the geometry of the inverted-L patch antenna which has a thick air substrate and a short probe. The inverted-L patch is formed by a horizontal portion of a square patch and a vertical portion of an isosceles triangular shape. It is mounted above a ground plane with a distance of  $h$  and a size of  $150 \times 150 \text{ mm}^2$ . The thickness  $h$  is usually selected to be about 10% of the wavelength of the centre operating frequency [8]. Circular polarization is obtained by cutting a pair of opposite corners on the square radiating patch of the inverted-L patch. Good impedance matching over a wide frequency range can be obtained by using the isosceles triangular vertical portion and a probe feed with a short probe pin. The optimal side length of the truncated corners ( $\Delta L$ ) is about 40% of the side length of the square patch ( $L$ ). The optimized values of the dimensions are shown in Figure 3.6.

The measured return loss and AR results of the prototype are shown in Figures 3.7 and 3.8, respectively. As observed, the impedance bandwidth defined by  $\text{VSWR} \leq 1.5$  is about 19.5% (2.272–2.747 GHz) and the obtained 3-dB AR bandwidth reaches 7.0% (2.37–2.54 GHz). It is also noted that, the bandwidth of the antenna covers the 2.4-GHz WLAN band (2.4–2.484 GHz).

Measured spinning linear radiation patterns in two principal planes at 2.442 GHz are shown in Figure 3.9. Slight asymmetry of the radiation pattern can be observed, which is mainly caused by the presence of the isosceles triangular vertical part of the inverted-L



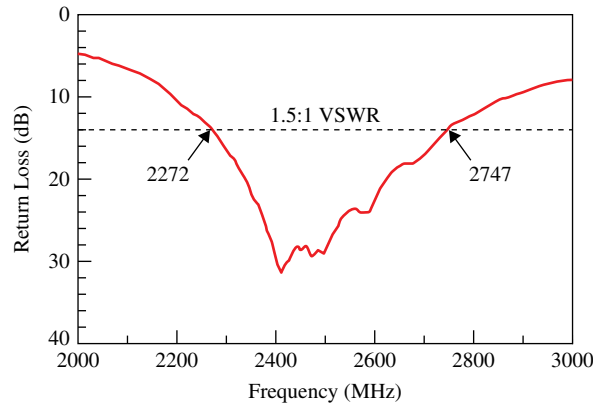
**Figure 3.6** Geometry of a broadband CP patch antenna using an L-shaped patch [8].  $d = 1$  mm,  $h = 12$  mm. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

patch. Good CP radiation at boresight is obtained. The average gain level is around 8.0 dBi within the obtained CP bandwidth.

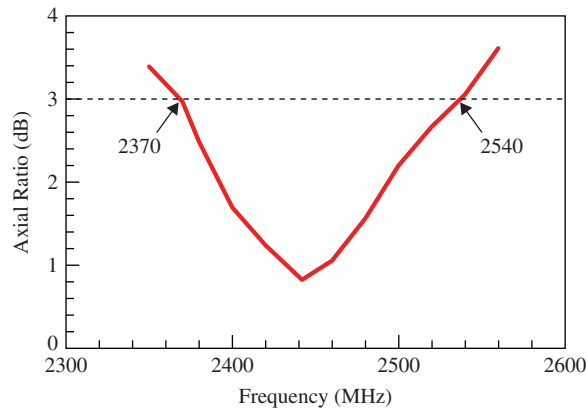
The broadband CP patch antennas discussed so far employ a single probe feed and a thick air substrate. It is also possible to employ other feeding techniques, such as slot coupling, electromagnetic coupling, and so on. Basic configurations of slot-coupled CP patch antennas have already been discussed in Chapter 1. If slot coupling is employed for patch with a thick air substrate, it will not have the large inductance problem caused by the long probe as in the case of probe-fed patch antennas. However, the coupling slot will need to be sufficiently long so that there is strong coupling between the patch and the slot. This will lead to high backward radiation and a low antenna gain.

### 3.2.1.2 Slot Loading

The broadband CP antennas utilizing a thick air substrate are bulky and thus not suitable for compact-size electronic systems. An alternative method for achieving a single-feed

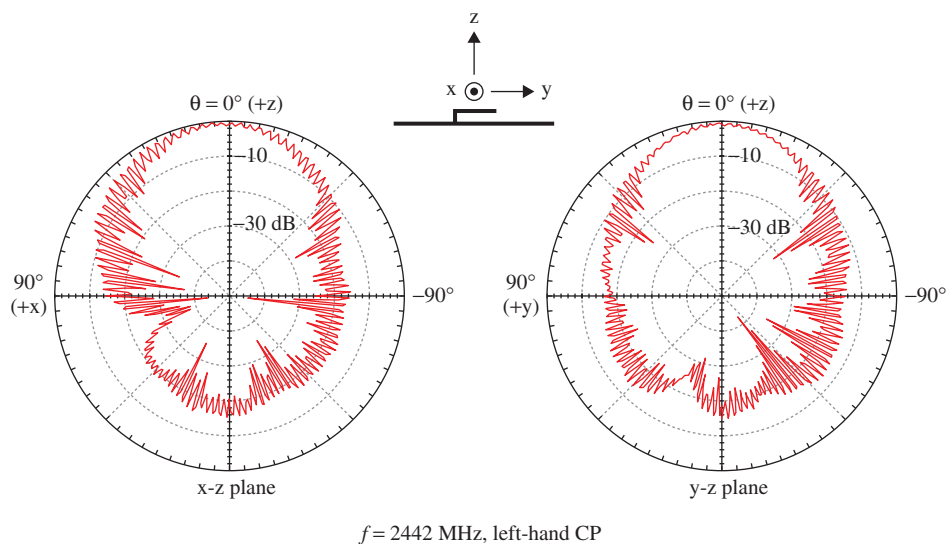


**Figure 3.7** Measured return loss for the inverted-L patch antenna [8]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

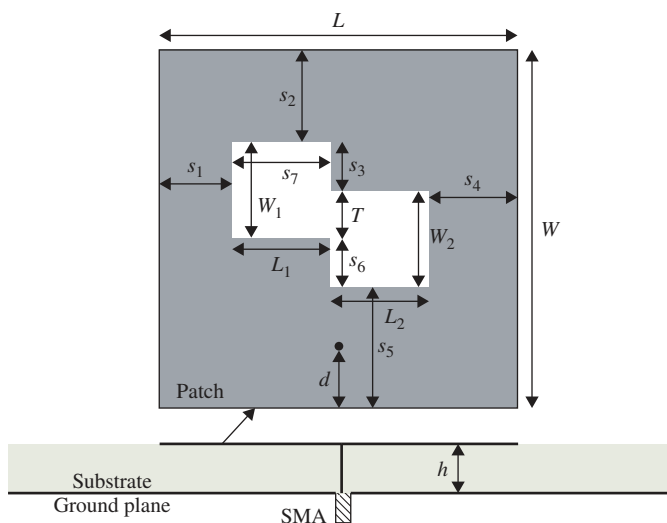


**Figure 3.8** Measured AR for the inverted-L patch antenna [8]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

broadband CP patch antenna is the use of slot loading. Figure 3.10 presents the configuration of a slot-loaded patch antenna. The probe-fed patch with a width of  $W$  and a length of  $L$  is printed on a substrate with a thickness of  $h$ . Two connected slots are inserted in the square patch. The physical dimensions of the slots are  $L_1$ ,  $W_1$ ,  $L_2$ ,  $W_2$  and  $T$ , as illustrated in Figure 3.10. When connected to each other, the two slots translate into two distinct resonant frequencies close to each other, leading to a wide frequency band. Circular polarization can be obtained when selecting proper values of the parameters for the slots. Two cases with different substrates (Rohacell foam and Duroid 5880) have been studied in this design. The optimized parameters in the case of using Rohacell with a relative permittivity of  $\epsilon_r = 1$  are:  $L = 21$  mm,  $W = 21.2$  mm,  $s_1 = 4.5$  mm,  $s_2 = 5.6$  mm,  $s_3 = 3.5$  mm,  $s_4 = 4.5$  mm,  $s_5 = 5.6$  mm,  $d = 3.6$  mm,  $s_6 = 3$  mm,  $s_7 = 6.5$  mm and  $h = 3.0$  mm. The simulated and measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) are 11.9% (5.14–5.79 GHz) and 11.8%



**Figure 3.9** Measured spinning linear radiation patterns at 2442 MHz in two principal planes for the inverted-L patch antenna [8]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.



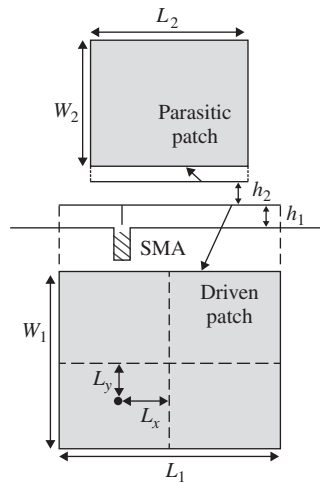
**Figure 3.10** Geometry of a slot-loaded patch antenna [9]. Reproduced with permission of © 2005 John Wiley & Sons, Inc.

(5.17–5.82 GHz), respectively, which are much broader than the impedance bandwidth of the same antenna without slot loading (4.2%).

When the Duroid 5880 substrate ( $\epsilon_r = 2.2$ ) is applied in the design, the optimized parameters of the antenna are:  $L = 14.8$  mm,  $W = 14.8$  mm,  $s_1 = 2.95$  mm,  $s_2 = 3.65$  mm,  $s_3 = 2.45$  mm,  $s_4 = 2.95$  mm,  $s_5 = 3.65$  mm,  $d = 2.84$  mm,  $s_6 = 2.1$  mm,  $s_7 = 5.0$  mm and  $h = 3.175$  mm. It is noticed that, the antenna size is significantly reduced when Duroid 5880 substrate is used. The measured operating frequency of the antenna is decreased by about 90 MHz compared to the simulated result, which is probably owing to the tolerance limitation of the dielectric material. The measured impedance bandwidth ( $|S_{11}| \leq -10$  dB) is 11.6%. It is also noted that, when using Rohacell foam, the AR is less than 3 dB only within a small frequency range, while the 3-dB AR bandwidth can reach 8% when the Duroid 5880 substrate is utilized.

### 3.2.1.3 Stacked Patches with a Single Feed

The technique of stacked patches has been employed to achieve broadband CP patch antennas [10–19]. It does not require a thick air substrate as that for antennas in Section 3.2.1.1. Figure 3.11 presents the geometry of a CP stacked-patch antenna. The antenna consists of two patches suspended above a ground plane. Both patches are supported by two layers of foam ( $\epsilon_r \approx 1.05$ ) with thicknesses of  $h_1$  and  $h_2$ , respectively. A perpendicular probe feeds the bottom patch which is denoted as a driven patch, whereas the top patch behaves as a parasitic patch. Circular polarization is generated by diagonally feeding ‘nearly square’ patch. A broadband performance is achieved due to the use of two patches with slightly different dimensions: the two patches can resonate at different frequencies close to each other. In order to achieve minimal AR and return loss over the frequency band of interest, the optimized dimensions of the antenna in [15] are:  $L_1 = 78.1$  mm,  $W_1 = 64.8$  mm,  $L_2 = 64.6$  mm,  $W_2 = 60.8$  mm,  $L_x = 30$  mm,  $L_y = 12.8$  mm,  $h_1 = 6.4$  mm and  $h_2 = 7$  mm. The obtained impedance bandwidths defined by  $VSWR \leq 2$  and 2.5-dB AR bandwidth are 20% and 13.5%, respectively.

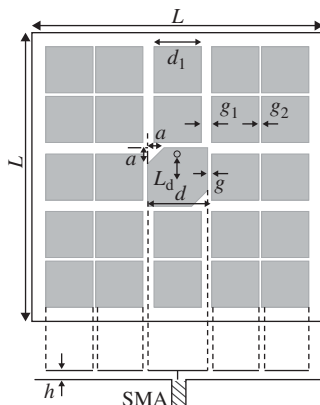


**Figure 3.11** Geometry of the single-feed CP stacked-patch antenna

In practice, it is difficult to fabricate patches on the foam substrates accurately. Thus, usually the patches are printed on different dielectric substrates and a foam substrate is inserted between these layers, as proposed in [10–14]. The stacked patches in [10,11] are printed on separate substrates with an air gap inserted between them to increase the bandwidth. In order to increase the antenna gain, the stacked-patch antenna in [12] has employed a quasi-planar surface-mounted short horn and achieved a 3-dB AR bandwidth of 15.5%, a 10-dB return loss bandwidth of 25% and an antenna gain of more than 11.5 dBi over the impedance bandwidth. Conversely, the driven patch in [13,14] is fabricated on the substrate in the middle layer between two air layers while the parasitic patch on the top is supported by an air substrate. The single-feed antenna in [13] consists of a lower notched circular patch with a capacitive feeding and an upper parasitic notched patch, and can achieve a CP bandwidth of 10% with conical radiation patterns. The antenna in [14] can provide an impedance bandwidth ( $VSWR \leq 2$ ) of 66.26% and a 3-dB AR bandwidth of 20.2%, respectively. Furthermore, in [16], the impedance bandwidth ( $VSWR \leq 1.5$ ) and 3-dB AR bandwidth of the antenna are improved to be 25.8% and 13.5% by using a meandered strip feed instead of a probe feed. The design in [17] has applied two parasitic patches, in which one is to improve the bandwidth and the other one acts as a director for increasing the antenna gain. Thus it can achieve both a broad bandwidth and a high antenna gain. Instead of using a probe feed, aperture-coupled feed with stacked patches can be used to increase the bandwidth, as discussed in [18]. More recently, the aperture-coupled stacked-patch antenna in [19] consists of four parasitic patches with each one being rotated by an angle of  $30^\circ$  relative to its adjacent patches for achieving circular polarization. The obtained CP bandwidth of the antenna in [19] can reach 33.6% (7.2–10.11 GHz).

### 3.2.1.4 Coplanar Parasitic Patches with a Single Feed

The technique of stacked patches requires the antenna to use a multi-layer structure. In some applications, it is desirable to use a single-layer structure for keeping a low profile. One technique of achieving single-layer broadband CP antennas is to use parasitic patches on the same plane as the driven patch. It is shown in [20] that multiple frequency bands can be obtained if the sizes of the driven patch and parasitic patches are equal. However, a broadband performance can be achieved by using the driven and parasitic patches with variable sizes. Figure 3.12 presents the configuration of a gap-coupled multi-resonator CP patch antenna which is printed on an FR4 substrate with a thickness of  $h = 1.6$  mm and a relative permittivity of  $\epsilon_r = 4.06$ . The antenna consists of a probe-fed square patch which is denoted as the driven patch and  $5 \times 5$  parasitic patches. Two corners are cut from the square driven patch for exciting CP radiation. All parasitic patches are square in shape and the side length of each parasitic patch ( $d_1$ ) is smaller than the side length of the driven patch ( $d$ ). This can help improve the AR performance. Two parameters  $g$  and  $g_1$  are introduced to denote different separations between elements. In most cases,  $g_2$  can be equal to  $g$ . If  $g_2$  is too large, little energy can be coupled to the parasitic patches. However, the antenna may have poor CP performance when  $g_2$  has a small value. In order to achieve a broad bandwidth, the optimized values of the parameters are:  $L = 50$  mm,  $L_d = 4.1$  mm,  $a = 3.7$  mm,  $d = 11$  mm,  $d_1 = 8$  mm,  $g = g_2 = 0.25$  mm and  $g_1 = 1.75$  mm. The diameter of the probe pin is 1.3 mm. The measured impedance bandwidth ( $VSWR \leq 2$ ) reaches 22% (5.44–6.787 GHz). It is



**Figure 3.12** Geometry of the gap-coupled multi-resonator CP antenna [20]. Reproduced with permission of © 2009 EMW

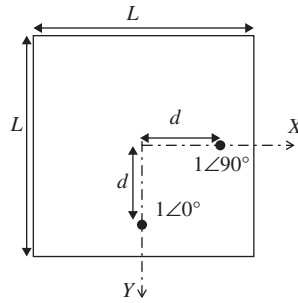
found that the measured 3-dB AR bandwidth is 8.1% (5.85–6.4 GHz). The antenna gain is about 9.3 dBic within the CP bandwidth.

### 3.2.2 Broadband Multi-Feed CP Patch Antennas

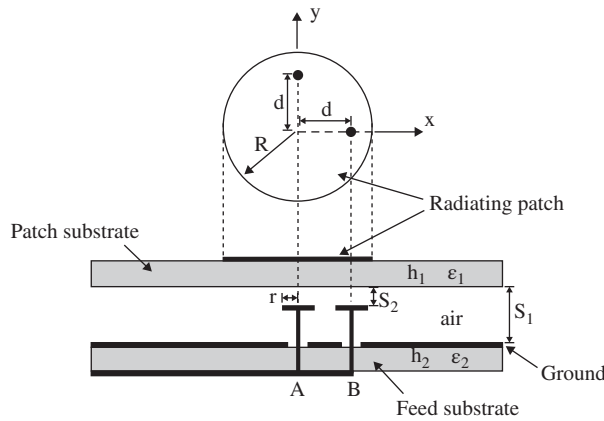
Many examples of broadband CP patch antennas using a single feed have been discussed in previous sections. Circular polarization can also be achieved by using two or more feeds. In the case of dual feeds, two orthogonal modes are excited by two feeds located at the centre of orthogonal edges of a square patch, and a  $90^\circ$  phase shift is required at two feeds, as shown in Figure 3.13. As two feeds are at the position of null electric field in the orthogonal mode, a high isolation between the two feeds can be obtained. Compared to the single-feed type, much wider impedance and AR bandwidths can be achieved by using the dual-feed technique. Multi-feed antenna designs can be realized by using an integrated feeding circuit including Wilkinson power divider [21–24], cross-slot coupling with an L-shaped feed [25], two-stub  $90^\circ$  hybrid coupler [26],  $120^\circ$  phase shifter [27], three-stub  $90^\circ$  hybrid coupler [28],  $90^\circ$  broadband balun [29], or external two-way  $0^\circ$  and  $90^\circ$  power divider [30]. The methods to increase the bandwidth of single-feed CP patch antennas can also be applied in the multi-feed CP patch antennas.

#### 3.2.2.1 Thick Air Substrate

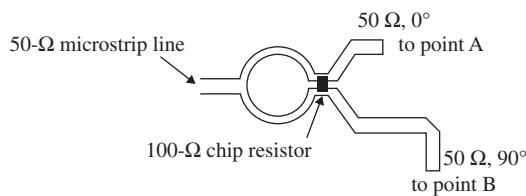
Figure 3.14 presents the geometry of a broadband CP antenna with two capacitively coupled feeds. The design of a Wilkinson power divider is also studied in [21]. It is printed on a grounded FR4 substrate with a relative permittivity of  $\epsilon_2 = 4.4$  and a thickness of 0.8 mm. The two output feedlines of the power divider have a length difference of one quarter-wavelength for generating a quadrature phase shift and are connected to two capacitor-plate-loaded feeds through the via holes at points A and B. High isolation between the two feeds can be obtained by applying a  $100\text{-}\Omega$  chip resistor in the power divider. The capacitor feeds loaded on the two feeds have the same radius of  $r$ . The radiating circular patch, also



**Figure 3.13** Configuration of a dual-feed CP patch antenna.



(a)



(b)

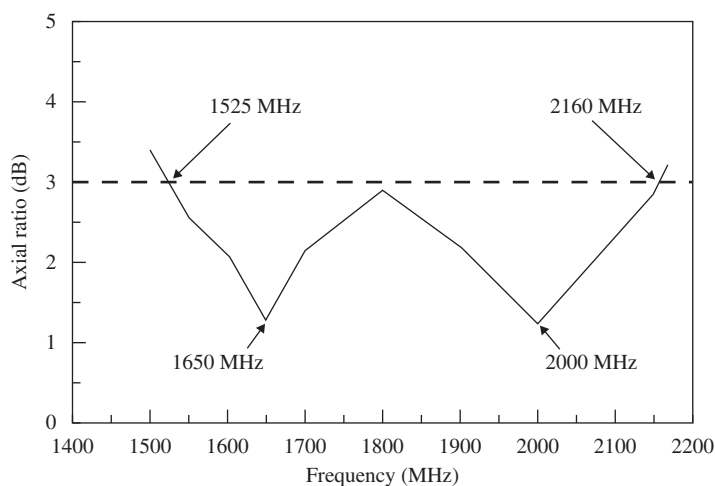
**Figure 3.14** Configuration of the CP microstrip antenna with dual capacitively coupled feeds [21]. Reproduced with permission of ©2001 IEEE

fabricated on an FR4 substrate with a thickness of 0.8 mm, is capacitively excited by the two feeds. Both feeds are orthogonally located and have the same distance of  $d$  away from the centre axis of the circular radiating patch. The patch substrate is mounted above the ground plane ( $100 \times 100 \text{ mm}^2$ ) of the feed substrate by non-conducting posts with a distance of  $S_1$ . The distance between the patch substrate and the two feeds is  $S_2$ . The Wilkinson power

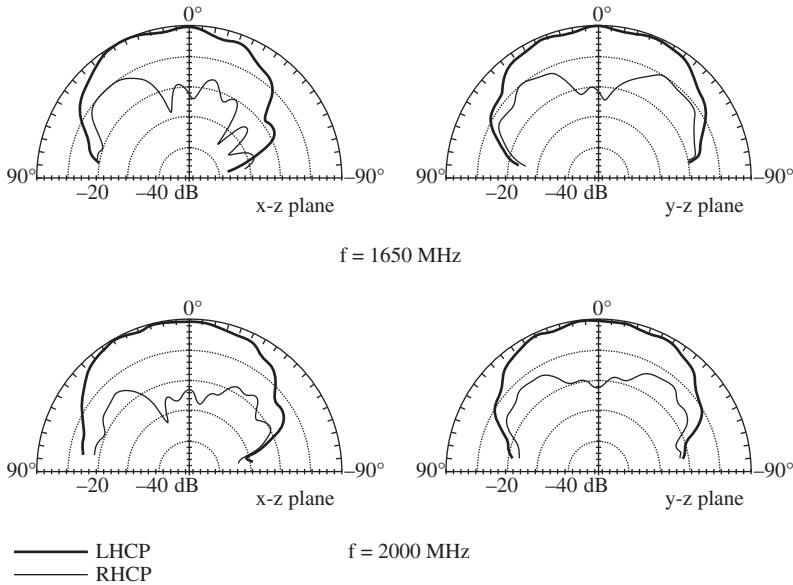
divider with two feedlines having different lengths provides two signals of equal amplitudes and a  $90^\circ$  phase shift as required for circular polarization. Moreover, a broadband CP bandwidth can be obtained when choosing large values of  $S_1$ . It is found that an optimal CP bandwidth can be reached when  $S_1$  is about  $0.08\lambda_0$  ( $\lambda_0$  is the free space wavelength at the centre operating frequency). The optimized parameters of the antenna for achieving an optimal bandwidth are:  $R = 28.75$  mm,  $S_1 = 12.8$  mm,  $r = 4.5$  mm,  $S_2 = 2.4$  mm and  $d = 22$  mm.

The obtained impedance bandwidth ( $VSWR \leq 1.5$ ) for the antenna integrated with the Wilkinson power divider is 986 MHz or 49% with respect to the centre operating frequency at 2.007 GHz. The measured 3-dB AR bandwidth is 35% (1.525–2.16 GHz), as shown in Figure 3.15. The measured radiation patterns in two principal planes at 1650 and 2000 MHz are shown in Figure 3.16. Good LHCP radiation has been obtained. The variation of the obtained antenna gain over the entire CP bandwidth is less than 3 dB and the peak antenna gain is found to be 7.2 dBi. The gain variation in 1.63–2.16 GHz band (about 28%) is less than 1 dB.

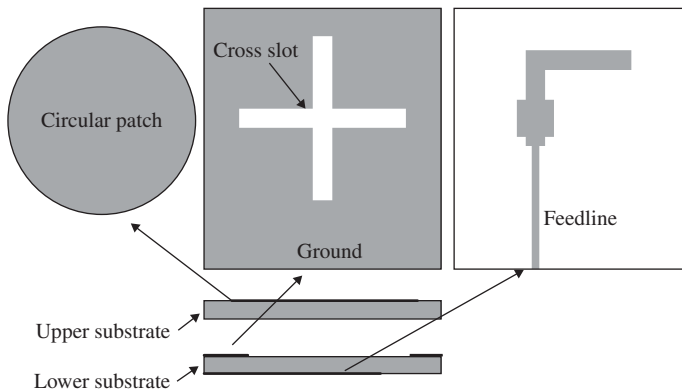
Other feed techniques such as slot coupling or electromagnetic coupling [22,25] can also be employed in multi-feed broadband CP antennas. Figure 3.17 shows a cross-slot-coupled patch antenna. By using a crossed slot to couple the electromagnetic energy from the microstrip feed to the radiating element, circular polarization can be easily obtained. The crossed slots are used to excite the two orthogonal modes in the antenna, with equal amplitudes and a  $90^\circ$  phase shift between them is obtained by using an L-shaped microstrip feed which has a length of a quarter-wavelength. To achieve a broadband operation, a circular microstrip patch is placed above an air layer with a thickness of 10 mm. It is electromagnetically coupled to the crossed slots in the ground plane, which is fed by an L-shaped microstrip feed below the ground plane. The proposed design achieves RHCP radiation at a centre frequency of 2 GHz. To ease the fabrication, an inverted structure is used, and a circular patch is printed on the Duroid 5880 substrate, which has a thickness of 0.79 mm and a relative permittivity of 2.2. This substrate behaves as a kind of protection



**Figure 3.15** Measured AR for the broadband CP antenna [21]. Reproduced with permission of © 2001 IEEE



**Figure 3.16** Measured radiation patterns in two principal planes at 1650 and 2000 MHz [21]. Reproduced with permission of © 2001 IEEE



**Figure 3.17** Geometry of a slot-coupled CP antenna [25]. Reproduced with permission of © 2007 IEEE

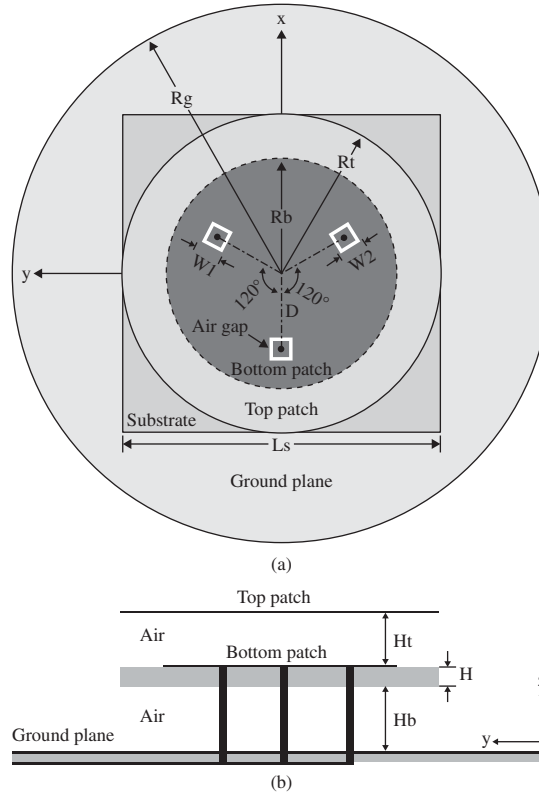
for the antenna against snow and fog. An L-shaped feed line is printed on the bottom of the lower Duroid 5880 substrate, which has a thickness of 0.79 mm and a relatively permittivity of 2.2. Crossed slots with unequal dimensions are etched in the ground plane. The substrate between the patch and ground plane determines the bandwidth of the proposed design. The antenna in [25] can achieve a 10-dB return loss bandwidth of 26% (1.76–2.29 GHz) and a 3-dB AR bandwidth of 9.6% (1.88–2.07 GHz). The bandwidth for gain greater than 6 dBi is 30% (1.86–2.27 GHz). In addition, in [23], an L-shaped probe and an aperture are

utilized to orthogonally excite the rectangular patch. The rectangular patch is supported by four plastic posts above the ground plane. The L-shaped probe is connected to a microstrip line through a via hole in the ground plane for exciting one edge of the patch. The aperture is used to feed the centre of the patch to provide the linear polarization orthogonal to that generated by the L-probe. The microstrip lines used to feed the L-probe and the aperture have the same length. In order to realize circular polarization, the output feedlines of a Wilkinson power divider are connected to the inputs of the dual-polarized patch antenna. The antenna using air as the dielectric substrate can achieve a 14-dB return loss bandwidth of 35% (1.49–2.12 GHz) and a 3-dB AR bandwidth of 20.4% (1.67–2.05 GHz). Furthermore, a CP conical-beam patch antenna is studied in [24]. The circular patch is supported by two plastic posts in the air. The hybrid feeding is implemented by using a pair of L-probes and apertures, which are connected by the feeding network to excite the circular patch in orthogonal directions. The feeding network is formed by three Wilkinson power dividers to provide signals with equal amplitudes and a  $90^\circ$  phase shift. The antenna in [24] can achieve an impedance bandwidth ( $VSWR \leq 1.5$ ) of 28.2% and a 3-dB AR bandwidth of 28.3%. Also, it has LHCP conical patterns with a 3-dB gain bandwidth of 32.4%.

### 3.2.2.2 Stacked Patches with Multiple Feeds

A CP stacked-patch antenna fed by a two-stub branch-line hybrid coupler is proposed in [26]. The antenna is formed by three square patches which are made by brass. The lowest one is a driven patch and the other two are parasitic patches. The hybrid coupler printed on an FR4 substrate ( $\epsilon_r = 4.4$ ,  $\tan \delta = 0.02$ , thickness = 0.8128 mm) has four ports: an input port, a 50- $\Omega$  loaded port and two other ports which are used to excite the driven patch. Such a feed network has a high isolation between the two feeding ports and less reflection to the input port as the reflected power is absorbed by the resistor. The feeding positions are located symmetrically with respect to the centre of the antenna with a same distance away from the edge of the driven patch. The measured 15-dB return loss bandwidth is from 800 to 980 MHz. The measured 3-dB AR bandwidth is from 820 to 1000 MHz. The AR is lower than 2.6 dB across 860–960 MHz. The measured gain is more than 6.5 dBic over 800–980 MHz.

Most CP antennas employ two orthogonal feeds. It is also possible to excite CP fields using three or four feeds. A stacked-patch antenna with three feeds is presented in Figure 3.18 [27]. The driven patch is excited by three metallic probes mounted via an FR4 substrate and an air gap. An air gap embedded in the bottom patch is equivalent to a serial capacitive component to compensate the inductance caused by the probe, thus improving the impedance matching. Three feeding positions are oriented in circumferential directions with the same distance from the patch centre. The upper patch is added to enhance the bandwidth and supported by a foam layer. Traditionally, circular polarization requires two orthogonal linear polarizations with a  $90^\circ$  phase shift. The feeding positions of the antenna in Figure 3.18 employ three non-orthogonal linear polarizations instead. The investigation in [27] has shown that CP radiation can be obtained when the phase difference between succeeding ports is  $120^\circ$ . The feed network is printed on the bottom layer of a dielectric substrate and presented in Figure 3.19. The meta-material transmission line at Port 2 gives the phase-advanced response to achieve a  $120^\circ$  difference compared to the transmission line at Port 3. It is realized by cascading two T-network with  $L = 9.5$  nH and  $C = 3.8$  pF. In addition, the length difference between the

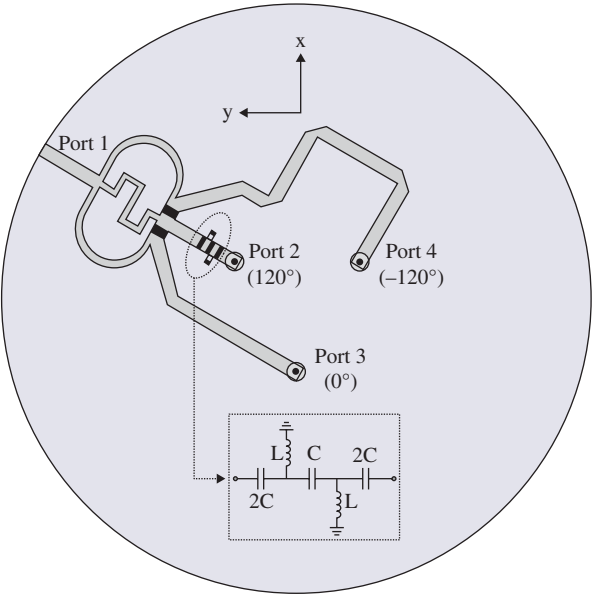


**Figure 3.18** Geometry of the CP stacked patch antenna with three feeds [27]. Reproduced with permission of © 2010 IEEE

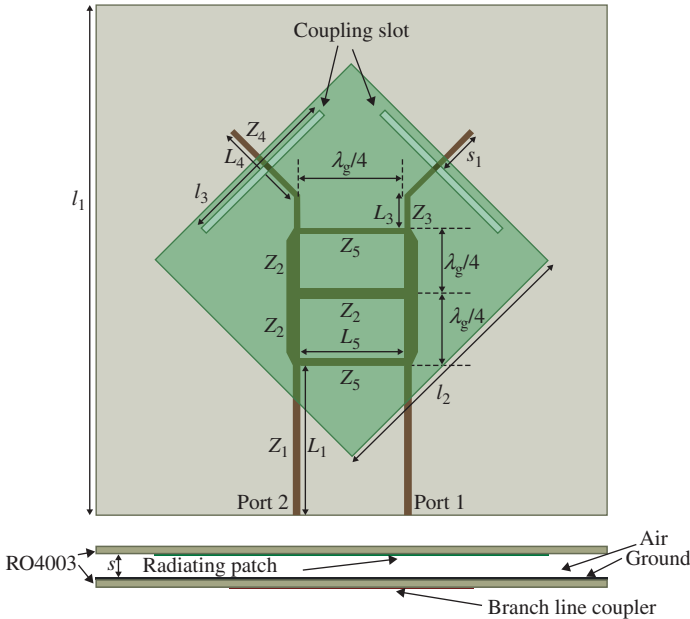
transmission lines for Port 3 and Port 4 is selected to be one third wavelength for providing another  $120^\circ$  phase difference. The antenna in [27] can achieve a 10-dB return loss bandwidth of 54.75% and a 3-dB AR bandwidth of 47.88%. The obtained 3-dB gain bandwidth can reach 40% with a peak gain of 8.8 dBic.

### 3.2.2.3 Broadband Coupler

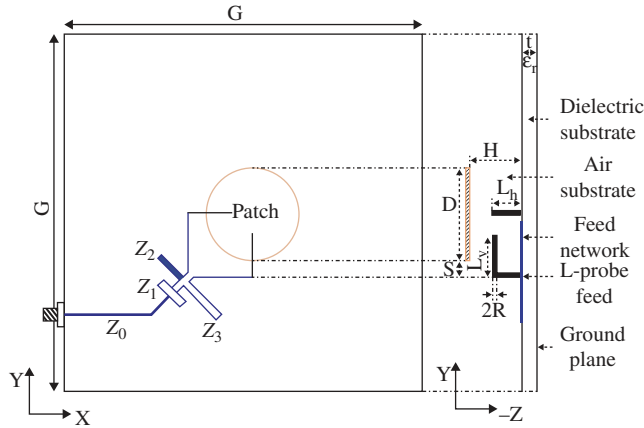
The bandwidth of CP patch antennas can be further increased by using a broadband coupler. Figure 3.20 presents the configuration of an aperture-coupled CP patch antenna fed by a three-stub hybrid coupler. Compared to a two-stub branch-line coupler which has a bandwidth of around 25%, the bandwidth of a three-stub branch-line coupler can go up to 40%. The hybrid coupler is printed on one side of the lower 0.813-mm-thick RO4003 substrate ( $\epsilon_r = 3.38$ ,  $\tan \delta = 0.002$ ). The characteristic impedance of the two wider parallel strips and the inner shunt strip is  $Z_2 = 35.4 \Omega$ . The outer shunt strip has a characteristic impedance of  $Z_5 = 120.8 \Omega$  and the characteristic impedance of all other lines is  $Z_1 = Z_3 = Z_4 = 50 \Omega$ . Port 2 is terminated with a  $50 \Omega$  resistor to absorb the reflected power. Two rectangular slots with a width of 1.88 mm are embedded in the square ground plane with a side length of  $l_1$  on the opposite side and separately perpendicular to the feed strip of the hybrid coupler.



**Figure 3.19** Geometry of the feed network [27]. Reproduced with permission of © 2010 IEEE



**Figure 3.20** Geometry of the aperture-coupled CP patch antenna using a three-stub branch-line coupler [28]. Reproduced with permission of © 2004 John Wiley & Sons, Inc.

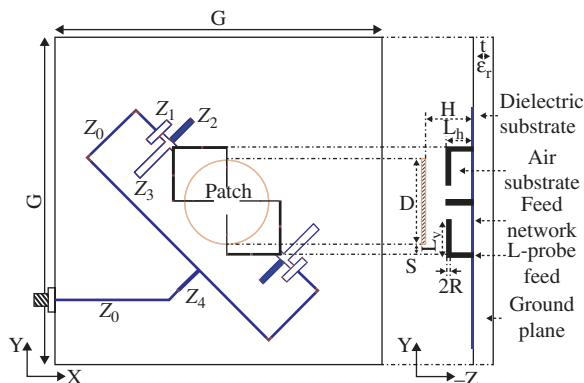


**Figure 3.21** Geometry of the dual L-probe fed circular patch antenna [29]. Reproduced with permission of © 2008 IEEE

The square radiating patch is printed on the bottom side of the upper substrate. The energy is coupled to the square radiating patch through the slots. Two dielectric substrates are connected by some plastic posts. The optimized values of the dimensions are:  $L_1 = 31.4$  mm,  $L_3 = 10$  mm,  $L_4 = 26.5$  mm,  $L_5 = 30.38$  mm,  $s_1 = 12.15$  mm,  $l_1 = 156$  mm,  $l_2 = 83$  mm and  $l_3 = 50$  mm. The obtained impedance bandwidth ( $VSWR \leq 2$ ) and 3-dB AR bandwidth are 32.3% (1.3–1.8 GHz) and 42.6% (1.22–1.88 GHz), respectively. The maximum antenna gain is found to be 8.8 dBic at 1.55 GHz.

In addition, a broadband dual L-probe fed CP antenna with a broadband balun is demonstrated in [29], as shown in Figure 3.21. The circular patch with a diameter of  $D$  is supported by an air substrate of thickness  $H$  above a grounded Rogers RO4003 dielectric substrate. The feed network is printed on the top layer of the substrate. Two L-shaped probes formed by its vertical and horizontal portions are orthogonally oriented and positioned at a distance of  $S$  away from the circumference of the patch. The L-probe feeds are soldered to the output lines of the feed network and excite the radiating patch by proximity coupling. The broadband feed network consists of a cascade of a 3-dB Wilkinson power divider for wideband impedance matching and balanced power splitting, and a novel broadband  $90^\circ$  Schiffman phase shifter for wideband consistent  $90^\circ$  phase shifting. The measured 10-dB return loss and 3-dB AR bandwidths of the dual L-probe fed CP antenna can reach 60.24% and 37.7%, respectively.

It is reported in [29] that the antenna bandwidth can be further increased by using four feeds instead of two feeds, as shown in Figure 3.22. The quadruple L-probe antenna has the same antenna parameters with the dual L-probe antenna. The feed network consists of a pair of the aforementioned  $90^\circ$  broadband baluns connected by a  $180^\circ$  transformer. The length difference of the microstrip branches is selected to be  $\lambda_g/2$  ( $\lambda_g$  is the guided wavelength at the centre operating frequency) for providing a  $180^\circ$  phase shift. The input transmission line is connected to the two microstrip branches through a quarter-wavelength transformer. In the case of four feeds, the 10-dB return loss and 3-dB AR bandwidths can reach 71.28% and 81.6%, respectively. However, the multi-feed CP antenna requires a complicated feed network which also occupies more space. For example, the four-feed CP



**Figure 3.22** Geometry of the quadruple L-probe fed circular patch antenna [29]. Reproduced with permission of © 2008 IEEE

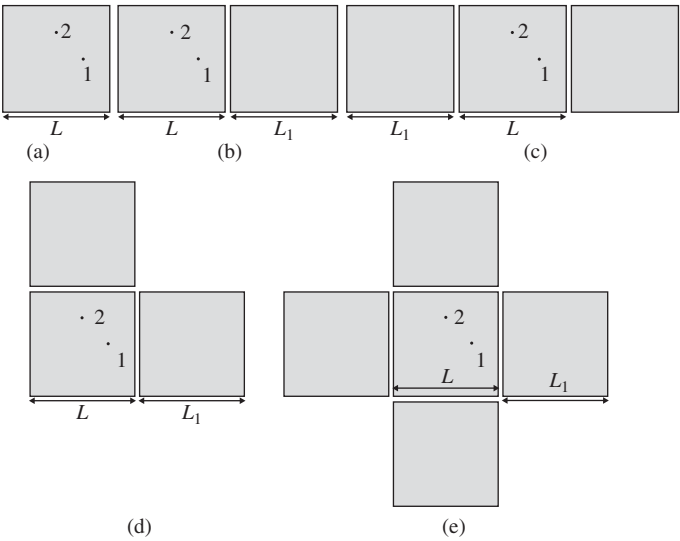
antenna requires a complicated feed network to provide  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  phase shift at four feeds, respectively.

### 3.2.2.4 Coplanar Parasitic Patches with Multiple Feeds

Similar to the case of single-feed gap-coupled multi-resonator CP antennas, the method of coplanar parasitic patches can also be employed in the multi-feed CP antenna [30]. Figure 3.23(a) presents the geometry of a patch antenna fed by two feeds. Circular polarization can be obtained by using an external hybrid coupler or power divider. To obtain a higher gain and larger bandwidth, one, two or four parasitic square patches are coupled to the driven patch on the same layer, as shown in Figures 3.23(b–e). A comparison of bandwidth, directivity and AR for these configurations is illustrated in Table 3.1. The directivity of CP antenna using a thick air substrate is stable over a wider frequency range than that of antenna using other dielectric substrates. It is noted that the AR value is less than 2 dB in the whole band for antenna using other dielectric substrates, whereas the corresponding value increases to be 4 dB in the upper band for antennas using a thick air substrate. By combining the techniques of using a thick air substrate and coplanar parasitic patches as shown in Figure 3.23(e), the CP patch antenna can achieve a bandwidth of 400 MHz (19%) with a directivity of 10.66 dB.

### 3.2.3 Other Broadband CP Patch Antennas

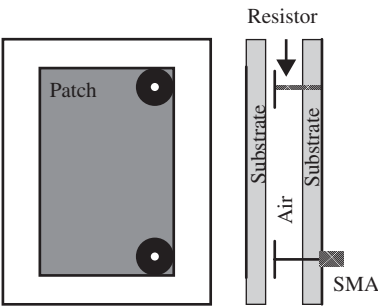
Recently, several typical bandwidth enhancement methods have been combined to increase the bandwidth of CP patch antennas. In [31], the bandwidth of the patch antenna is enhanced by inserting an air layer and using the resistor load. As depicted in Figure 3.24, the antenna is formed by a rectangular radiating patch, two identical substrates with an air layer inserted between them and a capacitively coupled feed. Another capacitive chip is connected at the opposite position with respect to the feed location through a resistor. Good CP performance can be obtained by tuning the feed position along the diagonal line of the rectangular patch. The bandwidth of the antenna can be enhanced with the loading of an external chip resistor,



**Figure 3.23** Various CP antennas with parasitic patches fed by two feeds [30]. Reproduced with permission of © 2000 John Wiley & Sons, Inc.

**Table 3.1** Comparison of various configurations in terms of Bandwidth, D and AR [30]. Reproduced with permission of © 2000 John Wiley & Sons, Inc.

| Configurations | $\epsilon_r = 2.33$ and $h = 1.6$ mm |                     |            | $\epsilon_r = 1$ and $h = 5$ mm |                     |            |
|----------------|--------------------------------------|---------------------|------------|---------------------------------|---------------------|------------|
|                | Bandwidth<br>(MHz)                   | Directivity<br>(dB) | AR<br>(dB) | Bandwidth<br>(MHz)              | Directivity<br>(dB) | AR<br>(dB) |
| (a)            | 30                                   | 7.36                | 0.3        | 110                             | 9.75                | 1.2        |
| (b)            | 60                                   | 8.78                | 2          | 220                             | 10.21               | 4          |
| (c)            | 70                                   | 9.02                | 2          | 280                             | 10.4                | 4          |
| (d)            | 90                                   | 9.31                | 2          | 350                             | 10.37               | 2          |
| (e)            | 100                                  | 10.79               | 0.8        | 400                             | 10.66               | 2          |



**Figure 3.24** Configuration of the broadband CP patch antenna with chip-resistor loading [31]. Reproduced with permission of © 2010 IEEE

as the quality factor of the microstrip patch is decreased. It is found that the 3-dB AR bandwidth of the antenna without the resistor loading is less than 7%. Both capacitive plates have the same radius.

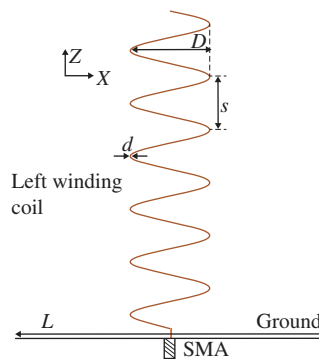
In [31], the optimum value of the resistor is  $30\ \Omega$ . The simulated impedance bandwidth ( $|S_{11}| \leq -10\text{ dB}$ ) is 1.06–2.54 GHz or 82% with respect to the centre frequency of 1.8 GHz. The measured impedance bandwidth ranges from 1.209 to 2.697 GHz or 78.7% at 1.891 GHz. The simulated 3-dB AR bandwidth is 36% (1.64–2.36 GHz) whereas the measured AR bandwidth is 26% (1.74–2.26 GHz).

### 3.3 Broadband Helix, QHAs and PQHAs

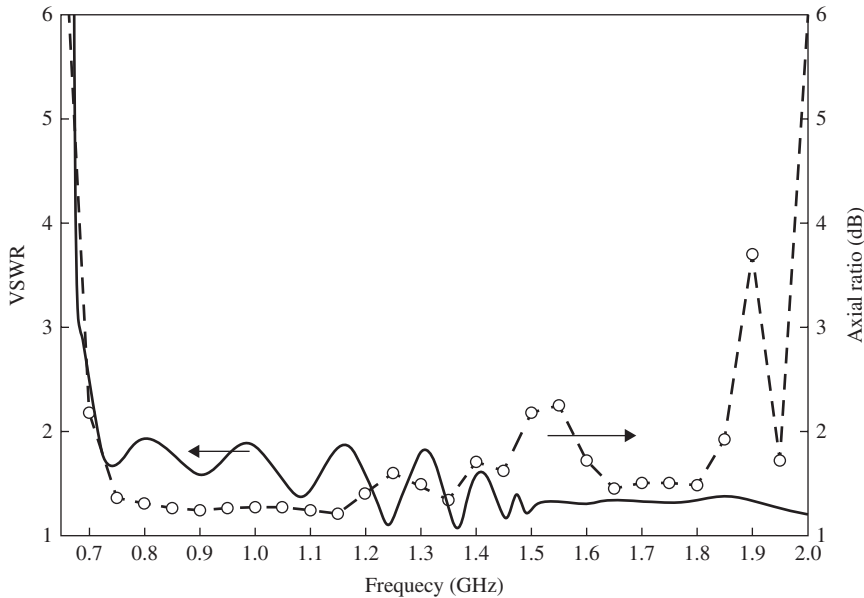
#### 3.3.1 Broadband Helix Antennas

The axial-mode helix antenna [32] invented by Kraus in 1946 has a good CP performance and high gain over a wide frequency range. However, the physical size can become rather large at low frequencies. In order to achieve low-profile helix antennas, the helix antenna of low pitch and a small number of turns or stub loaded helix has been studied in [33–35]. The low profile helix antennas can provide a CP bandwidth of more than 10%, though the height is significantly reduced. Basics of helix antennas have been discussed in Chapter 1. The helix antennas [36] include monofilar axial-mode helix antenna, monofilar normal-mode helix antenna, multifilar axial-mode helix antenna and multifilar normal-mode helix antenna. The QHA [37–39] consists of two pairs of bifilar helices and produces a cardioids-shaped CP pattern with a half-power beamwidth (HPBW) of around  $120^\circ$ . The QHA belongs to the class of resonant antennas and the bandwidth is narrow ( $\approx 4\%$ ). Moreover, the PQHA presented in [40] has the advantage of easy fabrication.

Figure 3.25 presents the geometry of a typical monofilar axial-mode helix antenna. The parameters to describe the helix [36] are:  $D$  = diameter of helix,  $C$  = circumference of helix =  $\pi D$ ,  $s$  = spacing between turns,  $d$  = diameter of helix conductor. As given in equation (1.18), Chapter 1, the antenna can radiate in the axial-mode for  $C$  between approximately  $0.75\ \lambda$  and  $1.25\ \lambda$ . The simulated results of VSWR and AR are shown in Figure 3.26. The



**Figure 3.25** Geometry of a typical six-turn monofilar helix antenna. ( $D = 9.5\text{ cm}$ ,  $s = 6.6\text{ cm}$ ,  $d = 1.7\text{ cm}$ ,  $L = 34\text{ cm}$ )



**Figure 3.26** Simulated VSWR and AR results versus frequency for the helix antenna

obtained impedance bandwidth for  $VSWR \leq 2$  ranges from 0.7 to over 2 GHz, whereas the 3-dB AR bandwidth is from 0.7 to 1.95 GHz. Symmetrical directional radiation patterns with maximum radiation in the axial direction can be observed in Figure 3.27. It is also found that the RHCP levels are 15 dB below the LHCP fields at 1 GHz.

### 3.3.2 Broadband PQHAs

As mentioned in the previous section, PQHA inherently has a narrower bandwidth compared to the monofilar axial-mode helix. This section will discuss different techniques of increasing the bandwidth of PQHA.

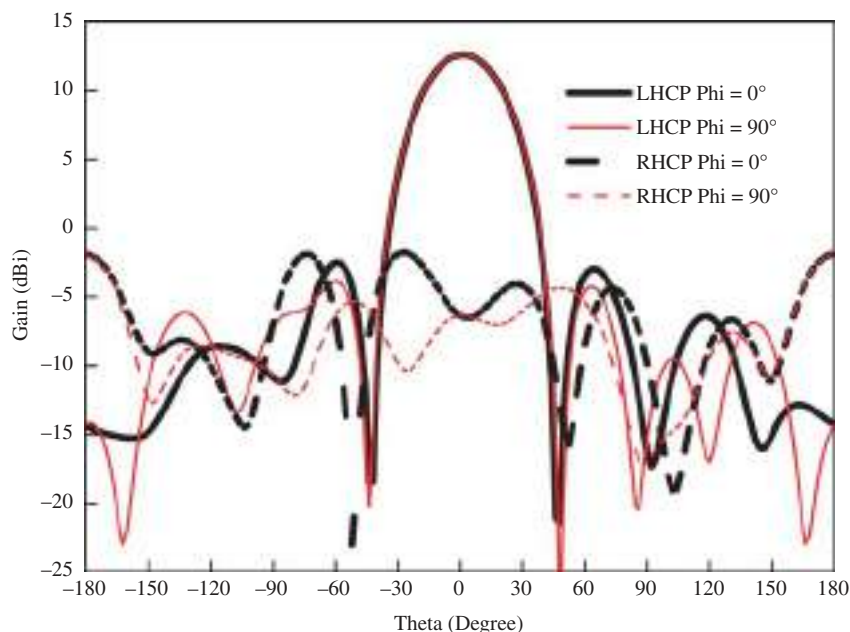
#### 3.3.2.1 Short PQHA with a Large Diameter

The relationship between the impedance bandwidth and the geometry of the PQHA is studied in [41], and given as:

$$BW \approx 19 \left( \frac{D}{H} \right) \% \quad (3.3)$$

where  $D$  and  $H$  are the diameter and height of the antenna, respectively. The relationship demonstrates that the bandwidth of the antenna can be increased by designing a short antenna with a large diameter.

Figure 3.28(a) presents the configuration of the low-profile PQHA which is mounted above a grounded FR4 substrate with a thickness of 0.8 mm, a relative permittivity of 4.4 and a size of  $100 \times 100 \text{ mm}^2$ . Its four radiating arms are printed on a flexible plastic

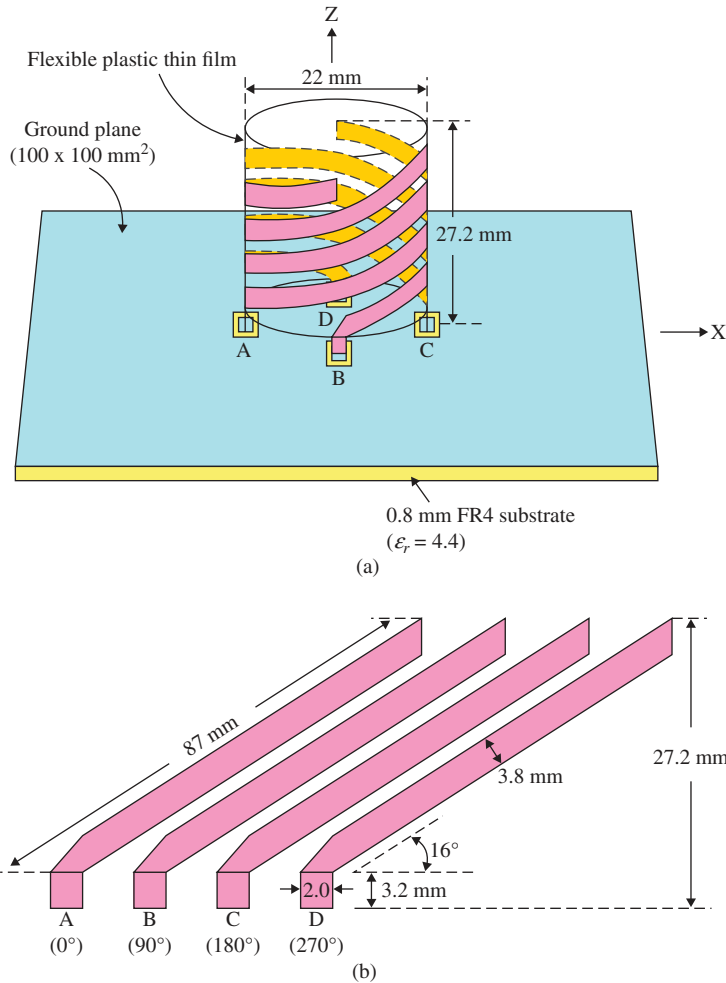


**Figure 3.27** Simulated co-polarized and cross-polarized radiation patterns at 1 GHz

thin sheet with the unrolled structure shown in Figure 3.28(b). The PQHA is designed to provide CP operation over a broadband frequency band covering 2310–2360 MHz, 2500–2655 MHz and 2655–2690 MHz for broadcasting satellite applications. The lengths of arms are selected to be 90.2 mm (around 70% of the wavelength at the desired operating frequencies). Each arm has a shorter vertical strip with a length of 3.2 mm for easy connection to the feed and the main radiating strip is 87 mm long. It is also worthwhile to mention that, at the connection to the shorter vertical strip, the main radiating strip with a width of 3.8 mm is chamfered to have the same width (2 mm) as that of the shorter vertical strip. The PQHA has a small pitch angle of  $16^\circ$  and its four strips are rolled into a helix shape on a foam cylinder with a diameter of 22 mm and a height of 27.2 mm (about 22% of the wavelength at the desired frequencies).

Figure 3.28(d) shows the layout of the Wilkinson power divider which can provide quadrature signals for the PQHA. In order to produce two output signals with equal output amplitudes but a  $90^\circ$  phase difference, two output feedlines of the power divider have a length difference of one quarter-wavelength. In this case, the signals at points A, B, C and D can have equal amplitudes and phase shifts of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$ , respectively.

The measured results of return loss and AR are shown in Figures 3.29 and 3.30. As observed, the impedance bandwidth ( $VSWR \leq 1.5$ ) is about 800 MHz at the centre frequency of 2.4 GHz. For the three frequency bands of interest, the measured return loss is more than 15 dB across 2310–2360 MHz, 20 dB across 2500–2655 MHz and 2655–2690 MHz. The obtained 3-dB AR bandwidth reaches about 670 MHz (2060–2730 MHz) or 27.9% at 2.4 GHz. The measured axial ratios are less than 1.2, 1.8 and 2.3 dB across the three frequency bands, respectively. Measured radiation patterns in two principal planes at different

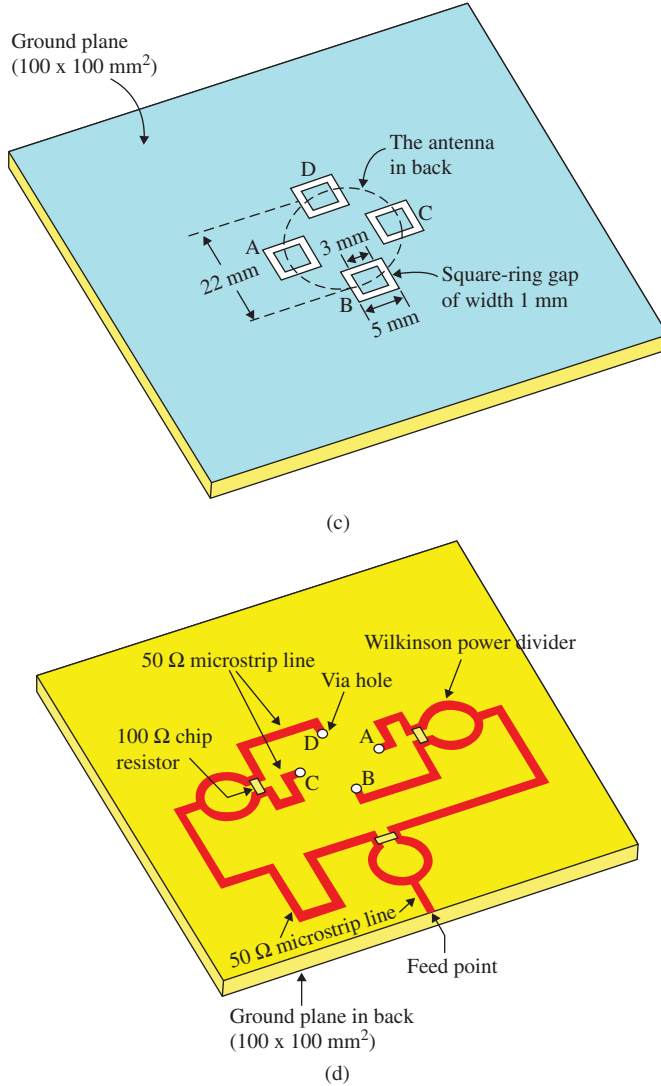


**Figure 3.28** (a) Configuration of the PQHA; (b) four radiating helix strips unrolled into a planar structure; (c) the upper side of the FR4 substrate; (d) the bottom side (feed network) of the FR4 substrate [40]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.

frequencies are illustrated in Figure 3.31, in which good CP radiation has been obtained. The measured antenna gain is about 3.2–4.3, 4.0–5.5 and 3.3–4.0 dBic over the three frequency bands, respectively.

### 3.3.2.2 Tapered PQHA

The bandwidth of the PQHA can be enlarged by using a tapered arm width, as shown in Figure 3.32. The four tapered arms can be printed on a thin substrate. The tapered arm is realized by changing the width of the arm along the antenna. In [42], the conventional PHQA with a constant arm width can operate in dual bands and the bandwidths are 7% and 5% at

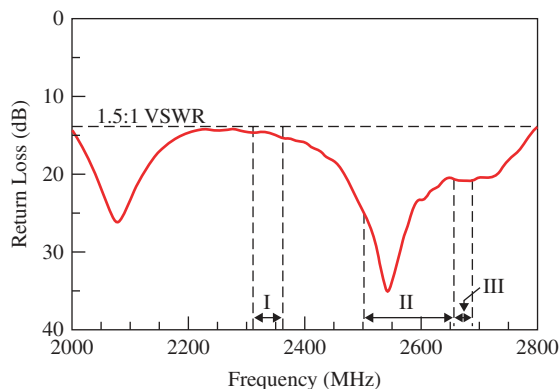


**Figure 3.28** (continued)

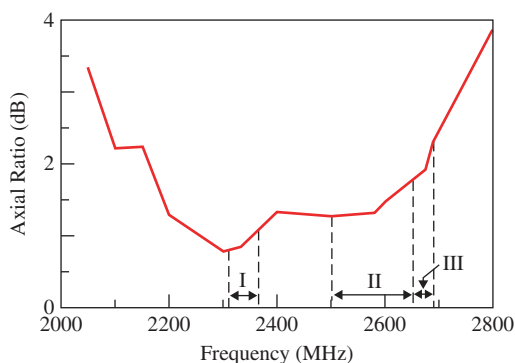
the L- and S-band, respectively. Compared to the conventional PQHA, the bandwidths of the tapered PQHA are increased to be 14% and 16% for  $VSWR \leq 2$ .

### 3.3.2.3 Conical PQHA

Another approach to increase the bandwidth of the PQHA is to utilize a conical arrangement [43]. The printed quadrifilar helix arms are folded in a conical shape. Figure 3.33 presents the geometry of the unwrapped conical PQHA. Each arm has a width of 4 mm and a length of  $0.778\lambda_0$  ( $\lambda_0$  is the wavelength at 2.16 GHz in this design). The conical PQHA is excited by a



**Figure 3.29** Measured return loss for the PQHA [40]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.



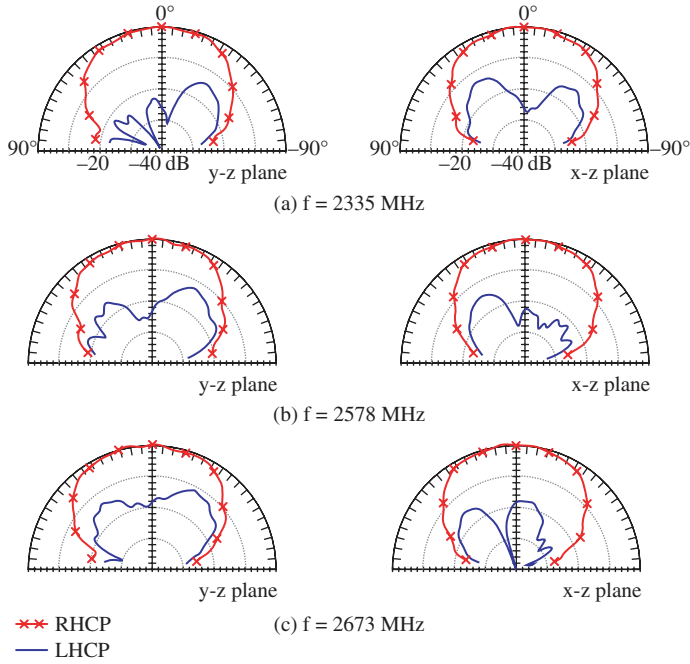
**Figure 3.30** Measured AR for the PQHA [40]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.

four-port feed network consisting of one  $180^\circ$  rat-race coupler and two  $90^\circ$  hybrid couplers which are printed on an FR4 substrate of thickness 1.575 mm and relative permittivity 4.4. The four output ports of the feed network have equal amplitudes but a  $90^\circ$  phase difference for achieving circular polarization. A 3D HFSS model for the integrated conical PQHA is presented in Figure 3.34.

The simulated and measured results of return loss are illustrated in Figure 3.35. The obtained 10-dB return loss bandwidth is 32.4%. In addition, the measured 3-dB AR bandwidth is 18.5% (1.9–2.3 GHz). It is also noted that, the conical PQHA possesses a wide HPBW, about  $150^\circ$  at 1.9 GHz and  $168^\circ$  at 2.3 GHz.

### 3.3.2.4 Parasitic Strips

The PQHA with parasitic helix strips can achieve a wide bandwidth, as the parasitic helix strips can produce another resonant frequency. The mutual coupling effect between the conductor and the parasitic strip can improve the impedance matching. The ends of the

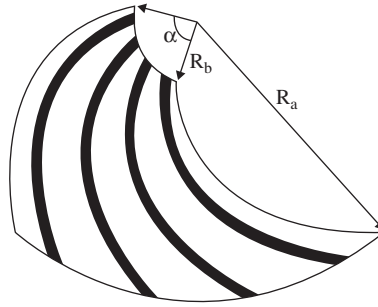


**Figure 3.31** Measured radiation patterns in two principal planes at different frequencies [40]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.

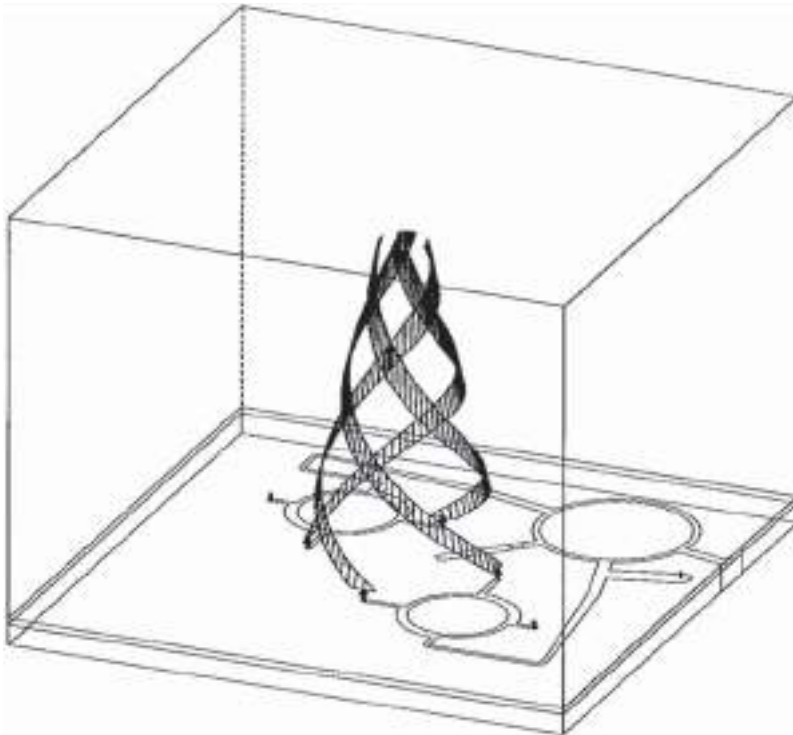


**Figure 3.32** Geometry of the unwrapped PQHA with tapered arms

conductor and the parasitic strip can be shorted [44,45] or separated [46]. Figure 3.36 presents the geometry of an unwrapped PQHA with parasitic strips. It is printed on a thin dielectric substrate of thickness 0.127 mm and relative permittivity 2.2, and wrapped around a cylindrical support. Each strip is formed by a conductor strip with a width of  $W_a$  and a parasitic strip with a width of  $W_b$ . The conductor strip is excited and the parasitic strip is shorted

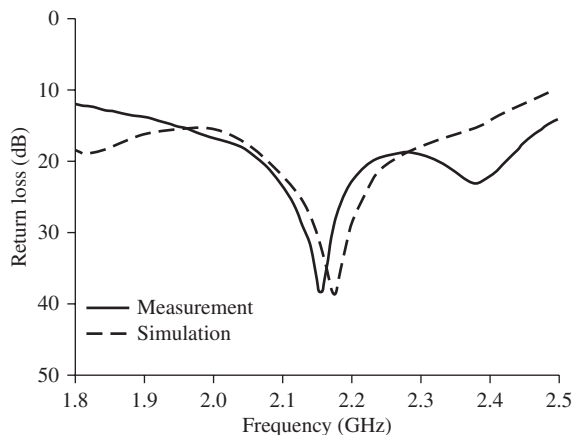


**Figure 3.33** Geometry of the unwrapped conical PQHA [43]. ( $R_a = 104.67$  mm,  $R_b = 24.53$  mm and  $\alpha = 82.55^\circ$ ) Reproduced with permission of © 2002 John Wiley & Sons, Inc.

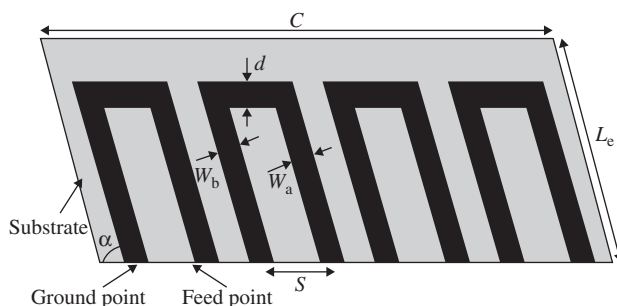


**Figure 3.34** 3D HFSS model for the integrated conical QHA [43]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

to the ground. The top ends of the conductor strip and the parasitic strip are shorted. The spacing between the conductor strip and the parasitic strip is  $s$  and the width of the shorted section is  $d$ . The parasitic study in [44] has clarified that the parameters of  $W_a$ ,  $W_b$ ,  $d$  and  $s$  play an important role on the impedance matching of the antenna. The obtained impedance bandwidth of  $VSWR \leq 2$  reaches about 30% (1.19–1.62 GHz) in axial-mode in the case of



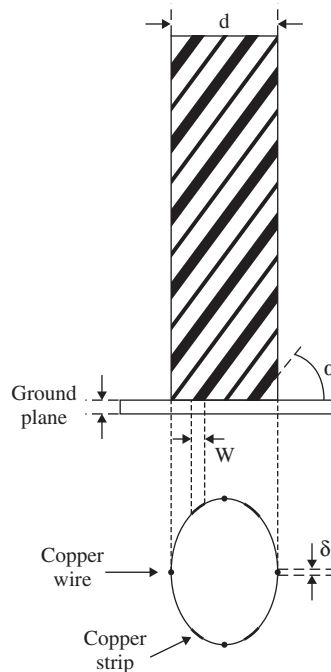
**Figure 3.35** Simulated (—) and measured (—) return loss against frequency [43]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.



**Figure 3.36** Geometry of the unwrapped PQHA with parasitic strips [44]. Reproduced with permission of © 2006 IEEE

$W_b = 2$  mm,  $W_a = 16$  mm,  $s = 18$  mm,  $d = 0.5$  mm,  $\alpha = 50^\circ$  and  $L_e = 166$  mm. The measured radiation patterns have shown a large beamwidth of  $150^\circ$  with an AR less than 2 dB at different frequencies. The measured antenna gain is over 1.5 dBic across the CP bandwidth.

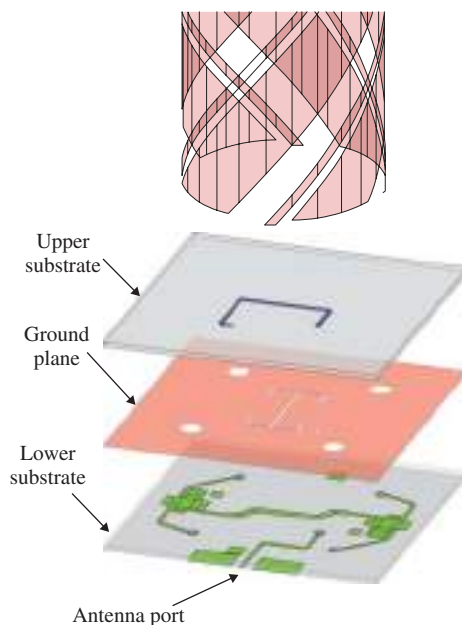
Figure 3.37 presents the configuration of another type of PQHA with parasitic strips [46]. Four equal-length copper wires are wound as four helices on a cylindrical core. The open ends of the four helix wires are left open. In order to increase the bandwidth, four equal-length copper strips are wound between every two helices. The four copper strips are shorted to the ground plane with the top ends open. The four helices are excited by a T-splitter power divider circuit printed on the bottom layer of the ground. The helices are connected to the feeding circuit through small via holes in the ground plane. The obtained impedance bandwidth ( $VSWR \leq 2$ ) reaches 39% and the 3-dB AR bandwidth is found to be 170 MHz (3.57–3.74 GHz). The CP bandwidth will be changed to be 40 MHz (3.63–3.67 GHz), when the parasitic strips are removed. In addition, the 3-dB beamwidth of the PQHA can be enlarged due to the presence of the parasitic strips.



**Figure 3.37** Geometry of a broadband PQHA with parasitic strips [46]. Reproduced with permission of © 2001 John Wiley & Sons, Inc.

### 3.3.2.5 Broadband Feed Network

In order to realize the broadband folded PQHA, a compact broadband feeding circuit is required to integrate with the folded PQHA. In [45], the feeding circuit consists of a compact aperture-coupled microstrip transition enclosed in the bottom section of the antenna, as shown in Figure 3.38. The dual-layer structure has the advantage of providing broadband performance. From the antenna port on the bottom layer of the lower substrate, a microstrip to coaxial transition allows connection to a line placed on the upper substrate. This line excites the aperture-slot transition which couples to a microstrip line printed on the bottom layer of the lower substrate. The outputs of the microstrip line have a  $180^\circ$  phase difference. Commercial surface mount technology (SMT)  $90^\circ$  hybrids are connected to both outputs of the aperture-slot transition to allow for a four port circuit providing a  $90^\circ$  phase sequential rotation. The prototype of the integrated folded PQHA is illustrated in Figure 3.39 and good CP performance with a 29% bandwidth has been obtained. The performances of the integrated PQHA at F1 (1.225 GHz), F2 (1.375 GHz) and F3 (1.575 GHz) are summarized in Table 3.2. As observed, the maximum gains at F1, F2 and F3 are 1.8, 2.3 and 1.5 dBic, respectively. The boresight gains are lower than the maximum gains and found to be 1.5, 1.3 and 0 dBic at three frequencies, respectively. The HPBW increases slightly against frequency, with  $145^\circ$  at F1,  $160^\circ$  at F2 and  $180^\circ$  at F3. The AR at boresight is 1.5 dB at F1, 0.7 dB at F2 and 2 dB at F3. The beamwidth (AR < 3 dB) is varying from  $200^\circ$  for F1 and F2, to  $150^\circ$  for F3. The front-to-back ratio reaches 8, 15 and 6 dB at F1, F2 and F3, respectively.



**Figure 3.38** Geometry's exploded view of the compact feeding circuit integrated with the folded PQHA [45]. Reproduced with permission of © 2010 IEEE



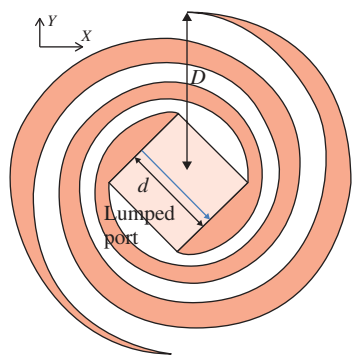
**Figure 3.39** Prototype of a broadband folded PQHA with a wideband feeding circuit [45]. Reproduced with permission of © 2010 IEEE

### 3.4 Spiral Antennas

The planar log-spiral antenna belongs to the class of frequency-independent antennas [36]. They can produce a good CP performance over a wide frequency range. Figure 3.40 presents the HFSS model of a planar log-spiral antenna in free space. The presented arrangement can excite RHCP outward from the page. The highest operating frequency is limited by the spacing  $d$  of the input terminal, whereas the lowest operating frequency is determined by

**Table 3.2** Performance of the broadband integrated PQHA at F1, F2 and F3 [45]. Reproduced with permission of © 2010 IEEE

| Frequency                      | F1  | F2  | F3  |
|--------------------------------|-----|-----|-----|
| Max. Gain (dBic)               | 1.8 | 2.3 | 1.5 |
| Gain at boresight (dBic)       | 1.5 | 1.3 | 0   |
| HPBW (deg.)                    | 145 | 160 | 180 |
| AR at boresight (dB)           | 1.5 | 0.7 | 2   |
| Beamwidth for AR < 3 dB (deg.) | 200 | 200 | 150 |
| AR max. in the HPBW (dB)       | 1.9 | 1.5 | 4.0 |
| Front-to-back ratio (dB)       | 8   | 15  | 6   |



**Figure 3.40** Geometry of a planar log-spiral antenna

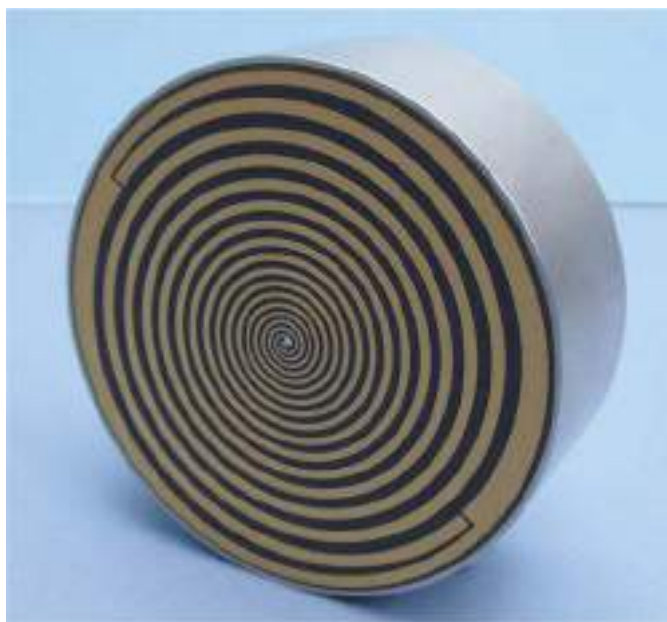
the overall diameter  $2D$ . The structure in Figure 3.40 can achieve bidirectional radiation patterns. Other configurations [36] include planar Archimedean antennas and sinuous antennas. A reflector or cavity can be attached in the bottom to achieve directional patterns, which has been investigated in [47,48]. The low-profile two-arm Archimedean spiral antenna in [48] is backed by a conducting cavity. The height of the cavity is extremely small and is chosen to be 7 mm. Only one arm is directly excited with the other one being parasitically excited, or the spiral arms are excited in an unbalanced mode. In this arrangement, a balun is not required for this unbalanced-mode spiral. The unbalanced spiral antenna in [48] can achieve a wide frequency range from 3 to 9 GHz. The radiation is circularly polarized around the antenna axis normal to the spiral plane due to the presence of the cavity. It shows similar gain performance with that of the balanced-mode spiral.

In most cases, the number of arms is chosen to be two and these two arms are excited in balanced mode by a broadband balun. In order to reduce the complexity of the design and fabrication, the spiral antenna with two arms can be externally excited by an integrated balun including tapered stripline [49] or microstrip line [50]. The antenna in [49] has a dual-layer substrate and three metal layers. The finite-ground stripline feed structure is synthesized based on the gap-fed spiral and realized by moving one arm of the gap-fed spiral to the top layer of the upper substrate and creating a copy on the bottom layer of the lower substrate. These two arms act as top and bottom grounds in the finite-ground stripline feed structure.

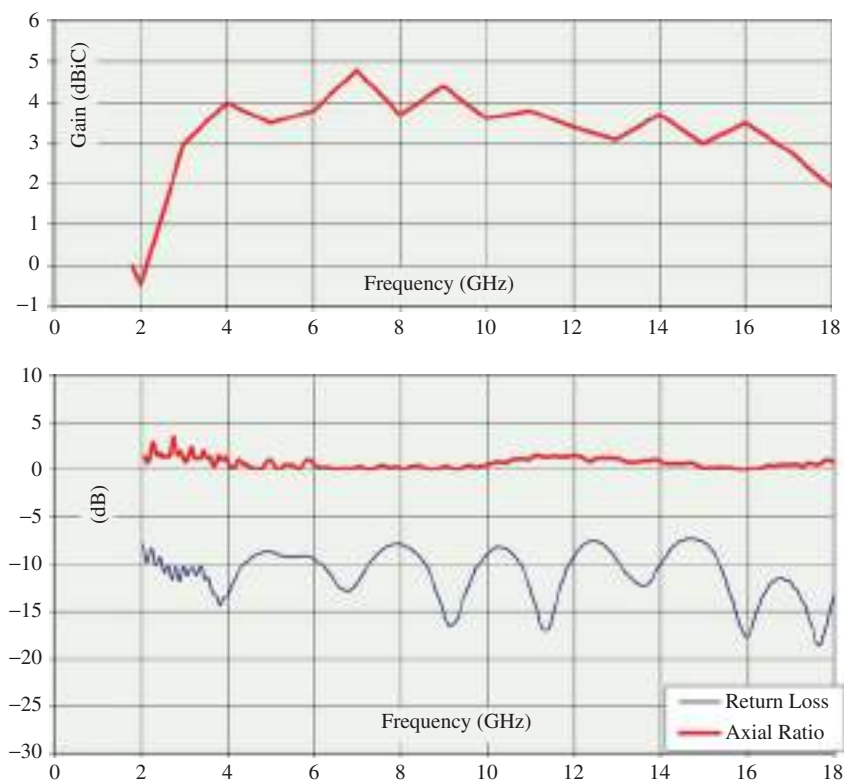
A signal line is centred and conformal to the spiral winding of the top and bottom grounds and extends through the feed gap until it connects with the second arm. In order to satisfy the characteristic impedance to be  $50\ \Omega$ , the signal line is modified to be a tapered microstrip line. The antenna in [49] can provide CP performance across a broad frequency range from 2 to 20 GHz. The equiangular spiral antenna in [50] has employed a tapered microstrip balun which is integrated into one spiral arm. The spiral-shaped radiator is printed on the bottom side of the Duroid substrate and the tapered microstrip line is on the top side. One arm behaves as the ground plane of the microstrip line and the other one is connected to the tapered microstrip line through a metal via. The 10-dB return loss and 3-dB AR bandwidths of the antenna in [50] are from 3.75 to 18.6 GHz and from 3 to 14.5 GHz, respectively.

Recently, a three-arm spiral antenna has been studied in [51]. It presented broadband radiation characteristics and CP performance. Also, it has two advantages. Firstly, unlike the conventional spiral antennas, the 3-arm spiral antenna can be fed by an unbalanced transmission line and an external balun is not required. Secondly, the antenna can be directly matched to the transmission line when selecting proper dimensions of the arms and the external matching network is not needed. Moreover, the spiral antenna can also be formed by four arms. As discussed in [52], the externally fed four ports spiral antenna consists of four arms and can radiate a CP wave over a wide frequency band within a wide angular range. Interestingly, the polarization can be right/left circular or linear vertical/horizontal by adjusting the amplitudes and phases of different ports.

Figure 3.41 presents a prototype of a cavity-backed spiral antenna which is proposed by Q-par Angus Ltd. The antenna presented in [53] is manufactured from aluminium and engineering composites. It features various advantages including compact size, light weight,



**Figure 3.41** Photo of a 2–18 GHz cavity-backed spiral antenna [53]. Reproduced with permission of © 2013 Q-par Angus Ltd



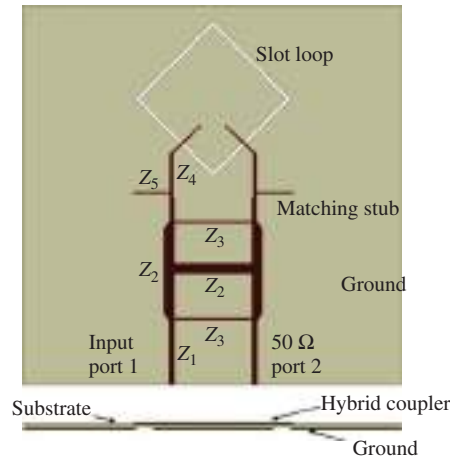
**Figure 3.42** Measured results of antenna gain, reflection coefficient and axial ratio [53]. Reproduced with permission of © 2013 Q-par Angus Ltd

smooth and stable radiation patterns, and purity of CP radiation. Thus it is suitable for many airborne and direction-finding applications. The diameter and height of the antenna are 78 and 46 mm, respectively. Measured results of antenna gain, reflection coefficient and AR are shown in Figure 3.42. As observed, the obtained impedance bandwidth ( $VSWR \leq 2.5$ ) ranges from 2 to over 18 GHz. The typical and maximum ARs in the operating frequency band are less than 1.8 and 4 dB, respectively. The antenna gain varies from  $-0.5$  to  $4.8$  dBic. The obtained 3-dB beamwidth across the frequency range of interest is from  $100^\circ$  to  $60^\circ$ .

### 3.5 Broadband CP Slot Antennas

#### 3.5.1 Ring Slot with a Broadband Coupler

The hybrid coupler which is used in the broadband CP patch antenna can also help the CP ring slot antenna increase the bandwidth. Figure 3.43 presents the geometry of a ring slot antenna with a hybrid coupler. A square ring slot is embedded in the ground which is printed on the bottom layer of an FR4 substrate. Two adjacent sides of the slot loop are excited by a three-stub hybrid coupler which is printed on the other side of the substrate. It is noted that two stubs are added to improve impedance matching. The characteristic impedances of the



**Figure 3.43** Geometry of a ring slot antenna with a hybrid coupler

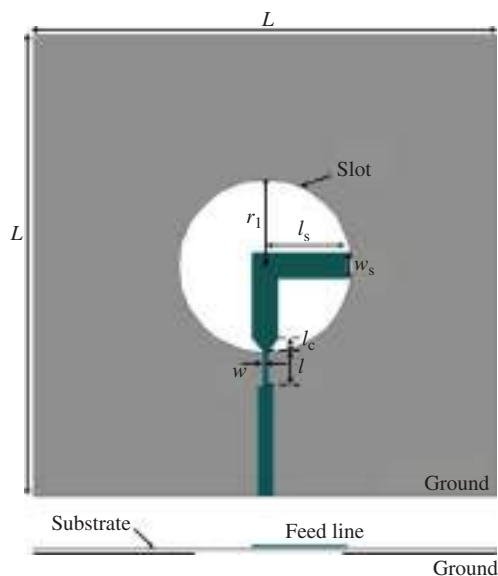
stubs are:  $Z_1 = 50 \Omega$ ,  $Z_2 = 35.4 \Omega$ ,  $Z_3 = 120.8 \Omega$ ,  $Z_4 = 67.6 \Omega$  and  $Z_5 = 92.8 \Omega$ . The antenna in [54] can achieve an impedance bandwidth ( $VSWR \leq 2$ ) of 55.6% (3.9–6.9 GHz). The measured 3-dB AR bandwidth reaches around 22.2% (4.8–6.0 GHz). The antenna gain ranges from  $-0.5$  to  $1.4$  dBi across the operating frequency range.

### 3.5.2 Circular/Square Slot with an L-Shaped Feed

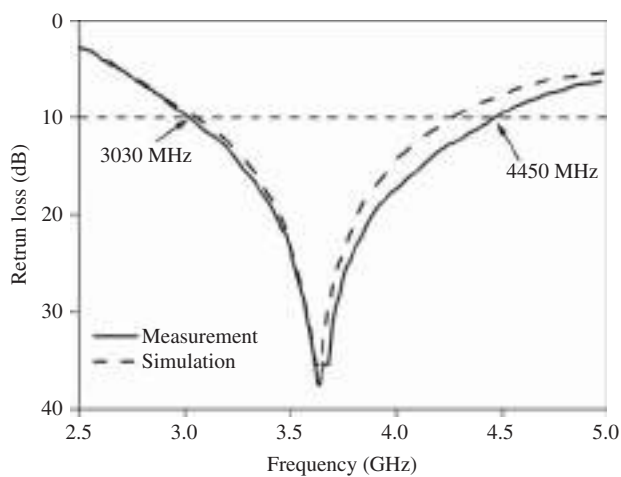
Compared to microstrip antennas, the printed slot antennas, especially wide slot antennas, can have broader impedance bandwidths. Several approaches have been taken to achieve good CP performance over a wide frequency range. These include the L-shaped feedline [55,56] or grounded strip [57–60]. In addition, circular polarization of the slot antenna can be obtained by applying modifications on the feedline with grounded strips [61–63].

Figure 3.44 presents the geometry of a microstrip-fed CP slot antenna. A circular slot with a radius of 18.8 mm is embedded in the ground plane which is printed on one layer of an FR4 substrate with a thickness of 1.6 mm and a relative permittivity of 4.4. An L-shaped strip with a taper end is fabricated on the other side of the substrate to excite two orthogonal fields in phase quadrature. CP operation is obtained due to the L-shaped strip. A narrow line with a length of  $l$  and a width of  $w$  behaves as an impedance transformer between the taper end of the L-shaped strip and the  $50\Omega$  microstrip line. The length of the tapered portion is  $l_c$ .

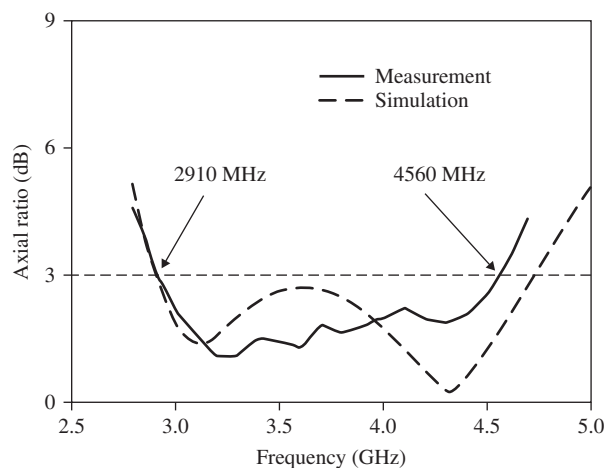
The simulated and measured results of return loss and AR are illustrated in Figures 3.45 and 3.46. The measured 10-dB return loss bandwidth is 1.42 GHz or 37.97% at 3.74 GHz. The measured result exhibits a 3-dB AR bandwidth of 1.65 GHz or 44.1% at 3.74 GHz. The slot antenna radiates towards both sides, and the patterns at both sides have different senses of circular polarization. The measured antenna gain varies from 4 to 5 dBi across the operating frequency band. In addition, a CPW-fed slot antenna with an embedded square slot and an L-shaped strip connected to the edge of the slot is studied in [56]. The width of the L-strip is larger than that of the signal line. The obtained 10-dB return loss and 3-dB AR bandwidths of the CPW-fed slot antenna are 43% and 17%, respectively.



**Figure 3.44** Geometry of a microstrip-fed CP slot antenna. Antenna dimensions:  $r_1 = 18.8$  mm,  $l_s = 17.7$  mm,  $w_s = 5.5$  mm,  $l_c = 11.7$  mm,  $l = 7.2$  mm,  $w = 0.5$  mm,  $L = 100$  mm [55]. Reproduced with permission of © 2008 John Wiley & Sons, Inc.



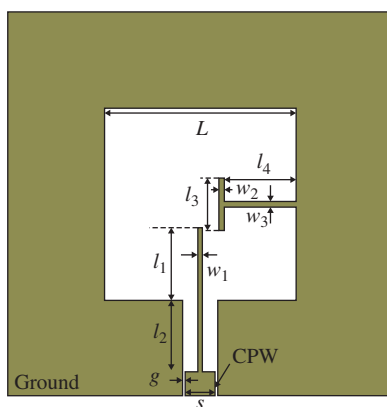
**Figure 3.45** Simulated and measured return loss for the slot antenna [55]. Reproduced with permission of © 2008 John Wiley & Sons, Inc.



**Figure 3.46** Simulated and measured AR for the slot antenna [55]. Reproduced with permission of © 2008 John Wiley & Sons, Inc.

### 3.5.3 Grounded Strip

Figure 3.47 presents the geometry of a CPW-fed CP slot antenna which is printed on a single layer of an FR4 substrate with a thickness of 1.6 mm and a relative permittivity of 4.4. A square slot with a side length of  $L = 40$  mm is embedded in a square ground with a side length of 70 mm. A T-shaped strip with a width of  $w_2 = w_3 = 1$  mm is protruded toward the slot centre at the centre of one slot edge. The horizontal and vertical portions of the T-shaped strip have the lengths of  $l_4$  and  $l_3$ , respectively. A narrow strip with a width of  $w_1$  and a length of  $l_1 + l_2$  is protruded from the end of the signal line with a width of  $s = 6.4$  mm. The gap between



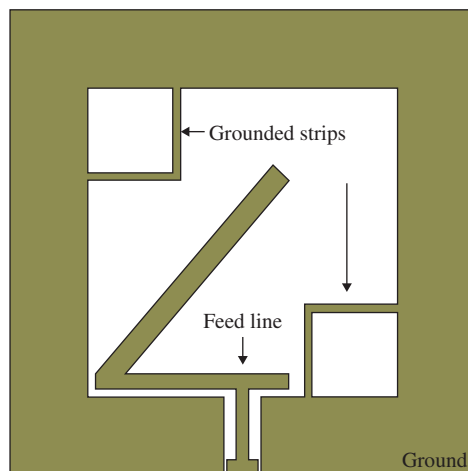
**Figure 3.47** Geometry of a CPW-fed CP slot antenna with a T-shaped strip protruding the ground [57] ( $l_1 = 15$  mm,  $l_2 = 9.7$  mm,  $l_3 = 9.2$  mm,  $l_4 = 15$  mm and  $w_1 = 1$  mm). Reproduced with permission of © 2003 IEEE

the signal strip and the ground plane has a length of  $g = 0.5$  mm. Good CP performance over a wide frequency range can be achieved by tuning the dimensions of  $l_1$ ,  $l_2$ ,  $l_3$ ,  $l_4$  and  $w_1$ . It is also worthwhile to mention that the design in Figure 3.47 gives LHCP radiation.

The obtained impedance bandwidth defined by  $VSWR \leq 2$  is 874 MHz or 44.5% at 1.965 GHz. The measured 3-dB and 1-dB AR bandwidths are 18.8% and 12.7%, respectively. The slot antenna without a reflector has bi-directional radiation patterns. The gain variation is less than 1 dB with a peak gain of around 3.7 dBi. Similarly, the circular CPW-fed slot antenna with a circular slot and having a broader CP bandwidth is proposed in [58]. Circular polarization is obtained by protruding a metallic mono-strip from the circular ground plane towards the slot centre. The circular slot antenna in [58] has an impedance bandwidth ( $|S_{11}| \leq -10$  dB) and a 3-dB AR bandwidth of 50% and 36%, respectively.

Instead of using CPW feed, the microstrip-fed slot antenna in [59] consists of a microstrip T-junction, a rectangular wide-slot and a T-shaped stub protruded at the slot edge. A portion of microstrip line with a high characteristic impedance is arranged between the T-junction and the 50- $\Omega$  feedline for improving the impedance matching. The microstrip-fed slot antenna in [59] can achieve a 10-dB return loss bandwidth of 58% and a 3-dB AR bandwidth of 22.2%. Furthermore, a CPW-fed square slot antenna with two grounded inverted-L metallic strips is studied in [60]. A square slot is embedded in the square ground plane and excited by a 50 $\Omega$  CPW. A strip with a larger width protrudes from the signal line of the CPW into the slot. The CP operation is mainly attributed to the two grounded inverted-L metallic strips located around two opposite corners of the square slot. The obtained impedance bandwidth ( $VSWR \leq 2$ ) and 3-dB AR bandwidth are 52% and 25%, respectively.

The impedance and AR bandwidths of the slot antennas can be further increased by using grounded strips and modifying the feedline, which has been investigated in [61–63]. Figure 3.48 presents the geometry of a CPW-fed square slot antenna [61]. The design has two main features: one for increasing the AR bandwidth and the other mainly for enhancing the impedance bandwidth. The AR bandwidth is enlarged due to the combination of the



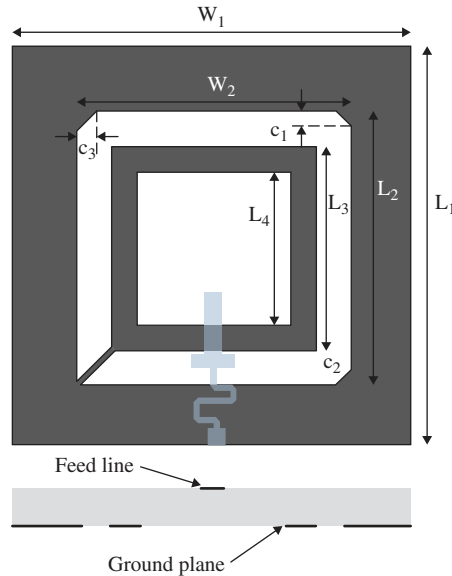
**Figure 3.48** Geometry of a square slot antenna with a lightning-shaped feedline and inverted-L grounded strips [61]. Reproduced with permission of © 2010 IEEE

lightening-shaped feedline and two inverted-L grounded strips on the opposite corners of the square slot. The lightening-shaped feedline is realized by extending the signal line of the CPW to the lower left corner of the slot and then protruding into the slot at an inclined  $45^\circ$ . In order to increase the impedance bandwidth, two tuning stubs are embedded in the feedline. The vertical tuning stub is formed by shrinking a section of the CPW's signal line while the horizontal one is realized by extending the horizontal feed section. The second resonant mode will be shifted to a lower frequency band and combined with the first resonant mode to achieve a wider impedance band, when choosing proper dimensions for the tuning stubs. The antenna in [61] can achieve a 10-dB return loss and 3-dB AR bandwidths of 50.9% and 48.8%, respectively.

More recently, a CPW-fed regular-hexagonal slot antenna is proposed in [62]. The ground plane is formed by a hexagonal ring and two L-shaped strips. An L-shaped monopole patch is connected to the signal line of the CPW. Wide impedance and AR bandwidths can be obtained by employing an L-shaped monopole patch protruding into the slot and two inverted-L grounded strips. Moreover, the impedance and AR bandwidths can be further enlarged by inserting a rectangular notch in the ground plane. The antenna in [62] can achieve a 10-dB return loss and 3-dB AR bandwidths of 86% and 50%, respectively. Compared to adding two equal-size inverted-L strips, the square slot antenna in [63] has a square ground plane with a wide square slot and two unequal-sized inverted-L shaped strips at two opposite corners. Two orthogonal fields in quadrature can be generated by the grounded strips. A rectangular patch is connected to the signal line of the CPW. Also, a tuning slit is inserted in the feedline and a tuning vertical stub is attached to the signal line. The impedance bandwidth is significantly enlarged owing to these modifications. The slot antenna in [63] can operate over a wide frequency range with an impedance bandwidth ( $\text{VSWR} \leq 2$ ) of 132% and a 3-dB AR bandwidth of 32.2%.

### 3.5.4 Shorted Ring Slot

Regarding the ring slot antenna, the investigations in [64,65] have revealed that circular polarization can be obtained by using a shorted strip [64] at a proper position or a protruded strip [65]. A broadband CP shorted ring slot antenna has been proposed in [66]. As shown in Figure 3.49, the shorted square-ring slot antenna is printed on an FR4 substrate with a thickness of 1.6 mm and a relative permittivity of 4.4. The ground plane is fabricated on the bottom side of the substrate and has a length of  $L_1$  and a width of  $W_1$ . The outer and inner lengths of the square ring slot are  $L_2$  and  $L_3$ , respectively. In order to generate CP fields, a shorted strip is diagonally included. It is noted that, the meandered impedance transformer and three tuning stubs are utilized to reduce the size and match the 50- $\Omega$  transmission line. Moreover, a square slot with a length of  $L_4$  is etched at the centre of the antenna to increase the bandwidth. Three small triangles (with widths of  $c_1$ ,  $c_2$  and  $c_3$ ) are located at the corners of the shorted ring slot to provide more symmetric radiation patterns. It is worthwhile to mention that the impedance bandwidth is determined by the meandered transformer and three tuning stubs, while the AR bandwidth depends on the dimensions of the square ring slot. The antenna in [66] can provide an impedance bandwidth ( $|S_{11}| \leq -10$  dB) of 38.7% at 2.4 GHz (1.83–2.76 GHz). The obtained 3-dB AR bandwidth is 2.21–2.52 GHz or 12.9% at the centre frequency. The minimum AR value is 0.1 dB at 2.4 GHz. A relatively constant gain response of about 3.3 dBi has been obtained within the 3-dB AR bandwidth.

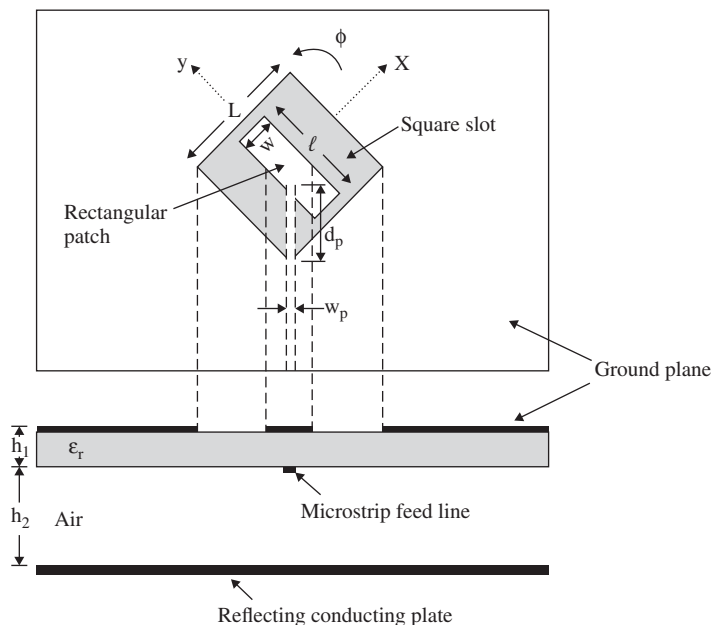


**Figure 3.49** Geometry of the shorted ring slot antenna [66]. Reproduced with permission of ©2007 John Wiley & Sons, Inc.

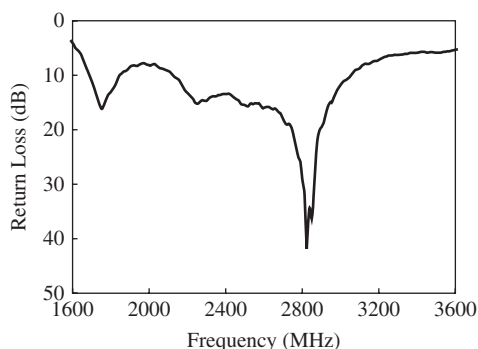
### 3.5.5 Wide Slot with a Loading Patch

Circular polarization of the broadband slot antennas can also be achieved by loading a patch [67–69]. When selecting proper dimensions and positions of the patch, two orthogonal resonant modes with a  $90^\circ$  phase difference for CP operation can be generated. In order to block the backward radiation and increase the antenna gain, a conducting patch can be employed below the antenna [68,69]. Figure 3.50 presents the geometry of a microstrip-fed square slot antenna. It is printed on an FR4 substrate with a relative permittivity of 4.4 and a thickness of 1.6 mm. A square slot with a side length of  $L$  is embedded in the ground plane with a size of  $120 \times 120 \text{ mm}^2$ . The loading rectangular patch having a length of  $l$  and a width of  $w$  is oriented along the  $y$ -axis. A microstrip line with a width of  $w_p$  and a tuning length of  $d_p$  is located at  $45^\circ$  with respect to the rectangular patch. CP fields can be generated by choosing suitable dimensions of the loading patch. In order to reduce the backward radiation, a conducting plate is located at a distance of  $h_2$  away from the microstrip line. Normally, the distance  $h_2$  can be a quarter wavelength of the centre operating frequency. The values of optimized dimensions for the slot antenna are given as:  $h_2 = 31.5 \text{ mm}$ ,  $L = 44.7 \text{ mm}$ ,  $l = 31.5 \text{ mm}$ ,  $w = 12.2 \text{ mm}$ ,  $d_p = 24.2 \text{ mm}$  and  $w_p = 3.1 \text{ mm}$ .

The measured results of return loss and AR are shown in Figures 3.51 and 3.52. The obtained 10-dB return loss bandwidth reaches 924 MHz and the 3-dB AR bandwidth is found to be 284 MHz. Or the impedance and AR bandwidths are 39% and 12% at the centre frequency of 2.375 GHz, respectively. The gain variation of less than 1.3 dB is obtained and the antenna gain at the centre frequency is about 4.8 dBi. The measured radiation patterns in



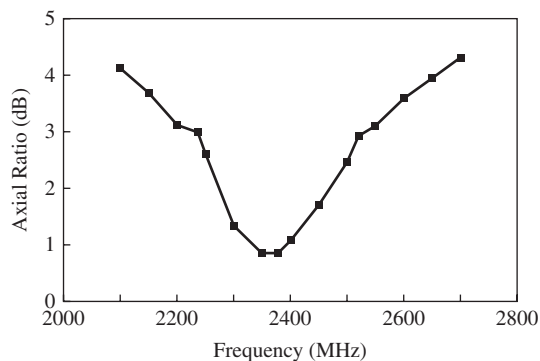
**Figure 3.50** Configuration of the square slot antenna with a rectangular conducting patch [69]. Reproduced with permission of ©1999 John Wiley & Sons, Inc.



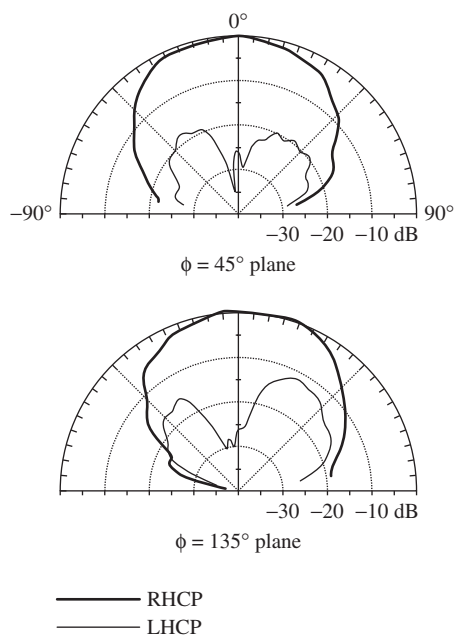
**Figure 3.51** Measured return loss versus frequency [69]. Reproduced with permission of ©1999 John Wiley & Sons, Inc.

two orthogonal planes at 2.375 GHz are shown in Figure 3.53, in which good CP radiation has been obtained.

Moreover, a CPW-fed square slot antenna is proposed in [67]. It consists of a square ground plane with an embedded square slot and a cross patch loaded at the centre of the slot. The cross patch is inclined with respect to the signal line at an angle of  $45^\circ$ . It is noted that, the loaded-cross-patch with unequal arm lengths makes the fundamental resonant mode of the square slot split into two near-degenerate orthogonal modes for CP operation.

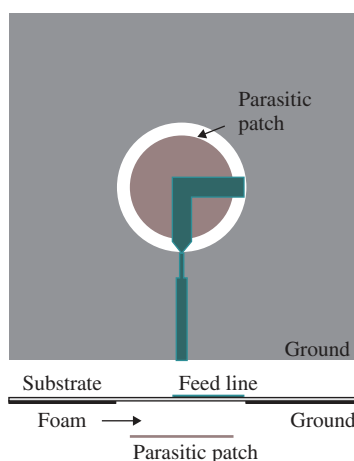


**Figure 3.52** Measured AR versus frequency [69]. Reproduced with permission of ©1999 John Wiley & Sons, Inc.



**Figure 3.53** Measured radiation patterns at 2.375 GHz [69]. Reproduced with permission of ©1999 John Wiley & Sons, Inc.

Good CP radiation can be obtained by tuning the dimensions of the cross patch. The signal line of the CPW has two portions with different widths to improve the impedance matching. The square slot antenna in [67] can achieve a 10-dB return loss and 3-dB AR bandwidths of 39.6% and 12.4%, respectively. More recently, a patch-loaded square slot antenna with a largely enhanced AR bandwidth is investigated in [68]. A square slot is inserted in the square ground plane. A reference antenna with an asymmetric T-shaped strip protruded from the signal line of the CPW into the square slot can have a CP bandwidth of around 11%.



**Figure 3.54** Geometry of a microstrip-fed CP slot antenna with a parasitic patch [70]. Reproduced with permission of © 2008 IEEE

A modified antenna can be obtained by stretching the left end of the T-shaped trip upward, loading a rectangular patch in the square slot and implanting a grounded strip at the right edge of the slot. The modified antenna can achieve a much larger CP bandwidth of 39.1%. In order to achieve unidirectional radiation patterns, a simple reflector is attached below the antenna. The peak gain can be increased from 4.2 to 7.6 dBic while the CP bandwidth is slightly decreased to be 34.3%.

### 3.5.6 Other Broadband CP Slot Antennas

The bandwidth of the aforementioned microstrip-fed CP slot antenna in [55] can be further increased by loading a parasitic patch [70], as shown in Figure 3.54. Based on the microstrip-fed CP slot antenna shown previously in Figure 3.44, a circular patch is printed on another substrate and supported by a foam material below the ground plane. The CP bandwidth can be enlarged when the resonant frequencies excited by the circular slot and the circular patch are close to each other. The bandwidth of the patch-loaded slot antenna can vary from 5% to 45% by adjusting the distance between the slotted ground and the parasitic patch. In addition, the antenna gain can be improved when the parasitic patch is located at a proper position. Moreover, the antenna gain can be further enlarged and the unidirectional radiation can be obtained by adding a reflector, which has been studied in [71]. The antenna in [71] consists of a triangular-shaped slotted ground plane, a triangular parasitic patch, a reflecting conducting plate and is fed by an L-shaped feedline. It can provide an average gain of 10 dBi within a CP operating bandwidth of 23% and a front-to-back ratio of over 10 dB.

In addition to circular or square slots, CP slot antennas can also be obtained by employing Spidron-shaped slot, which is studied in [72]. The Spidron is a plane formed by equilateral and isosceles triangles. The slot of the microstrip-fed antenna is generated by merging downscaled copies of a right-angled triangular slot. A simple 50- $\Omega$  microstrip line is utilized to excite the Spidron-slot. The Spidron-shaped slot can provide CP radiation with a

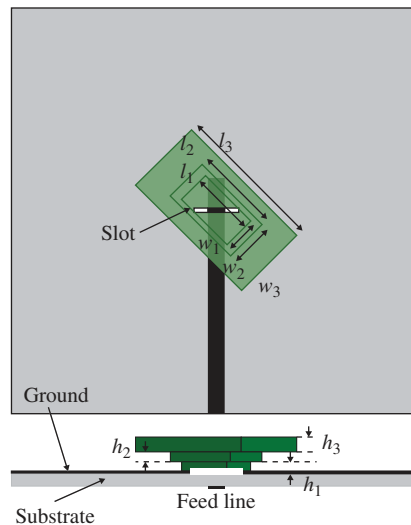
broadband bandwidth. The slot antenna in [72] can achieve 10-dB return loss and 3-dB AR bandwidths of 78.3% and 15.2%, respectively.

### 3.6 Broadband CP DRAs

#### 3.6.1 Single-Feed Stacked DRAs

Figure 3.55 presents the geometry of an aperture-coupled CP stair-shaped DRA which is fabricated from a ceramic material with a dielectric constant of 12. The DR is in a flipped two-step stair form with rectangular cross section. The length and width of each layer is  $l_1 = 7.86$  mm,  $w_1 = 4.14$  mm,  $l_2 = 10.48$  mm,  $w_2 = 5.52$  mm and  $l_3 = 18.34$  mm,  $w_3 = 9.66$  mm. They have the same length to width ratio of 1.9. The thickness of each layer is  $h_1 = 1.3$  mm,  $h_2 = 1.3$  mm and  $h_3 = 1.8$  mm. A wide impedance bandwidth can be obtained as the stair DRA supports multiple resonances. The DRA is excited by a 50- $\Omega$  microstrip line through a rectangular slot with a length of 5.6 mm and a width of 0.5 mm. The microstrip line is terminated with an open circuited stub of 4 mm long and printed on the bottom layer of a substrate with a thickness of 0.813 mm and a relative permittivity of 3.38 while the slotted ground is on the top layer. Circular polarization can be realized by rotating the DR at an angle of  $45^\circ$  with respect to the slot, as the inclined slot can generate orthogonal degenerate modes.

The measured impedance bandwidth ( $|S_{11}| \leq -10$  dB) is about 36.6% (7.56–10.95 GHz). The measured 3-dB AR bandwidth reaches 10.6% (9.4–10.45 GHz). At 10 GHz, the obtained 3-dB beamwidth and 3-dB AR beamwidth are  $67^\circ$  and  $54^\circ$ , respectively. The RHCP level is 15 dB lower than LHCP. The front-to-back ratio is found to be 16 dB. Moreover, a trapezoidal DRA which uses the similar operating principle is studied in [74].



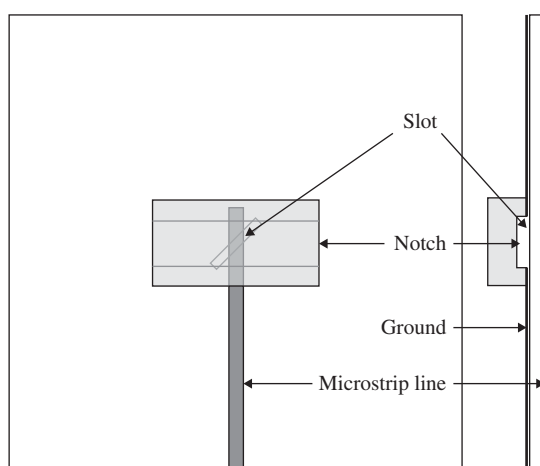
**Figure 3.55** Geometry of an aperture-coupled CP stair DRA [73]. Reproduced with permission of © 2006 IEEE

The trapezoidal DRA is excited by a microstrip line through a rectangular slot in the ground plane. Circular polarization is obtained by rotating the DR by  $45^\circ$  with respect to the slot. The measured 10-dB return loss and 3-dB AR bandwidths in [74] can reach 33.5% (2.88–4.04 GHz) and 21.5% (3.11–3.86 GHz), respectively.

### 3.6.2 Single-Feed Notched DRAs

The bandwidth of CP DRA can be increased by using notch DR due to the equivalent low  $Q$  value. Figure 3.56 shows the geometry of a microstrip line-fed slot-coupled notch DRA which has a notch DR made of DR and air. This configuration will reduce the equivalent dielectric constant and relatively lower the  $Q$  value, thus increasing the bandwidth. The inclined slot embedded in the ground can produce orthogonal degenerate modes to achieve CP radiation. The 10-dB return loss bandwidth can reach 20% and 3-dB AR bandwidth is 5%, when choosing proper dimensions.

More recently, CP DRAs with a notch DR and wider bandwidth have been proposed in [76,77]. A wideband omnidirectional CP DRA presented in [76] has an inclined slot on each DR sidewall. Due to the slots, the omnidirectional LP field is perturbed and converted into the CP field. A conducting parasitic strip is placed inside each slot to increase the AR bandwidth. In addition, a central circular cylinder is removed from the DR to enhance the impedance bandwidth. The antenna is centrally excited by a coaxial probe, which protrudes into the hollow region from the inner conductor of a SMA connector. The antenna in [76] can achieve 10-dB return loss and 3-dB AR bandwidths of 24.5% (3.08–3.94 GHz) and 25.4% (3.16–4.08 GHz), respectively. In [77], a rectangular DR ring with a relative permittivity of 10.2 excited by a microstrip line is embedded inside a thin dielectric substrate and the ground plane to reduce the total volume of the antenna. It is also noted that the dielectric substrate is cut to accommodate the DR. The microstrip line and ground plane are printed on both sides of the dielectric substrate. To generate the orthogonal fields, two parasitic strips are used to



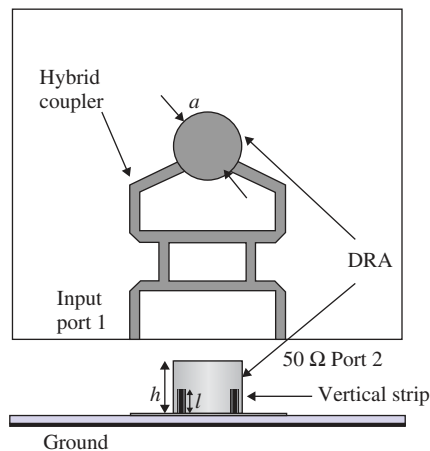
**Figure 3.56** Geometry of a CP notched DRA [75]. Reproduced with permission of © 2005 IEEE

form conducting strips and connected to the ground plane. Good CP performance can be obtained by choosing proper dimensions of the parasitic strips. The 10-dB return loss and 3-dB AR bandwidths of the antenna in [77] can be as large as 53% and 51%, respectively.

### 3.6.3 Multi-Feed DRAs with a Coupler

In addition to the dual-feed CP patch antennas and ring slot antennas, a hybrid coupler can be utilized in DRA to provide a broadband CP operation [78–82]. Figure 3.57 presents the geometry of a CP cylindrical DRA which is excited by dual conformal strips with a height of  $l = 5$  mm and a width of 2.3 mm. Two degenerate  $TM_{110}$  modes can be generated by the strips. In order to provide quadrature signals for achieving CP radiation, the output ports of a microstrip  $90^\circ$  hybrid coupler are connected to the strips. The hybrid coupler is printed on a substrate with a thickness of 0.787 mm and a relative permittivity of 2.33. The DR has a diameter of  $a = 7$  mm, a height of  $h = 10.8$  mm and a dielectric constant of 9.5. The obtained impedance bandwidth ( $|S_{11}| \leq -10$  dB) and 3-dB AR bandwidth are 13.7% and 20%, respectively.

The bandwidth of the CP DRA with a hybrid coupler can be further enlarged by using four-feed, as investigated in [79–81]. The CP cylindrical DRA in [79] is fed by four orthogonally-orientated vertical strips. A feed network is employed for providing signals with equal amplitudes and phase shifts of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$ . The feed network consists of a conventional  $180^\circ$  balun cascaded with a pair of conventional  $90^\circ$  hybrid couplers. The antenna in [79] can achieve an impedance bandwidth ( $|S_{11}| \leq -10$  dB) of 34.5% (1.75–2.48 GHz) and a 3-dB AR bandwidth of 25.9% (1.65–2.14 GHz). The cylindrical DRA in [80] is fed by four microstrip lines through four slots in the ground plane which is printed on the top layer of a dielectric substrate. A feed network providing quadrature signals is located on the bottom layer. The feed network comprises a wideband rat-race  $180^\circ$  hybrid coupler cascaded with two surface mount  $90^\circ$  branch-line hybrid couplers. The



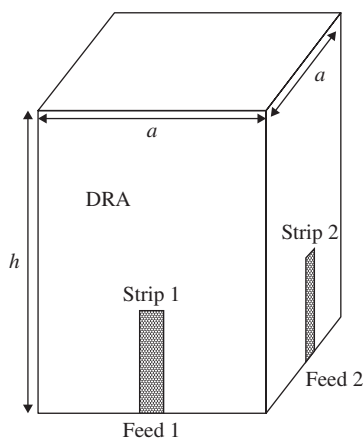
**Figure 3.57** Geometry of a CP DRA with a hybrid coupler [78]. Reproduced with permission of © 2000 IEEE

aperture-coupled four feed cylindrical DRA presented in [80] can achieve a 10-dB return loss bandwidth as high as 50% (1.08–1.82 GHz). The AR at boresight is less than 1 dB over the 1.13–1.63 GHz frequency range. Furthermore, a compact CP rectangular DRA with dual underlaid hybrid couplers is proposed in [81]. The antenna has a compact size, as no extra footprint is required for the couplers. The DRA is fed by four sequentially rotated conformal strips. In order to provide quadrature signals, the output ports of two hybrid couplers are connected to the strips. The dual underlaid hybrid couplers are arranged in parallel and connected through a meandered microstrip line with a length of  $\lambda_g/2$  ( $\lambda_g$  is the guided wavelength). The obtained 10-dB return loss and 3-dB AR bandwidths can reach 37.0% and 27.7%, respectively.

### 3.6.4 Multi-Mode DRAs

Only the fundamental  $TE_{111}$  mode is excited in the traditional DRA. Broadband DRAs can be achieved when using both the fundamental  $TE_{111}$  mode and the higher-order  $TE_{113}$  mode. This operating principle can also be applied in CP DRA [82–84]. Figure 3.58 presents the geometry of the dual-mode CP DRA. The DR has a dimension of  $a = 18$  mm,  $h = 29$  mm and a dielectric constant of 10. Dual identical feeding strips with a length of 10.7 mm and a width of 1.2 mm are attached to the middle of two side walls of the DRA. Circular polarization can be generated when the dual ports are fed by signals with equal magnitudes but quadrature phase. In this case, the theoretical resonant frequencies of the  $TE_{111}$  and  $TE_{113}$  modes are 3.22 and 4.03 GHz, respectively. When both modes are excited, the overall bandwidth of the antenna can be improved.

In measurement, the antenna in [82] can resonate at the frequencies of 3.04 and 3.65 GHz, respectively. The obtained impedance bandwidth ( $|S_{11}| \leq -10$  dB) is 32.8% (2.7–3.76 GHz). The average isolation between the two ports is better than 15 dB across the frequency band. The obtained 4-dB AR bandwidth is ranging from 2.6 to 3.78 GHz. The gains of the two CP modes are about 2.1 and 6.1 dBi, respectively.



**Figure 3.58** Geometry of a dual-mode CP DRA [82]. Reproduced with permission of © 2009 IEEE

### 3.6.5 Other Broadband CP DRAs

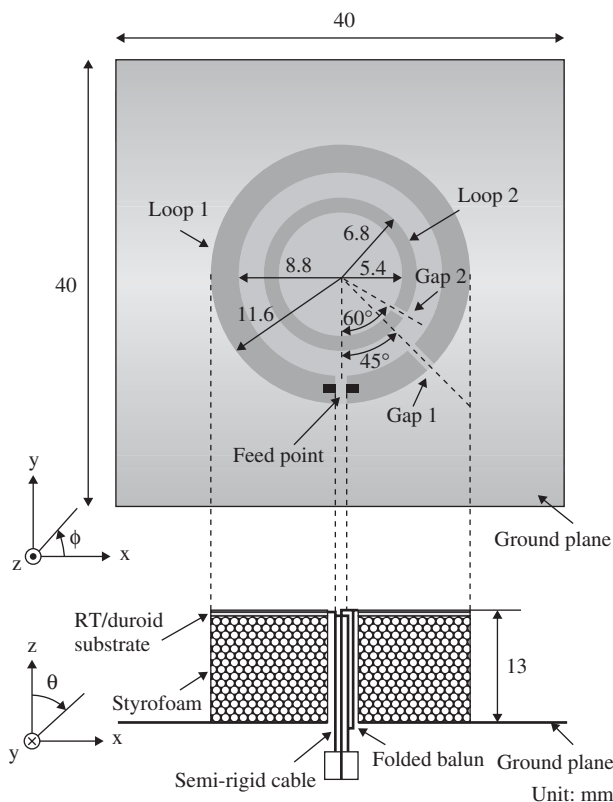
In addition to the above mentioned methods, the bandwidth of the CP DRA in [85] is increased by using concentric open half-loops. The rectangular DRA is excited by an open half-loop which is formed by three metallic strips. The AR bandwidth can be increased by adding another parasitic half-loop. The antenna in [85] can provide a 10-dB return loss bandwidth of 19% and a 3-dB AR bandwidth of 13%. More recently, a wideband CP dielectric bird-nest antenna with conical radiation patterns has been investigated in [86]. It comprises parasitic dielectric parallelepiped elements and a feeding probe. The probe protruding from a circular ground plane operates in its fundamental monopole mode which can generate omni-directional LP fields. A wave polarizer is realized by combining the dielectric parallelepipeds and converts the LP fields into CP fields. The antenna has a 10-dB return loss bandwidth of 41.0% and a 3-dB AR bandwidth of 54.9%, while the hybrid coupler is not required.

## 3.7 Broadband CP Loop Antennas

### 3.7.1 Parasitic Loop

Similar to broadband CP patch antennas, slot antennas and PQHAs, the bandwidth of CP loop antennas can be improved by introducing a parasitic loop, as studied in [87,88]. Figure 3.59 presents the geometry of the probe-fed CP loop antenna with a parasitic loop. The antenna is formed by two concentric wire loops: an outer loop (loop 1) behaves as the driven loop and an inner loop (loop 2) is a parasitic loop. A square ground plane with a size of  $40 \times 40 \text{ mm}^2$  is attached below the antenna to achieve unidirectional radiation. A small gap is embedded in the outer loop to excite CP fields and the inner loop (also with a small gap) is electromagnetically coupled to the outer loop. The gaps on loops 1 and 2 are placed at  $45^\circ$  and  $60^\circ$  away from the feed point, respectively. Both loops are printed on a thin Duroid 5880 substrate with a thickness of 0.254 mm and a relative permittivity of 2.2, and supported by a piece of Styrofoam ( $\epsilon_r = 1.03$ ,  $h = 13 \text{ mm}$ ). A quarter-wavelength folded balun is utilized to provide a balanced feed for the antenna. In order to cover a broad bandwidth, two folded baluns are employed in the design.

The 2-dB AR bandwidth is about 21%, while only 6% for the case without the parasitic loop. The combination of the two loops can lead to a significant enhancement for the CP bandwidth. The current distribution in [88] has shown that the current on loop 1 has a significant fluctuation in magnitude while the current in loop 2 has little feature of a travelling wave current distribution with a more uniform magnitude. The presence of the parasitic loop is to compensate the magnitude variation of the current on loop 1. It is also worthwhile to mention that the AR bandwidth will not be increased by placing the parasitic loop outside the driven loop, as the current on the parasitic loop shows little feature of a travelling-wave current distribution. The antenna gain varies from 7 to 8 dBi. It is also found that the VSWR of the antenna is around 3 over the operating frequency band. Thus an impedance transformer circuit is required for some practical applications.

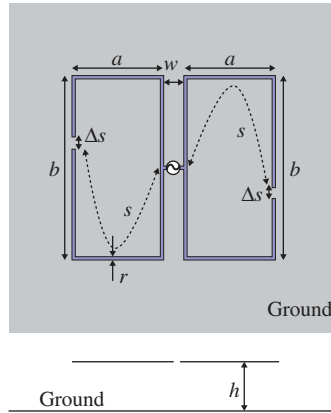


**Figure 3.59** Geometry of a probe-fed CP loop antenna with a parasitic loop [88]. Reproduced with permission of © 2005 IEEE

### 3.7.2 Dual Loops in Series or Parallel

Compared to the single-loop CP antenna, the dual-loop CP antenna can have a wider bandwidth [89–91]. Figure 3.60 presents the geometry of a dual-loop CP antenna. Two rectangular wire loops with the same size are fed in series with each loop is connected to each end of a short dipole. Circular polarization is obtained due to the small gaps which are located symmetrically with respect to the centre of the antenna. A ground plane is placed with a distance of  $h = 53$  mm below the antenna to enhance the antenna gain. The optimized values of the parameters are:  $a = 48.3$  mm,  $b = 96.7$  mm,  $s = 157.4$  mm,  $w = 10$  mm,  $\Delta s = 5.9$  mm and  $r = 2$  mm.

The results in [91] show that the real and imaginary values of the input impedance at 1.5 GHz are  $53.4$  and  $-0.3\Omega$ , respectively. The VSWR at 1.5 GHz reaches about 1.07. The obtained impedance bandwidth ( $\text{VSWR} \leq 2$ ) is around 22%. The minimum AR is 0.03 dB at 1.5 GHz. The 3-dB AR bandwidth is about 18%.

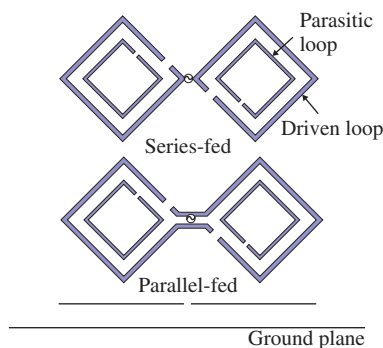


**Figure 3.60** Geometry of a dual-loop CP antenna [91]. Reproduced with permission of © 2004 IEEE

It is also worthwhile to mention that the sense of CP can be easily changed by switching the feed types or gap positions [89]. In addition, the parallel feed technique can be utilized in the dual-loop arrangement. Furthermore, a printed dual-loop antenna is proposed in [90]. The two spiral loops are printed on the opposite sides of a dielectric substrate. They are located symmetrically with respect to the centre of the antenna. A ground plane is attached below the antenna to achieve directional radiation. The lengths of two arms for each loop are properly differentiated to obtain CP radiation, which is similar to the air gaps in the dual rectangular loop antenna presented in Figure 3.60. Also the dual spiral loops are electrically fed by a short dipole. The antenna achieves a 10-dB return loss bandwidth of 6.7% and a 3-dB AR bandwidth of 15%. The size of the antenna is reduced by 24% compared to the antenna in [91].

### 3.7.3 Dual Rhombic Loops with Parasitic Loops

By combining both the methods discussed in the previous sections, the bandwidth of the CP loop antenna can be further increased with the dual loop configuration and parasitic loops [88,92,93]. Figure 3.61 presents the geometry of a series-fed dual rhombic-loop CP antenna which consists of two rhombic loops and two parasitic rhombic loops. The CP operation can be achieved by introducing the small gaps on the loops. A ground plane is applied below the antenna to increase the antenna gain. An optimal AR bandwidth of the dual rhombic-loop antenna can be achieved by optimizing the size of the parasitic loop, the positions of the gaps and the height of the loops above the ground plane. Compared to the series-fed dual rhombic-loop antenna without a parasitic element, the 2-dB AR bandwidth can be increased from 12% to 30% by adding the parasitic loops. In addition to the series-fed arrangement, the dual rhombic-loop antenna can be excited by a parallel-feed, as shown in Figure 3.61. The 2-dB AR bandwidth is enhanced from 25% to 50% by including the parasitic loops. It is also worthwhile to mention that the outer perimeters of the series-fed and parallel-fed dual rhombic-loop antennas are  $\sim 2.8$  and  $\sim 2.2\lambda_0$  ( $\lambda_0$  is the free-space wavelength at the centre frequency), respectively. In [92], a broadband balun is employed to provide the balanced mode for the dual rectangular loop antenna with parasitic loops. Good impedance matching can be obtained due to the coupling between the microstrip line on one side of



**Figure 3.61** Geometry of the series-/parallel-fed dual rhombic-loop CP antenna with parasitic loops [88]. Reproduced with permission of © 2005 IEEE

a thin substrate to the slot which is printed on the other side of the substrate. The antenna in [92] can achieve an impedance bandwidth ( $VSWR \leq 2$ ) of 50%, a 2-dB AR bandwidth of 46% with a gain about 8 dBi. In [93], the series-fed dual-rhombic loop antenna is also fed by a broadband balun and backed by a cylindrical cavity to improve the bandwidth and gain. The cavity-backed antenna can obtain a 3-dB AR bandwidth of 45% and an impedance bandwidth ( $VSWR \leq 2$ ) of 50% with an antenna gain around 11 dBi.

### 3.8 Other Broadband CP Antennas

#### 3.8.1 CP Antennas with Artificial Magnetic Conductors

The work of Yang et al. [94] has shown that, the reflection phase of the AMC varies continuously from  $180^\circ$  to  $-180^\circ$  versus frequency, not only  $180^\circ$  for a perfect electric conductor (PEC) surface or  $0^\circ$  for a perfect magnetic conductor (PMC) surface. This reflection phase characteristic has enabled the designs of broadband CP patch antennas with the AMC as a ground plane [95,96]. In [95], the unit cell of the AMC is a rectangular patch. Broadband circular polarization can be obtained by properly combining the transmitted wave from the square patch antenna and the reflected wave from the AMC structure. The antenna in [95] can achieve a 10-dB return loss bandwidth of 48.6% and a 3-dB AR bandwidth of 20.4%. By contrast, the impedance bandwidth of the antenna is only 5.2% when the AMC structure is removed. The design in [96] has also employed a rectangular patch as a unit cell for AMC structure which is located below the slotted patch. The 10-dB return loss and 3-dB AR bandwidths can reach 11% and 15%, respectively. Another potentially important application of this surface is its usage as the reflector of a spiral antenna for a low profile design, while it leads to a large thickness if a PEC ground is used. The spiral antenna can still remain good CP performance over a broadband frequency range when the AMC surface is located below the antenna with a small distance to achieve the unidirectional radiation [97].

#### 3.8.2 Active CP Antennas

In addition to broadband bandwidth, high efficiency is also a key requirement in modern wireless communications. Hence, it is necessary to design high-efficiency broadband

antennas. The losses in the connecting cables and matching networks can be eliminated when a high-efficiency power amplifier is integrated with the antenna. The designs of high-efficiency CP active antennas have been investigated in [98–100]. The design is obtained by integrating a broadband switching-mode class-E amplifier with a broadband CP slot coupled antenna. The obtained 3-dB AR bandwidth can reach about 9% (1.99–2.18 GHz) and the power added efficiency (PAE) is over 50% over a broadband bandwidth of 14% (1.92–2.21 GHz) [99].

### 3.8.3 CP Antennas for 60-GHz Wireless Communications

The 60-GHz unlicensed frequency band has attracted much attention, as it can provide much capacity of wireless networks. The demands on the 60-GHz systems have stimulated research into the design of the 60-GHz antennas. Accordingly, several broadband CP antennas operating at 60 GHz have been proposed in [101,102]. In [102], the CP on-chip antenna was fabricated using standard 0.18- $\mu\text{m}$  six metal-layer complementary metal-oxide-semiconductor (CMOS) technology. It has dimensions of  $1.8 \times 1.8 \times 0.3 \text{ mm}^3$ . The broadband CP antenna was obtained by utilizing an open-loop structure with a parasitic loop. The performance has been improved by introducing the modified AMC structure on the bottom layer. The obtained 3-dB AR bandwidth ranges from 57 to 67 GHz.

### 3.8.4 CP Antennas with Reconfigurable Polarizations

The broadband CP patch antenna with the capability of electrically switching between RHCP and LHCP has been studied in [103]. The L-probe fed corner-truncated patch antenna can achieve a bandwidth of over 10% with  $\text{VSWR} \leq 2$  and  $\text{AR} \leq 3 \text{ dB}$ . The switching function is realized by a single-pole double-throw (SPDT) switch which is constructed by PIN diodes.

## 3.9 Summary

This chapter has introduced several methods to increase the bandwidth of CP antennas which include patch antennas, PQHAs, slot antennas, DRAs and loop antennas. Monofilar axial mode helix antennas and spiral antennas are inherently broadband CP antennas. In order to compare different types of broadband CP antennas clearly, the performances in terms of bandwidth (covering 10-dB return loss bandwidth and 3-dB AR bandwidth), size, height and fabrication complexity are summarized in Table 3.3. It is shown that cavity-backed spiral and monofilar helix are suitable for CP applications requiring very wide bandwidth (>70%), but they are bulky. Single-feed patch antennas with a thick air substrate are useful for CP applications requiring a bandwidth below 25%, and have advantages of compact size, simple feed structure and low cost. Multi-feed stacked patch antennas with a broadband coupler can achieve a CP bandwidth up to 70%, and have advantages of compact size, easy fabrication and low cost. Other CP antenna technologies, for example, dual-loop antennas with parasitic loops and multi-feed DRAs, are also promising for achieving wideband CP performance up to 50%. Due to limited examples included here, the table is unable to give a whole picture of all CP antennas and it may serve as a rough guide only.

**Table 3.3** Comparisons among different CP antennas

| Ant. Type                  | Broadband methods                        |                                 | Bandwidth (%) | Area Size | Height | Fabrication Complexity |
|----------------------------|------------------------------------------|---------------------------------|---------------|-----------|--------|------------------------|
| Patch antenna              | Single feed                              | Thick air layer [2,4–6]         | 10~25         | Small     | Medium | Medium                 |
|                            |                                          | Slot loading [9]                | ~10           | Small     | Low    | Low                    |
|                            |                                          | Stacked patches [10–16,19]      | 10~30         | Small     | Low    | Medium                 |
|                            |                                          | Coplanar parasitic patches [20] | ~10           | Medium    | Low    | Low                    |
|                            | Multi feed                               | Thick air layer [21–24]         | 20~30         | Small     | Medium | Medium                 |
|                            |                                          | Stacked patches [26,27]         | 20~40         | Small     | Low    | Medium                 |
|                            |                                          | Broadband coupler [28,29]       | 30~70         | Small     | Low    | Medium                 |
|                            |                                          | Coplanar parasitic patches [30] | ~20           | Medium    | Low    | Medium                 |
| Helix                      | Monofilar axial mode [36]                |                                 | ~80           | Small     | Large  | Low                    |
|                            | PQHA                                     | Short with large diameter [40]  | ~30           | Small     | Large  | Medium                 |
|                            |                                          | Tapered PQHA [42]               | ~15           | Small     | Large  | Medium                 |
|                            |                                          | Conical PQHA [43]               | ~20           | Small     | Large  | Medium                 |
|                            |                                          | Parasitic strips [44]           | ~30           | Small     | Large  | Medium                 |
|                            |                                          | Broadband feed [45]             | ~30           | Small     | Large  | Medium                 |
| Spiral antenna             | Cavity-backed spiral [53]                |                                 | >100          | Large     | Medium | Medium                 |
| Slot antenna               | Ring slot with broadband coupler [54]    |                                 | ~20           | Small     | Low    | Medium                 |
|                            | L-shaped feed [55,56]                    |                                 | 20~40         | Small     | Low    | Low                    |
|                            | Grounded strip [57–60]                   |                                 | 10~40         | Small     | Low    | Low                    |
|                            | Shorted ring slot [66]                   |                                 | ~10           | Small     | Low    | Low                    |
|                            | Patch loading [69]                       |                                 | ~10           | Small     | Low    | Low                    |
| DRA                        | Single feed stacked DRA [73,74]          |                                 | 10~20         | Small     | Medium | Medium                 |
|                            | Single-feed notched DRA [76,77]          |                                 | 25~50         | Small     | Medium | Low                    |
|                            | Multi-feed DRA [78–81]                   |                                 | 15~50         | Small     | Medium | Medium                 |
|                            | Multi-mode DRA [82,84]                   |                                 | 25~30         | Small     | Medium | Medium                 |
| Loop antenna (with ground) | Parasitic loop [87]                      |                                 | ~15           | Small     | Medium | Low                    |
|                            | Dual-loop [89,91]                        |                                 | ~20           | Medium    | Medium | Medium                 |
|                            | Dual-loop with parasitic loop [88,92,93] |                                 | ~50           | Medium    | Medium | Medium                 |

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# 4

## Multi-Band Circularly Polarized Antennas

### 4.1 Introduction

Many wireless systems operate at several different frequency bands. For example, modern GNSS systems employ signals at several frequency bands including L1 (1.575 GHz), L2 (1.227 GHz), E5a and E5b bands (1.164–1.215 GHz), and so on. Satellite communication systems also employ different frequency bands for uplink and downlink applications: 6 and 4 GHz are the uplink and downlink frequencies used in C-band satellite communications; 14 and 12 GHz are the uplink and downlink frequencies used in Ku-band; and 30 and 20 GHz are the uplink and downlink frequencies used in Ka-band satellite communications. For these systems, it is possible to employ several single-band CP antennas with one antenna covering each frequency band separately. This, however, will lead to a large size, mass and high cost of antenna and wireless systems. It is far more desirable if a single multi-band CP antenna can be employed for these applications, as it can result in a significant reduction in size, mass, complexity and cost of wireless systems. Thus, multi-band CP antennas have become a hot research topic in recent decades, and a variety of techniques have been developed.

This chapter reviews different techniques for designing multi-band CP antennas, including multi-band CP microstrip patch antennas, multi-band CP slot antennas, multi-band CP DRAs, multi-band QHAs and PQHAs, and so on. Some other antennas such as dual-band dual-sense CP antennas and frequency reconfigurable CP antennas are also discussed. Many examples of antenna designs are illustrated and explained. A table summarizing the performance of some multi-band CP antennas is given at the end.

### 4.2 Multi-Band CP Microstrip Patch Antennas

Similar to the broadband CP patch antennas in Chapter 3, multi-band CP patch antennas can be realized by using single-feed or multi-feed patches. Single-feed design has advantages of a simple feed structure, compact size and low fabrication cost, but the bandwidth is usually narrow and the fabrication tolerance is a challenge. The use of multi-feed patches

can suppress undesired higher-order modes, and has advantages of high polarization purity and broad bandwidth, at the expense of complicated feed network and a large feed size. The following sections will start with a discussion of multi-band CP antenna designs using single-feed patches, and then the multi-band CP antennas using multi-feed patches.

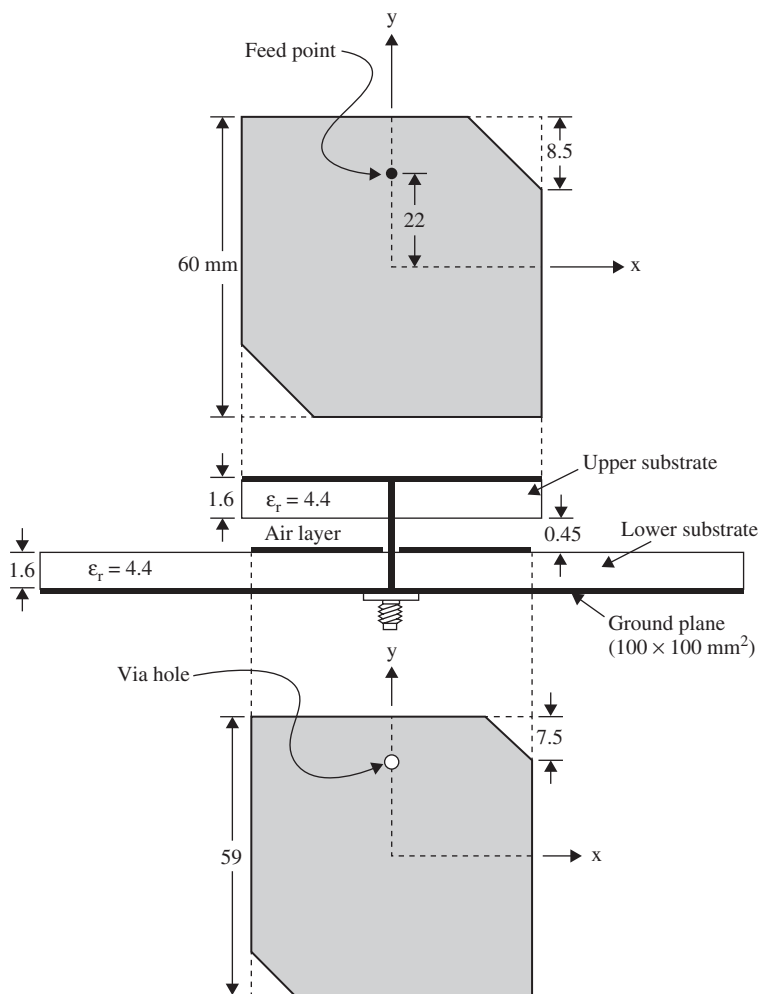
#### 4.2.1 Multi-Band Single-Feed CP Patch Antennas

##### 4.2.1.1 Stacked Patches

It is well-known that stacked patch antennas can be used for achieving dual-band linearly polarized operation. Similarly, multi-band CP antennas can be obtained by utilizing the stacked patch arrangement, as the added patch can provide another resonance. Different feed techniques can be employed. These include aperture-coupled CP stacked-patch antennas [1] and probe-fed CP stacked-patch antennas [2–5]. Figure 4.1 presents the geometry of a probe-fed stacked-patch antenna for achieving CP operations at 1.227 and 1.575 GHz. The antenna consists of two stacked square patches with corners truncated. The patches are printed on FR4 substrates with a thickness of 1.6 mm and a relative permittivity of  $\epsilon_r = 4.4$ . Circular polarization is realized due to the corner-truncated square patches. The upper patch is excited by a probe feed through a via hole in the lower patch. It is also noted that, there is an air layer with a thickness of 0.45 mm between the upper and lower substrates. The frequency ratio of the two operating frequencies can be tuned by varying the thickness of the air gap layer. It is worthwhile to mention that the sizes of two patches mainly depend on the lower operating frequency, which means the resonant lengths of the two patches have only a small difference between them. The resonant length of the upper patch is required to be larger than the lower one for exciting an additional operating frequency (upper). The optimized dimensions of the stacked patch CP antenna are shown in Figure 4.1. The ground plane has a size of  $100 \times 100 \text{ mm}^2$ .

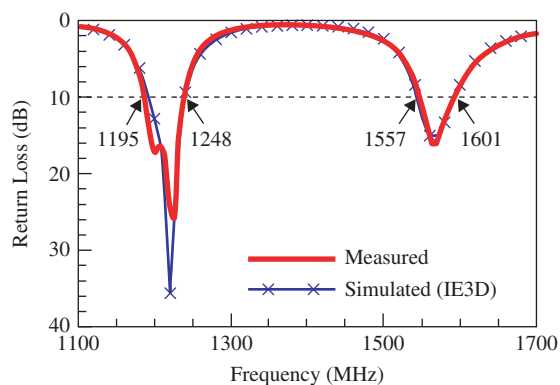
The simulated and measured results of return loss are shown in Figure 4.2 and good agreement between the measurement and simulation results can be observed. The obtained 10-dB return loss bandwidth is 53 MHz in the lower band, or about 4.3% at 1.227 GHz. For the upper band, the obtained impedance bandwidth is 44 MHz, which corresponds to 2.8% at 1.575 GHz. Figure 4.3 presents the result of measured AR. The obtained 3-dB AR bandwidths of the lower and upper bands reach about 15 MHz (1.2%) and 17 MHz (1.1%), respectively. The measured antenna gains at 1.227 and 1.575 GHz are about 1.5 and 4.5 dBi. The measured spinning linear radiation patterns in two principal planes at 1.227 and 1.575 GHz are shown in Figure 4.4. Good broadside CP radiation patterns have been obtained.

Apart from stacked square patches with truncated corners, multi-band CP radiation can be generated by using stacked patches with other shapes. The antenna in [2] consists of two elliptical patches which are printed on separate substrates. Both patches have the same major-axis length but different minor-axis lengths which lead to two different resonant frequencies. The top patch is excited by a probe pin through a via hole in the bottom patch which thus behaves as a parasitic patch. A small air gap is inserted between the substrates to achieve good impedance matching. Circular polarizations of both operating modes are obtained by locating the feed along the  $45^\circ$  line from the major axis, at a proper distance from the centre of top patch.

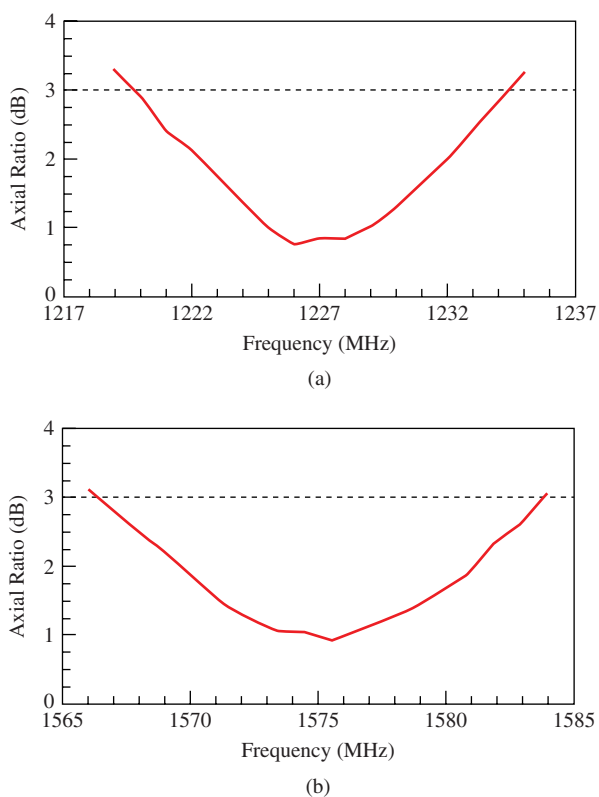


**Figure 4.1** Geometry of a dual-band CP stacked-patch antenna [3]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

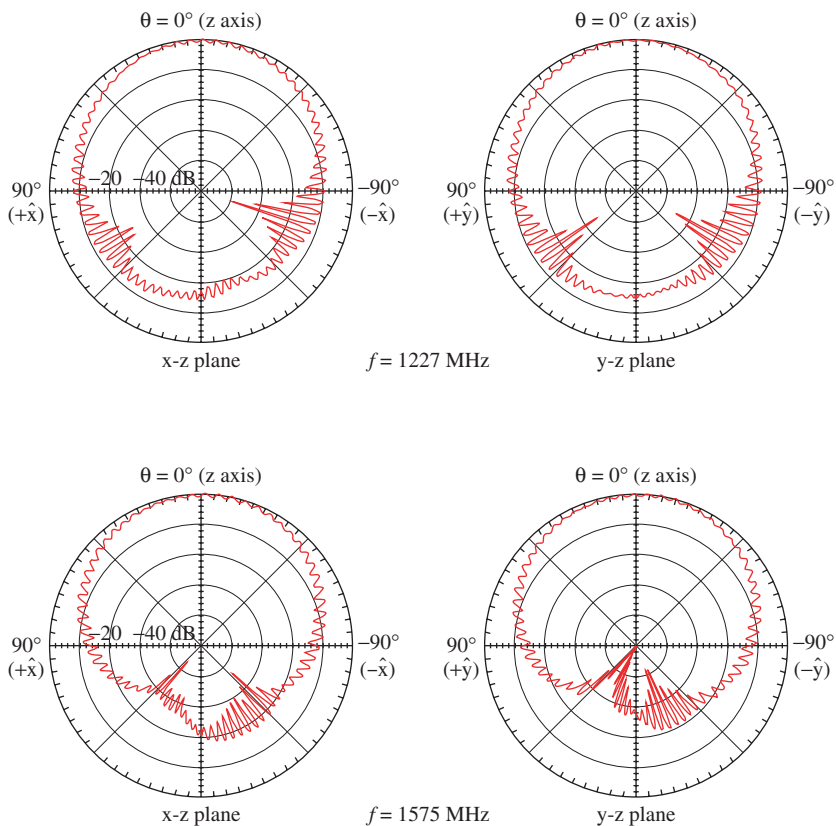
The designs in [2–5] have inserted an air gap between two dielectric substrates. The use of an air gap layer helps achieve good impedance matching at two frequency bands. Such an air gap layer, however, can lead to complexity of antenna fabrication. It is noted that the antenna performance is sensitive to the thickness of the air gap layer, thus accurate fabrication is required. In order to avoid these problems, the air gap is removed in [6]. As shown in Figure 4.5, the air-gap layer is removed and the total height of the antenna is decreased. In this design, the upper patch has a smaller size and resonates at 1.575 GHz, while the lower patch has a larger size and operates at 1.227 GHz. It is also noted that, the upper patch is excited and behaves as a driven patch while the lower patch is parasitically coupled to the top patch.



**Figure 4.2** Simulated and measured return loss for the dual-band CP antenna [3]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

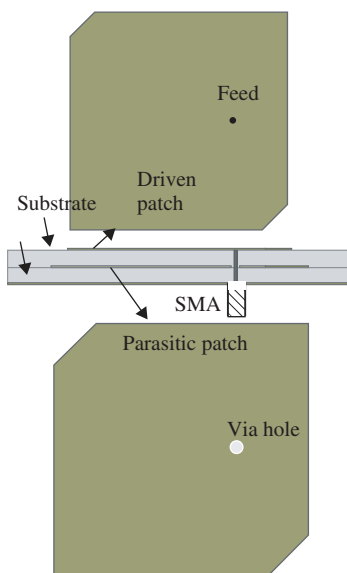


**Figure 4.3** Measured AR in the broadside direction for the dual-band CP antenna [3]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

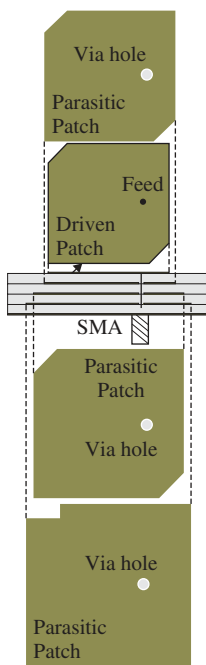


**Figure 4.4** Measured spinning linear radiation patterns in two principal planes at 1.227 and 1.575 GHz [3]. Reproduced with permission of © 2002 John Wiley & Sons, Inc.

In [7], a quad-band CP antenna is achieved by using four stacked patches without an air gap. As shown in Figure 4.6, the antenna consists of four square patches with the lower patch printed on the grounded Rogers RO4003 substrate and other patches printed on the FR4 substrates. The use of different dielectric substrates can reduce the mutual coupling between the lower patch and other patches. A slit cut is embedded in the lower patch while the remaining three patches are perturbed by truncated corners for circular polarization in four bands. The top patch is excited by a probe feed and behaves as a driven patch, whereas other patches are parasitic patches. It is worthwhile to mention that the via holes in the parasitic patches have a capacitive effect which can compensate for the inductive influence caused by the long feed probe. The quad-band stacked patch antenna can operate over 1.163–1.18 GHz, 1.212–1.237 GHz, 1.559–1.596 GHz and 2.277–2.331 GHz. Recently, it is shown in [8] that a dual-band EBG structure can be employed in a dual-band CP patch antenna for improving the performance of antenna gain, AR and impedance bandwidth. After including the EBG structure, the 3-dB AR bandwidths of the lower (1.227 GHz) and upper (1.575 GHz) bands are increased from 13 to 26 MHz and from 13 to 16 MHz, respectively. Besides, the antenna gains at 1.227 and 1.575 GHz are improved from 0.8 to 2.3 dBi, and from 2.4 to 5.4 dBi, respectively.



**Figure 4.5** Geometry of a dual-band CP stacked-patch antenna without an air gap [6]. Reproduced with permission of © 2005 John Wiley & Sons, Inc.

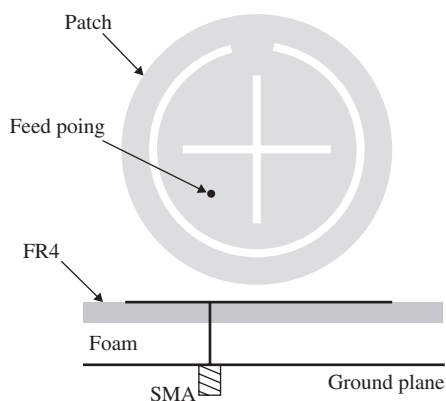


**Figure 4.6** Geometry of a quad-band stacked patch antenna [7]

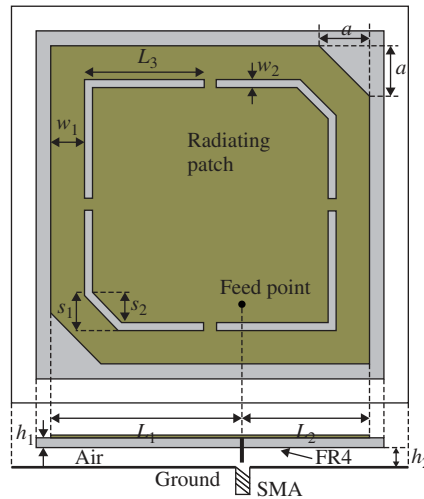
These multi-band CP patch antennas are easy to fabricate and have low cost, but they would be too large in size for applications in small portable terminals. Thus it is necessary to reduce the size of multi-band CP antennas. Some techniques have been proposed, such as the use of folded patches [9], and low temperature co-fired ceramic (LTCC) substrate [10]. In [10], good CP performance of the stacked-patch antenna is realized by adjusting the size of truncated corners and embedded slits of the hexagonal patches. Compared to the multi-band CP patch antenna using an FR4 substrate, the design using a LTCC substrate can achieve a size reduction of over 65%.

#### 4.2.1.2 Slot Loading

The dual-frequency operation of the slot-loaded rectangular patch antenna was studied in [11]. The dual-band operation was obtained by cutting two narrow slots close to the radiating edges. It is found that the resonance of the first operating mode ( $TM_{10}$  mode) is slightly affected while the resonant frequency of the second operating mode ( $TM_{30}$  mode) is significantly decreased due to the slots. Besides, the two modes of the operations show similar radiation properties. Based on this principle, several dual-band CP slot-loaded patch antennas have been investigated in [12–17]. Slot loading can be employed in dual-band patches with various shapes, such as circular patches [12,13] or square patches [14–17]. In the case of circular patches, the two operating modes of  $TM_{11}$  and  $TM_{12}$  are excited by embedding two pairs of arc-shaped slots with proper lengths close to the boundary of a circular patch [12]. The geometry of a dual-band slot-loaded circular patch antenna [13] is shown in Figure 4.7. It is printed on a two-layer substrate formed by an FR4 substrate and a layer of foam for improving the bandwidth. The circular patch is split into an inner patch and an outer patch by an embedded open ring slot for creating dual-band operations. The effective surface current path around the inner patch is slightly shorter than that along the outer patch. Circular polarizations at both bands are obtained by cutting a crossed-slot on the circular patch. The antenna in [13] can achieve a 10-dB return loss bandwidth of 45 MHz in the lower band (or 2.9% at 1.575 GHz). The impedance bandwidth of the upper



**Figure 4.7** Geometry of a dual-band circular patch antenna [13]. Reproduced with permission of © 2010 John Wiley & Sons, Inc.



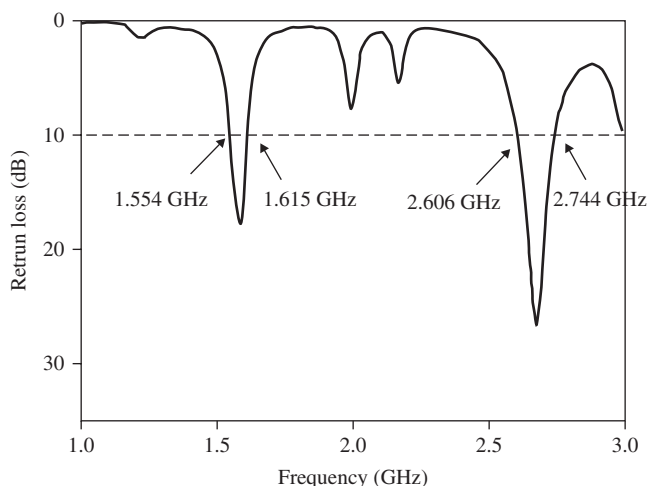
**Figure 4.8** Geometry of the dual-band CP slot-loaded patch antenna [14]. Reproduced with permission of © 2007 John Wiley & Sons, Inc.

band is 46 MHz (or 1.9% at 2.4 GHz). The 3-dB AR bandwidths of the lower and upper bands are 1.75% and 1.65%, respectively.

Figure 4.8 presents the geometry of a dual-band CP slot-loaded square patch antenna. A corner-truncated square patch with a side length of  $L_1 + L_2$  is printed on an FR4 substrate with a thickness  $h_1$  of 1.6 mm and a relative permittivity  $\epsilon_r$  of 4.4 to provide CP operation at 1.5 GHz. It is observed that, there is an air gap with a thickness of  $h_2$  between the ground plane and the substrate. Four slits are embedded in the square patch to achieve another resonance at 2.6 GHz and circular polarization is obtained by selecting the dimensions of the inner truncated corners ( $s_1$  and  $s_2$ ). The probe feed is located on the lower right side from the centre of the patch. The optimal dimensions to achieve dual-band CP operations are:  $L_1 = 38.85$  mm,  $L_2 = 25.05$  mm,  $L_3 = 24.2$  mm,  $a = 9$  mm,  $s_1 = 7.5$  mm,  $s_2 = 6.5$  mm,  $w_1 = 6.55$  mm,  $w_2 = 2$  mm and  $h_2 = 3.5$  mm.

The simulated and measured results of return loss for the presented antenna are shown in Figure 4.9. In the lower band, the obtained 10-dB return loss bandwidth is 61 MHz (3.84%) at 1.5 GHz whereas the corresponding bandwidth of the upper band is 138 MHz (5.15%) at 2.6 GHz. Minimum AR values are 0.28 dB at 1.581 GHz and 0.36 dB at 2.636 GHz, respectively. The measured maximum gains of the dual-band CP antenna are 4.08 dBi at 1.585 GHz and 8.08 dBi at 2.62 GHz, respectively.

The dual-band CP designs in [15,16] are realized by inserting four T-shaped slits at the patch edges, Y-shaped slits at the patch corners or T-shaped slits at the patch corners, respectively. In addition to operations in dual bands, these configurations can achieve very compact sizes, as the operating frequency of the first resonant mode (TM<sub>10</sub> mode) can be significantly decreased by the slits. The design in [15] can achieve a size reduction of 36% compared to a conventional patch antenna without inserting slits. In [16], the corner-truncated square patch antenna employing T-shaped slits with unequal lengths can achieve a dual-band CP performance and a size reduction of 42%. More recently, a dual-band aperture-coupled S-shaped slotted patch antenna can reach a small frequency ratio of 1.28, as investigated in [17]. The

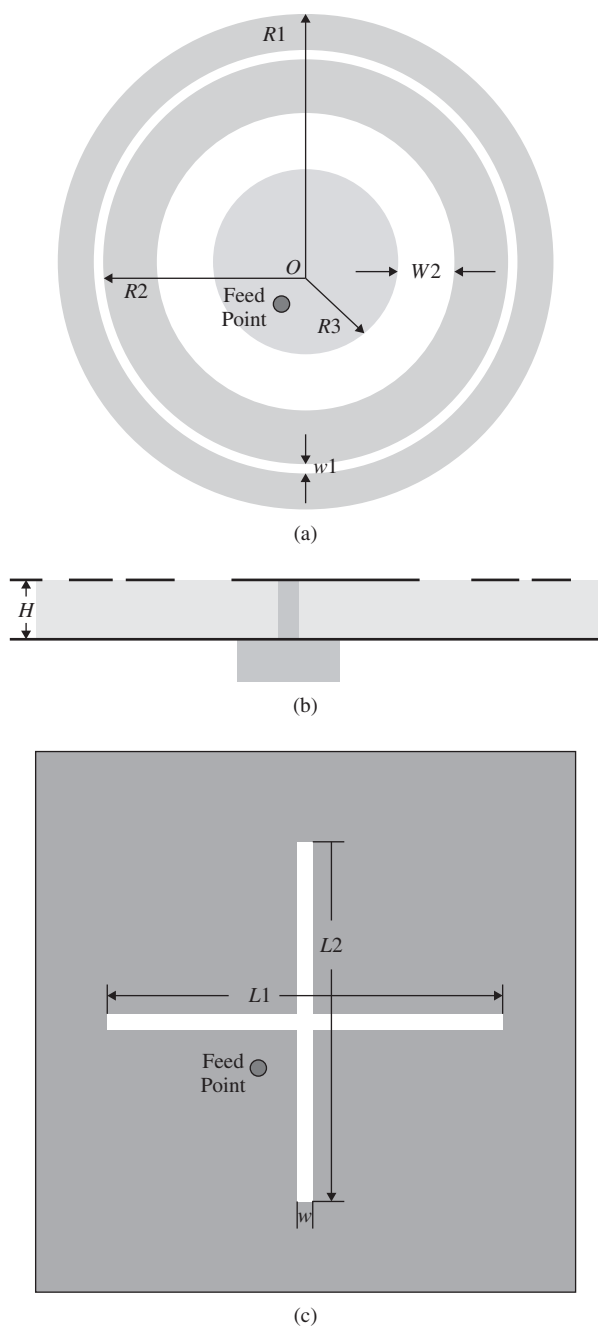


**Figure 4.9** Simulated and measured return loss for the dual-band CP slot-loaded patch antenna [14]. Reproduced with permission of © 2007 John Wiley & Sons, Inc.

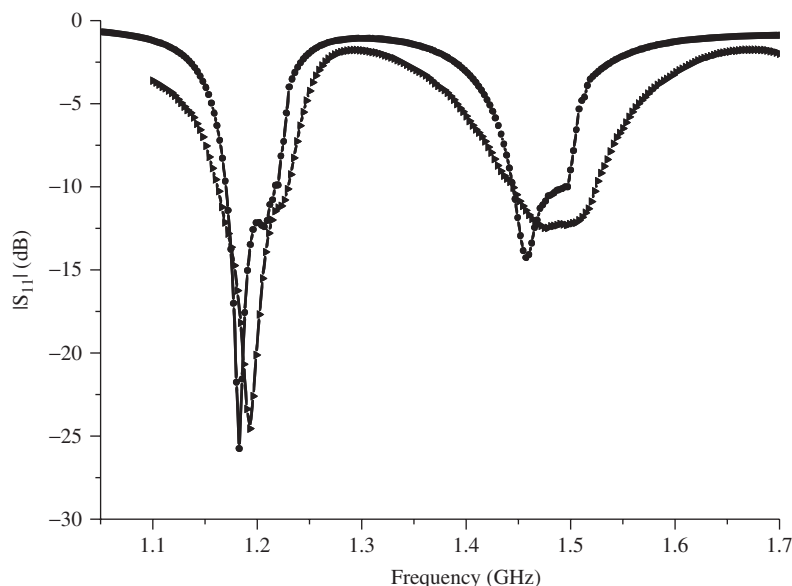
patch is coupled by a microstrip line through the slotted ground plane. Circular polarizations at the lower and upper bands are excited by embedding an asymmetrical S-shaped slot on the square patch. Besides, the S-shaped slot can resonate at the upper band. The measured 10-dB return loss bandwidths of the lower and upper bands are 16% (1.103–1.297 GHz) and 12.5% (1.444–1.636 GHz), respectively. The obtained 3-dB AR bandwidths are 6.9% (1.195–1.228 GHz) and 0.6% (1.568–1.577 GHz) in the lower and upper bands, respectively. The measured gain is over 5 dBic in both frequency bands.

#### 4.2.1.3 Coplanar Parasitic Patches

As discussed Chapter 3, coplanar parasitic patches can be utilized to increase the bandwidth of CP patch antennas. This technique can also be employed to create additional operating frequencies for multi-band CP antenna designs [18]. Figure 4.10 presents the geometry of a dual-band CP annular-ring antenna. The probe-fed antenna is printed on an FR4 substrate with a thickness of  $H = 1.52$  mm and a relative permittivity of 4.0. It consists of two annular rings surrounding a small circular patch on the top and a ground plane with an unequal lateral cross-slot at the bottom. The ground plane has a size of  $60 \times 60$  mm<sup>2</sup>. The lower operating frequency is mainly determined by the larger outer ring radius whereas the resonant frequency of the higher band depends on both the inner ring radius and the separation between the inner annular ring and the inner circular patch. A crossed slot with two slots having the lengths of  $L_1$ ,  $L_2$  and a width of  $w = 1$  mm is embedded in the ground plane. The two rings have the outer radii of  $R_1$ ,  $R_2$  and are separated by a gap  $w_1$ . The circular patch has a radius of  $R_3$  and the gap between the patch and the inner ring is  $w_2$ . The feeding point is located diagonally at  $(-3$  mm,  $-3$  mm) with respect to the centre of the patch. The CP operations at the dual frequency bands can be realized by adjusting the lengths of two slots in the ground plane. The dimensions of the antenna for achieving good performance are:  $R_1 = 24$  mm,  $R_2 = 18.1$  mm,  $R_3 = 6.5$  mm,  $w_1 = 0.8$  mm,  $w_2 = 6.3$  mm,  $L_1 = 40$  mm and  $L_2 = 42.4$  mm.



**Figure 4.10** Geometry of the dual-band CP annular-ring antenna [18]. Reproduced with permission of © 2007 IEEE



**Figure 4.11** Simulated and measured reflection coefficient for the dual-band CP antenna •: simulated result; ►: measured result [18]. Reproduced with permission of © 2007 IEEE

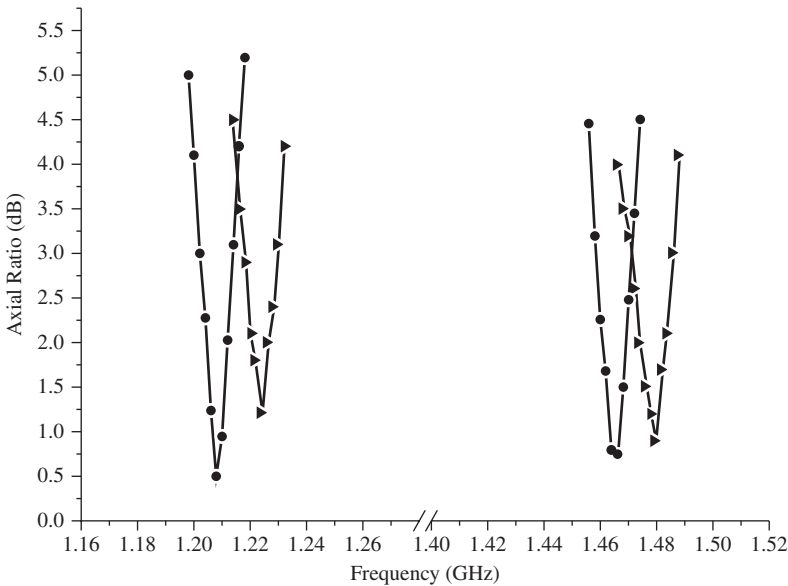
Simulated and measured results of reflection coefficient and AR are shown in Figures 4.11 and 4.12 respectively. The measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) of the lower and upper bands are 72 MHz (5.88% at 1.224 GHz) and 90 MHz (6.08% at 1.48 GHz), respectively. The measured 3-dB AR bandwidths of the lower and upper bands are 12 MHz (1.218–1.23 GHz) and 16 MHz (1.47–1.486 GHz). It is also noted that this antenna has bi-directional radiation due to the slotted ground plane. As shown in Figure 4.13, the LHCP levels at boresight are below  $-22$  dB at 1.21 GHz and  $-15$  dB at 1.467 GHz.

## 4.2.2 Multi-Band Multi-Feed CP Patch Antennas

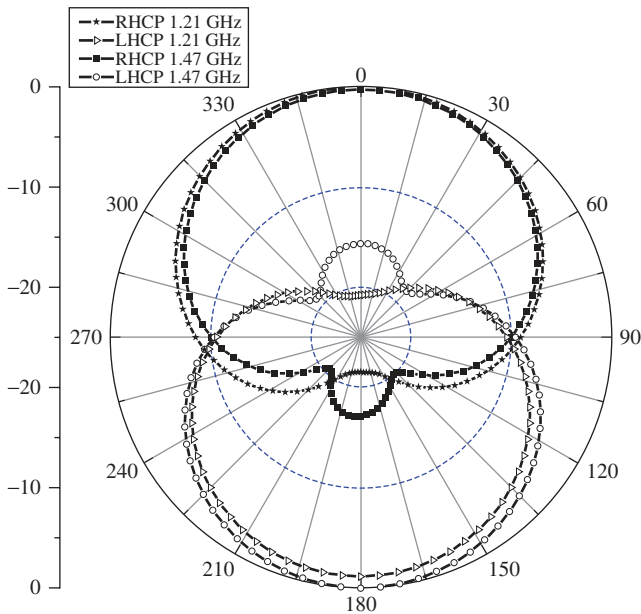
In order to improve the CP performance, multi-feed techniques can be utilized in the multi-band CP antennas. If a dual-feed patch is employed for CP operation, an integrated or external hybrid coupler is required to provide two signals with equal amplitudes but a  $90^\circ$  phase difference.

### 4.2.2.1 Stacked Patches

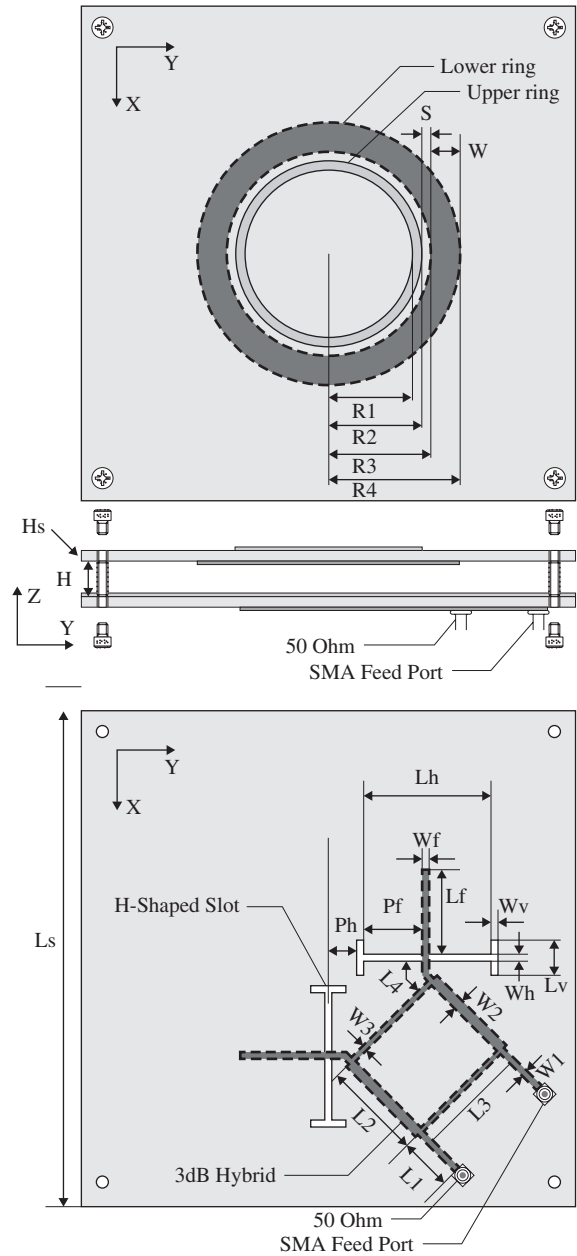
The multi-band multi-feed CP stacked-patch antennas can be realized by integrating a hybrid coupler [19–21] or using multi-port with an external hybrid [22]. Figure 4.14 presents the geometry of an aperture-coupled stacked ring patch antenna [19]. A photo of the fabricated prototype is shown in Figure 4.15. Two concentric annular ring patches are printed on two sides of an FR4 substrate: the small ring is on the top layer and the large ring is on the bottom layer. It is noted that the small ring is for upper band and the large one is for lower band. There is one air layer inserted between the antenna substrate and the ground plane. When



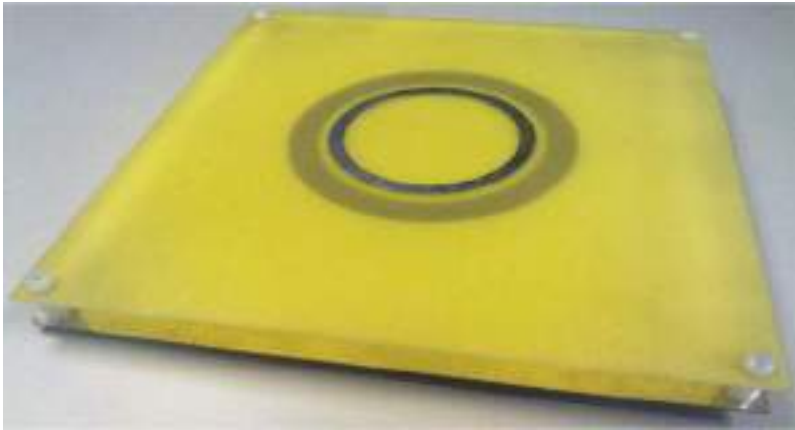
**Figure 4.12** Simulated and measured AR for the dual-band CP antenna •: simulated result; ►: measured result [18]. Reproduced with permission of © 2007 IEEE



**Figure 4.13** Simulated radiation patterns in  $xoz$  plane at 1.21 and 1.467 GHz [18]. Reproduced with permission of © 2007 IEEE



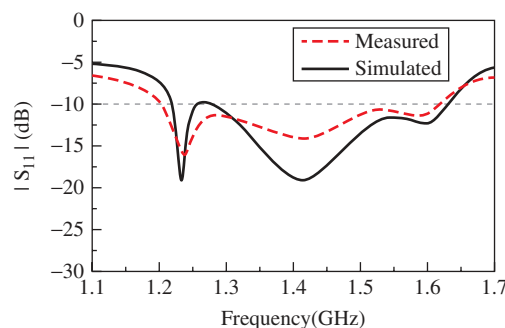
**Figure 4.14** Geometry of the aperture-coupled stacked ring patch antenna [19]. Reproduced with permission of © 2011 IEEE



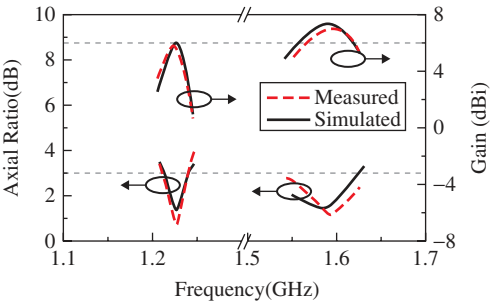
**Figure 4.15** Photo of the fabricated prototype [19]. Reproduced with permission of © 2011 IEEE

considering the aperture-coupled method for achieving circular polarization, two orthogonal H-shaped slots are embedded in the ground plane on the top layer of the feed substrate. A 3-dB microstrip 90° hybrid behaving as the feed network is on the bottom side of the feed substrate. The simulated and measured results of reflection coefficient, AR and antenna gain are shown in Figures 4.16 and 4.17. A wide impedance matching bandwidth from 1.227 to 1.575 GHz has been obtained, which is mainly owing to the broadband performance of the 3-dB hybrid coupler. The 50  $\Omega$  load can absorb the reflected power when the antenna is operating at mismatched frequencies. The AR values are 0.7 and 1.9 dB at 1.227 and 1.575 GHz, respectively, and the 3-dB AR bandwidths of both bands are over 2%. The measured gains at 1.227 and 1.575 GHz are around 6 and 7 dBi, respectively.

The stacked patches of the antenna in [20] are excited by four L-probes which are located sequentially under the lower patch. An integrated feed network consisting of three Wilkinson power dividers is utilized for providing signals with equal amplitudes but a 90° phase shift. The integrated antenna can achieve an impedance bandwidth ( $VSWR \leq 2$ ) of 43.9% and a 3-dB AR bandwidth of 33% in the lower band. For the upper band, the impedance bandwidth ( $VSWR \leq 2$ ) can reach 55.2% and a 3-dB AR bandwidth of 44.7% is obtained. The peak



**Figure 4.16** Simulated and measured reflection coefficient for the dual-band CP antenna [19]. Reproduced with permission of © 2011 IEEE

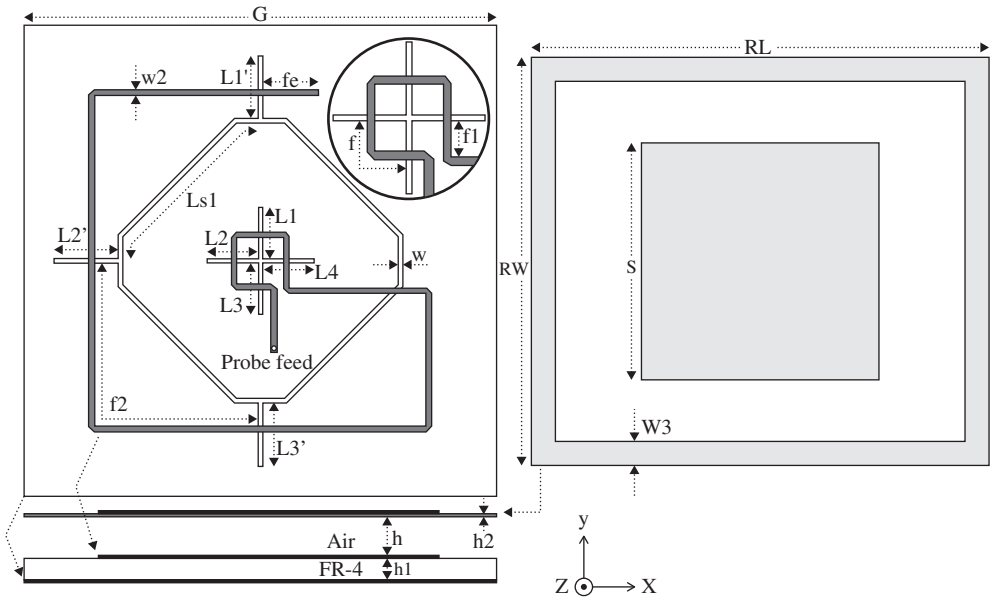


**Figure 4.17** Simulated and measured results of AR and gain [19]. Reproduced with permission of © 2011 IEEE

gains for both bands are around 7.8 dBi. In [21], the aperture-coupled stacked patch antenna is obtained by combining three Wilkinson power combiners and designed for operating at L1 and L2 bands. A proximity-fed stacked patch antenna using two feeding ports is presented in [22] to cover the L1, L2 and L5 bands. A high permittivity dielectric material is used to reduce the size and achieve a compact design.

**4.2.2.2 Coplanar Parasitic Patches**

Figure 4.18 presents the geometry of a serial aperture-coupled dual-band CP antenna. The structure is inherently suitable for dual-band operation due to the combination of a ring and a patch. A serial feed serves as a hybrid feed to excite the patch and the ring for achieving CP



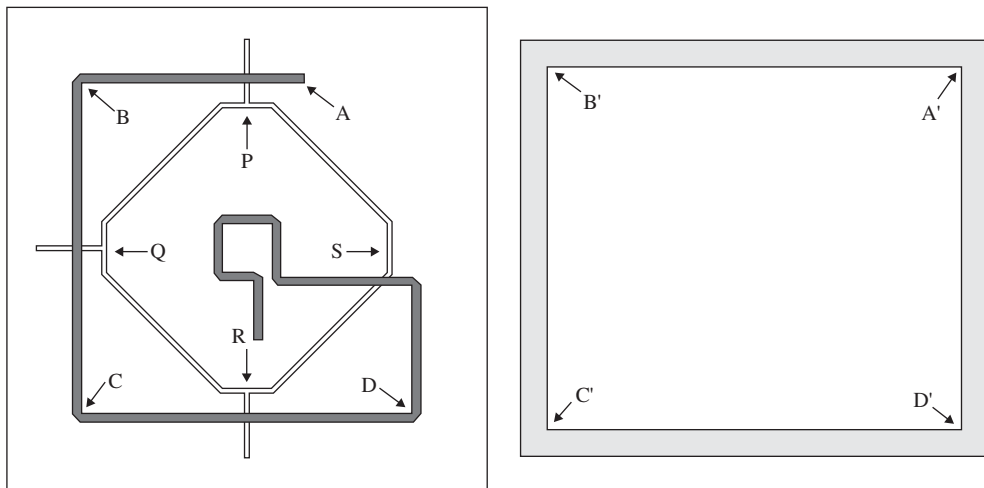
**Figure 4.18** Geometry of the serial aperture-coupled dual-band CP antenna [23]. Reproduced with permission of © 2011 IEEE

operation. Both the ring and patch are printed on the same substrate with the patch inside the ring. The slots are embedded in the ground plane which is on the bottom side of an FR4 substrate. The ring is coupled via three slots, whereas the patch is coupled through a crossed-slot. The extended microstrip line is located on the top side to couple energy to both the patch and the ring through these slots. The serial feed arrangement makes it easy to use a coaxial probe feed near the centre region. Figure 4.19 presents a structure modified from Figure 4.18. In the antenna part, the patch is removed and the ring is excited. In the feed structure part, the crossed-slot is removed whereas the long microstrip line and the diamond-shaped slot and slots L1', L2' and L3' remain unchanged. As observed, the ring is independently excited by the microstrip line. Alternatively, in Figure 4.20, the ring is removed and only the patch is excited. The diamond-shaped slot and three slots L1', L2' and L3' are removed. It is shown that the patch is independently excited by the microstrip line through a crossed-slot.

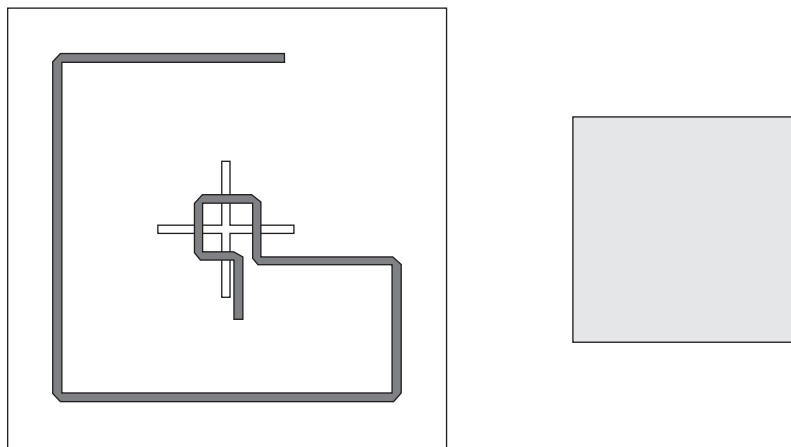
The simulated and measured results of return loss and AR are shown in Figure 4.21. The antenna can operate at both 0.915 and 2.45 GHz with good CP performance. For the lower band, the measured bandwidths for 10-dB return loss and 3-dB AR are 40 and 28 MHz, respectively. For the upper band, the obtained impedance and AR bandwidths are 205 and 155 MHz, respectively. The measured bandwidths can cover RFID operations over both 0.915 and 2.45 GHz bands. Measured peak gains are 3.8 dBic at 0.9 GHz and 9.08 dBic at 2.4 GHz, respectively. It is noted that, good CP performance at 0.915 and 2.45 GHz has been obtained within around  $\pm 30^\circ$  from the boresight direction in the radiation patterns.

#### 4.2.2.3 Meta-Material

Recently, the composite right and left handed meta-material has been employed in CP patch antennas [24,25] for achieving dual-band operations, as it can support the existence



**Figure 4.19** Feed structure (left side) used to independently excite the ring antenna (right side) [23]. Reproduced with permission of © 2011 IEEE

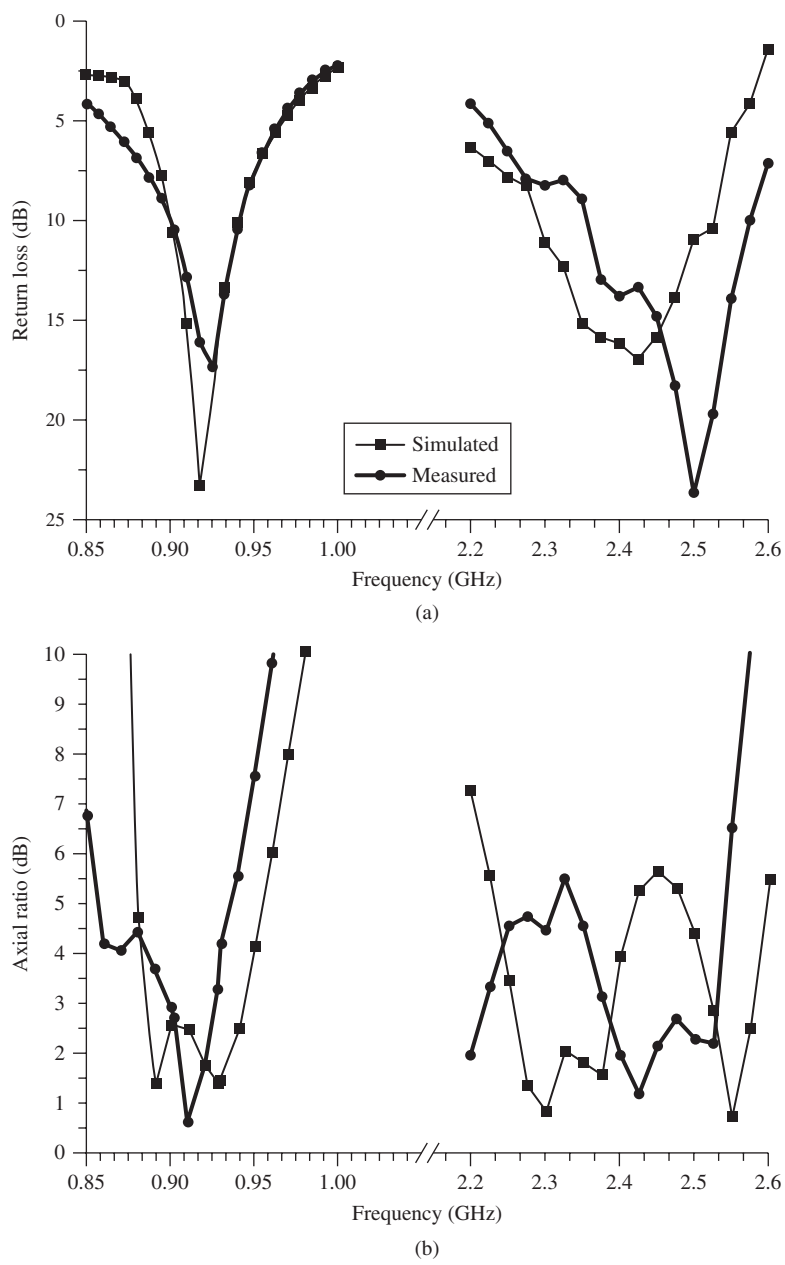


**Figure 4.20** Feed structure (left side) used to independently excite the patch (right side) [23]. Reproduced with permission of © 2011 IEEE

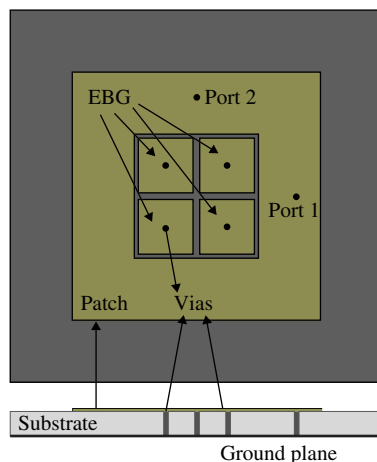
of backward waves, in other words, negative propagation constants. In addition, two modes with the same propagation constant (in the sense of absolute value) have similar radiation properties. As shown in Figure 4.22, a two-port design in [24] is formed by using a square patch antenna with  $2 \times 2$  mushroom-like EBG structures embedded on the radiating patch. The two ports are excited by an external three-arm hybrid coupler [24]. The obtained frequency ratio can be as small as 1.17. The dual-band CP antenna in [25] is realized based on the concept of composite right and left handed meta-materials. Circular polarization is achieved by feeding two vertical ports from external power dividers which can provide signals with equal amplitudes but a  $90^\circ$  phase difference.

#### 4.2.3 Other Multi-Band CP Patch Antennas

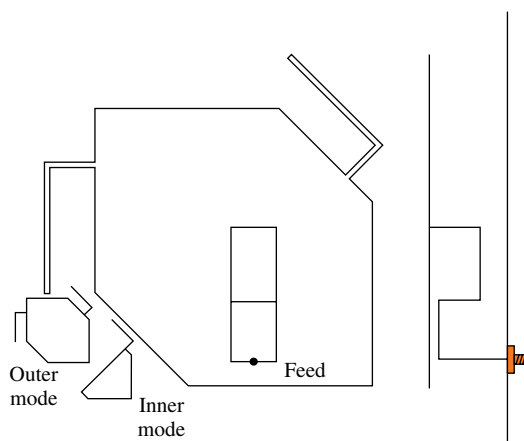
Other methods to obtain multi-band CP patch antennas have also been reported in [26–29]. The dual-band omni-directional CP antenna in [26] consists of three layers of metallization with two back-to-back patches and one middle layer working as the CPW feeding structure. The patches are additionally connected together by a thin copper strip. Circular polarization is obtained by embedding dual slits on both patches. A very low frequency ratio of 1.182 can be realized and it can be easily tuned by adjusting four lumped capacitors included in the antenna. The proximity-fed microstrip antenna in [27] consists of three layers with two nearly square patches locating on top of the first and third dielectric layers. The feedline electromagnetically exciting the patches is in an off-diagonal direction of the middle layer with respect to the two patches. In [28], the dual-band operations of the uni-directional CP patch antenna are obtained by loading a pair of L-shaped stubs outside the truncated patch. As shown in Figure 4.23, a pair of opposite corners is truncated on the square patch. One L-shaped stub is connected with upper right edge while the other one is connected with the left edge. The impedance bandwidth can be improved by using a meander probe having one end connected to the patch and the other end attached to the coaxial cable. Two CP modes can be realized when an outer mode and an inner mode are excited simultaneously. The lower



**Figure 4.21** Simulated and measured results of (a) return loss and (b) AR [23]. Reproduced with permission of © 2011 IEEE



**Figure 4.22** Geometry of a dual-band CP patch antenna with EBG structures [24]. Reproduced with permission of © 2007 EurAAP



**Figure 4.23** Geometry of the corner-truncated patch antenna loaded with two L-shaped stubs [28]. Reproduced with permission of © 2011 IEEE

band at 1.49 GHz has a measured 10-dB return loss bandwidth of 8% (1.43–1.55 GHz) and a 3-dB AR bandwidth of 3% (1.49–1.53 GHz). In the upper band, the impedance bandwidth is 19% (2.09–2.53 GHz) and the 3-dB AR bandwidth is 4% (2.34–2.43 GHz). The measured antenna gain is varying from 8 to 9 dBi in the lower band. For the upper band, the measured antenna gain remains at a stable value of 8 dBi. The summarized radiation pattern performance is shown in Table 4.1. The measured 3-dB beamwidths in both planes are  $65.7^\circ$  at 1.51 GHz. The corresponding values at 2.37 GHz are  $65.2^\circ$  and  $60.2^\circ$ , respectively. The cross-polarization levels are  $-20.9$  dB at 1.51 GHz and  $-17$  dB at 2.37 GHz. The obtained front-to-back ratio is found to be 21.2 and 19.8 dB at 1.51 and 2.37 GHz, respectively.

**Table 4.1** Comparison of the simulated and measured pattern characteristics [28]. Reproduced with permission of © 2011 IEEE

| Frequency | Measurement |           |                      |                   | IE3D simulation |           |                      |                   |
|-----------|-------------|-----------|----------------------|-------------------|-----------------|-----------|----------------------|-------------------|
|           | 3-dB        |           | X-pol.<br>Level (dB) | F/B ratio<br>(dB) | 3-dB            |           | X-pol.<br>Level (dB) | F/B ratio<br>(dB) |
|           | beamwidth   |           |                      |                   | beamwidth       |           |                      |                   |
|           | Phi = 0°    | Phi = 90° |                      |                   | Phi = 0°        | Phi = 90° |                      |                   |
| 1.51 GHz  | 65.7°       | 65.7°     | −20.9                | 21.2              | 67.0°           | 69.5°     | −24.5                | 20.1              |
| 2.37 GHz  | 65.2°       | 60.2°     | −17.0                | 19.8              | 68.5°           | 57.5°     | −20.2                | 21.8              |

More recently, a compact low multipath cross-plate reflector ground plane has been proposed and integrated with a dual-band CP antenna for high precision GNSS applications [29]. The proposed antenna can achieve a front-to-back ratio over 25 dB at L1 and L2 bands. The maximum backward cross-polarization levels are below −23 dB in both frequency bands. Antenna phase centre variation remains less than 2 mm at L1 and L2 bands. Compared to the choke ring and EBG ground planes, the cross-plate reflector ground plane has a compact size, low mass, wide operating frequency bandwidth and simple configuration.

### 4.3 Multi-Band QHAs and PQHAs

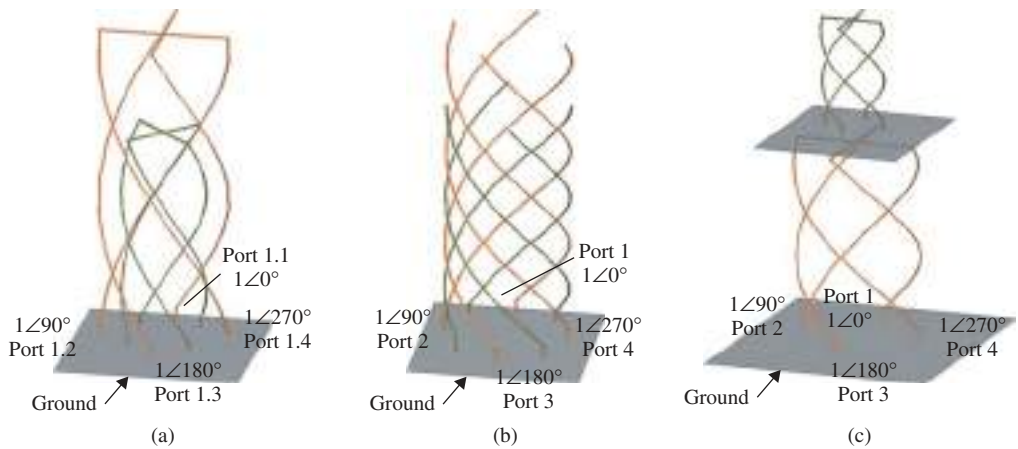
#### 4.3.1 Multi-Band QHAs

##### 4.3.1.1 Dual QHAs

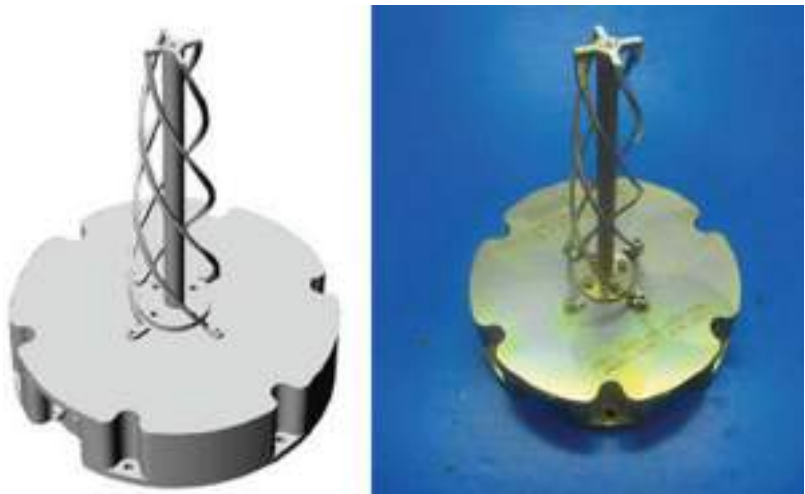
The dual-band operation of QHA can be achieved by incorporating two antennas into a single structure by coaxially mounting them together [30]. The two antennas operate at two different frequencies. The arrangements can be classified into three fashions: 1. ‘Piggyback’ (one antenna is on the top of the other one), as shown in Figure 4.24(a); 2. ‘Enclosed’ (one antenna is inside of the other one), as shown in Figure 4.24(b); and 3. ‘Rotational offset’ (the arms of the two antennas are interleaved), as shown in Figure 4.24(c), the two antennas have the same diameter but different arm length. A tri-band QHA operating at the L1, L2 and L5 bands has been studied in [31]. It is realized by combining three antennas in Piggyback and Rotational offset fashions. A dual-band broadband QHA using two QHAs in Piggyback fashion is integrated with a compact power divider, as investigated in [32]. The antenna is designed for operating at 1.615 and 2.492 GHz with the 10-dB return loss bandwidths of 39% and 28%, respectively. The obtained AR is less than 1.5 dB in both lower and upper frequency bands of the system (Uplink:  $1615.68 \pm 7$  MHz; Downlink:  $2491.75 \pm 15$  MHz).

##### 4.3.1.2 Hybrid Coupler

Figure 4.25 shows a compact dual-band conical-shaped QHA for satellite applications [33]. It is fed by a hybrid coupler. The geometrical model of the QHA consists of a cylinder as the base and four helical strips attached to the topside of the cylinder. The opposite strips are short-circuited at the top. The feeding circuit is located inside the cylinder to provide the signals with equal amplitudes but a 90° phase shift between each port. The feeding circuit is



**Figure 4.24** Structures of dual-band QHAs with two antennas

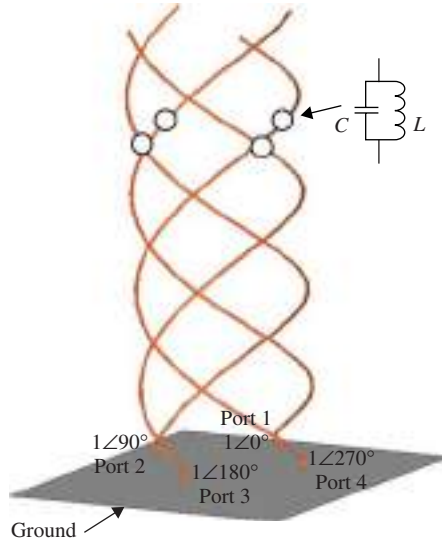


**Figure 4.25** Prototype of a dual-band conical QHA integrated with a broadband hybrid coupler [33]. Reproduced with permission of © 2012 IEEE

formed by a  $180^\circ$  hybrid ring and two  $90^\circ$  hybrid branch-line couplers. In order to increase the bandwidth, the couplers have two quadrature sections. The results in [33] have shown that the antenna operates at S-band and is suitable for satellite telemetry, tracking and control (TTC) applications.

**4.3.1.3 Lumped Element Loading**

Figure 4.26 presents the geometry of a dual-band QHA loaded with lumped elements. Each arm has one turn and is open at one end with the feed point at the other end. The four arms



**Figure 4.26** Geometry of a dual-band QHA loaded with lumped elements [34]. Reproduced with permission of © 2002 IEEE

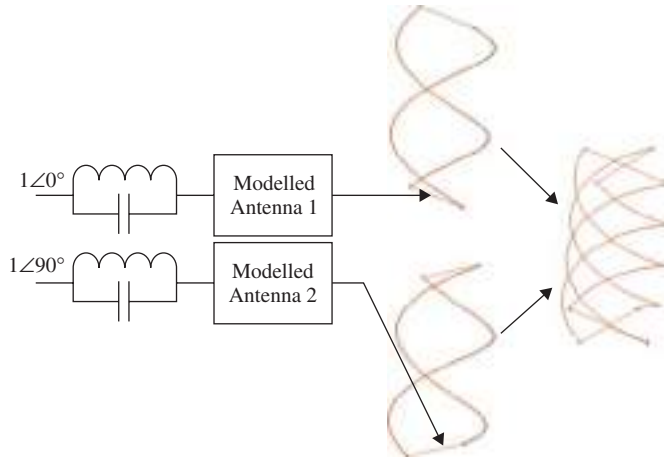
are excited in quadrature for obtaining circular polarization. The lower operating frequency is determined by the length of the helix. In order to generate a higher frequency band, a trap circuit is located at a proper location in each arm. It comprises a parallel LC circuit which has infinite impedance at the resonant frequency. At the lower frequency, the low reactive impedance introduced by the trap circuit can be compensated by a slight change in the overall length of the helix.

In [35], the QHA can operate in dual separate bands over 250–270 MHz and 290–310 MHz by changing the electrical length of the helix. PIN diodes are located at proper locations on the radiating helix. When shortening the antenna for the upper band is desired, the diodes are biased short circuiting segments of the antenna. On the other hand, diodes are unbiased to make the antenna work in the lower band.

The work in [36] has presented the design of a dual-band QHA with a very simple impedance matching network at two narrow frequencies. As shown in Figure 4.27, the QHA is formed by four coaxial helices placed in 90° intervals. The opposite helices form one branch with the top ends shorted and have one input port. Hence, the antenna has two similar input ports. The 90° phase difference is provided by a phase shifter. The input impedance after inserting the impedance matching network is given as [36]:

$$Z_{in} = (R_{in} + jX_{in}) + \frac{j}{1/\omega L - \omega C} = R_{in} + j \left[ X_{in} + \frac{j}{1/\omega L - \omega C} \right] \quad (4.1)$$

where  $R_{in}$  and  $X_{in}$  are the real and imaginary parts of the antenna input impedance at  $\omega$  respectively, and  $L$  and  $C$  are values of the inductor and capacitor of the matching network. As the real part of the antenna input impedance has been adjusted to be 50  $\Omega$  at both  $\omega_1$  and  $\omega_2$ , only the reactance of  $Z_{in}$  needs to vanish at both  $\omega_1$  and  $\omega_2$  to achieve optimal impedance



**Figure 4.27** Geometry of a dual-band QHA with an impedance matching network [36]. Reproduced with permission of © 2005 IEEE

matching. The optimal values of  $L$  and  $C$  can be extracted as follows [36]:

$$L = \frac{1 - \left(\frac{\omega_2}{\omega_1}\right)^2}{\omega_2 \left[ \frac{\omega_2}{\omega_1 X_{in1}} - \frac{1}{X_{in2}} \right]} \quad (4.2)$$

$$C = \frac{\frac{1}{L} + \frac{\omega_1}{X_{in1}}}{\omega_1^2} \quad (4.3)$$

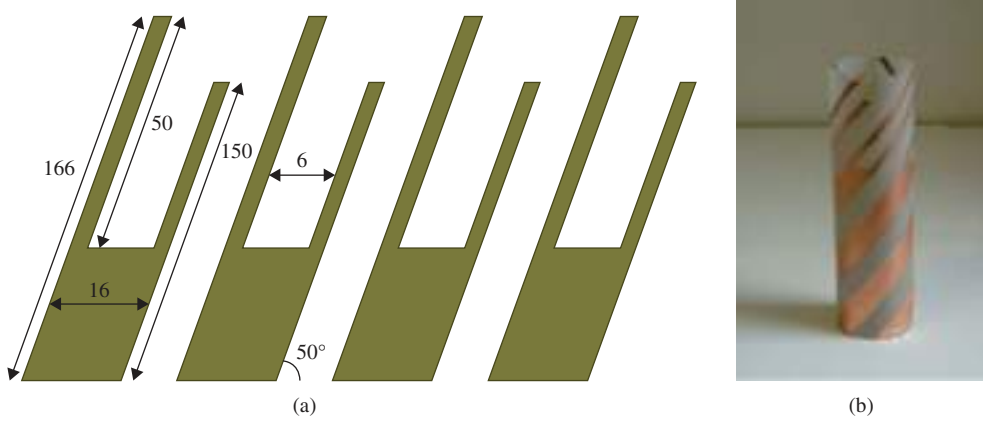
where  $X_{in1}$  and  $X_{in2}$  are the reactance values of the antenna input impedances at  $\omega_1$  and  $\omega_2$ , respectively. Thus, the required values of  $L$  and  $C$  for the optimum matching work can be easily obtained by using 4.2 and 4.3.

### 4.3.2 Multi-Band PQHAs

In order to ease the fabrication of the QHA, the helices can be printed on a thin substrate sheet. Various designs of multi-band PQHA will be introduced in this section.

#### 4.3.2.1 Multiple Fingers

For the ‘Rotational offset’ fashion in Figure 4.24(c), the antennas can be printed on a single thin substrate sheet [37] as they have the same diameter. This will ease the complexity of the fabrication. Figure 4.28 presents the unwrapped geometry and the fabricated prototype of a dual-band PQHA [37]. It is formed by four helix-shaped elements which are printed on a substrate with a thickness of 0.127 mm and a relative permittivity of 2.2, and wrapped

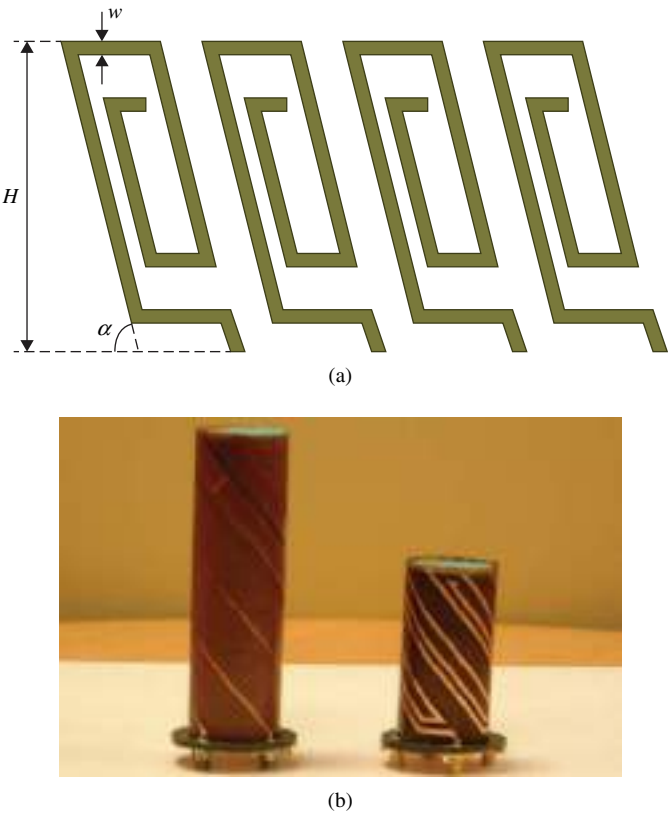


**Figure 4.28** (a) Geometry of the dual-band PQHA and (b) Fabricated prototype [37]. Reproduced with permission of © 2010 IEEE

around a cylindrical support. The antenna has a circumference of 113 mm and the number of turns is 0.9. Each helix consists of a wide arm and its end is divided into multiple parallel thin arms. Dual frequency operations can be obtained by adjusting the length of each thin arm. The measured result demonstrates that the antenna can operate at 1.25 and 1.42 GHz. The obtained bandwidths of the lower and upper bands defined by  $VSWR \leq 2$  are 7.2% and 3.5%, respectively. The antenna is excited by a commercial hybrid coupler. Stable radiation patterns with large beamwidths and good circular polarization can be observed at both frequencies. Compared to the co-polarizations, the cross-polarization levels are less than  $-15$  dB. The measured antenna gains are more than 2 dBi over both frequency bands.

#### 4.3.2.2 Folded PQHAs

In order to reduce the volume of PQHA, folded PQHAs have been studied in [38–40]. Compared to the traditional PQHA, the folded PQHA achieves a compact size. Moreover, additional resonant frequencies can be induced due to the mutual coupling between the folded arms of each helix. Figure 4.29 presents the geometry of the unwrapped dual-band folded PQHA and the fabricated prototype. It is formed by four helix-shaped elements which are printed on a 0.127-mm-thick dielectric substrate with a relative permittivity of 2.2 and wrapped around a cylindrical support. The arm of each helix is meandered and folded into the form of square spirals to increase the effective length, thus reducing the overall height of the antenna. The values of the geometrical parameters for the conventional and folded PQHAs are given in Table 4.2. As can be observed, the height of the antenna has been reduced by 43%. The simulated results of active return loss for both antennas are compared in Figure 4.30, in which the solid line is the folded PQHA and the dashed/dotted line represents the conventional PQHA. The multi-resonance behaviour is expected. Compared to the conventional PQHA, the folded PQHA has four resonant frequencies at 1.26, 1.85, 2.2 and 2.93 GHz. In the presented work, only the first two resonant frequencies have been studied. The measured 10-dB return loss bandwidths are 1.69% at 1.24 GHz and 4.86% at 1.8 GHz, respectively.

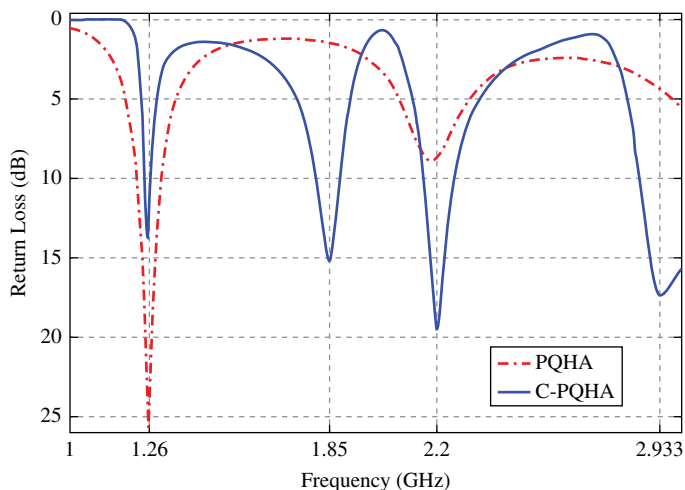


**Figure 4.29** (a) Geometry of the dual-band folded PQHA and (b) Fabricated prototype [39]. Reproduced with permission of © 2010 EurAAP

**Table 4.2** Geometrical parameters for conventional and folded PQHAs [39]. Reproduced with permission of © 2010 EurAAP

| Name               | Conventional PQHA | Folded PQHA |
|--------------------|-------------------|-------------|
| $H$ (mm)           | 127.16            | 72.48       |
| $L_c$ (mm)         | 166               | 302.66      |
| $R$ (mm)           | 18                | 18          |
| $\alpha(^{\circ})$ | 50                | 50          |
| $w$ (mm)           | 2                 | 2           |
| $N$                | 0.75              | 0.75        |

$R$  is the radius of the cylindrical support;  $L_c$  is the length of the helix;  $N$  is the number of turns.



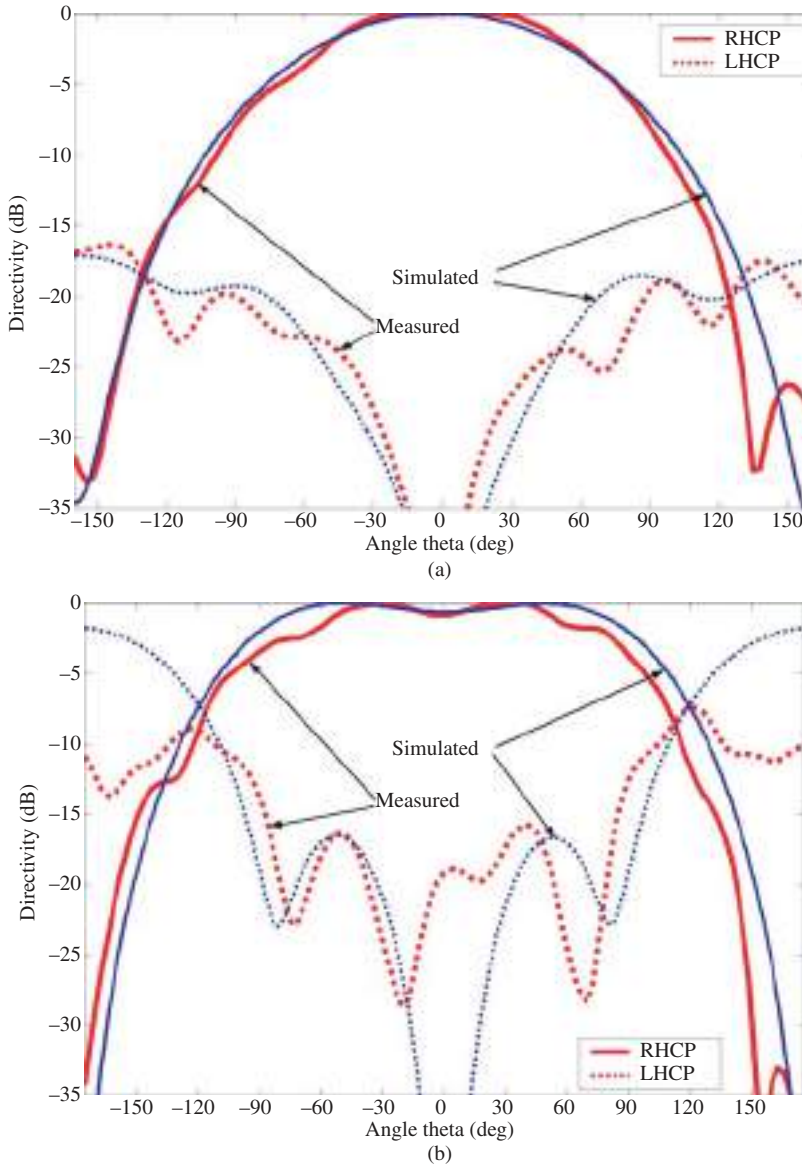
**Figure 4.30** Comparison of simulated return loss between conventional and folded PQHAs [39]. Reproduced with permission © 2010 EurAAP

The antenna is excited by a commercial hybrid coupler. The simulated and measured radiation patterns at 1.24 and 1.8 GHz are shown in Figure 4.31. The simulated radiation patterns display almost identical characteristics to the measured patterns at two frequencies. The measured 3-dB beamwidths at 1.24 and 1.8 GHz are  $120^\circ$  and  $196^\circ$ , respectively. The cross-polarization levels are much lower than the co-polarizations at boresight. The measured gains at 1.24 and 1.8 GHz reach around 6.2 and 3.75 dBic, respectively.

In order to further reduce the size of the PQHA, a miniaturized PQHA is developed in [41] with the combination of meander line technique and dielectrically loading technique. Four helix-shaped radiating elements are printed on a thin dielectric substrate, wrapped around a cylindrical dielectric support and mounted on a small ground plane. Each helix is formed by meandering and turning the helix arms into the form of square spirals. By introducing a dielectric rod with a relative permittivity of 10, the size of the antenna is significantly reduced. The presented antenna in [41] can operate over L1 and L2 bands with a size reduction of around 73% compared to the conventional PQHA. The measured bandwidths of the compact PQHA are 0.66% at 1.17 GHz and 0.5% at 1.54 GHz, respectively. Measured antenna gains are 2.22 dBic at 1.17 GHz and 1.15 dBic at 1.54 GHz, respectively.

#### 4.3.2.3 Square PQHAs with Folded Inverted-F Monopoles

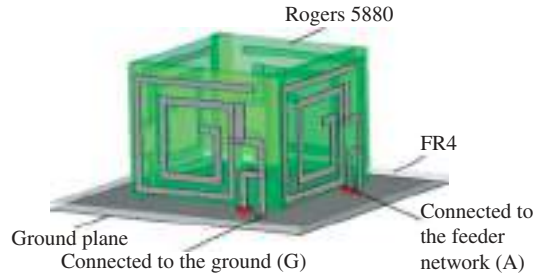
In [42], the folded inverted-F antenna has been used as a helix to realize QHA. Accordingly, dual-band PQHAs with folded inverted-F antennas have been investigated in [43,44]. Figure 4.32 presents the geometry of a dual-band PQHA which is mounted above an FR4 substrate board. The feed network is printed on the top of the FR4 substrate and the ground plane is on the opposite side. Each helix is formed by two folded inverted-F antennas and fabricated on the Duroid 5880 substrate. The dual-frequency operations can be achieved by adjusting the lengths of the two loops. The lower operating frequency is determined by the longer loop, whereas the higher resonant frequency depends on the shorter one. In order to



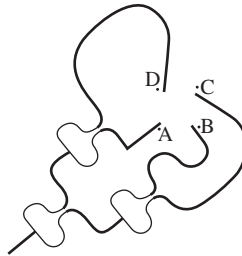
**Figure 4.31** Simulated and measured radiation patterns for the folded PQHA at (a) 1.24 GHz and (b) 1.8 GHz [39]. Reproduced with permission © 2010 EurAAP

improve the impedance matching of the antenna, the outer vertical metal strip is connected to the ground through a via and the inner vertical metal strip is connected to the output of the feed network directly. The layout of the feed network for the antenna is depicted in Figure 4.33. As observed, the feed network can provide four signals with equal amplitudes but quadrature phase. The presented arrangement can achieve RHCP radiation.

The simulated current distributions at 1.227 and 1.575 GHz are illustrated in Figure 4.34 to further understand the operating principle of the antenna. It is found that the current



**Figure 4.32** Configuration of the dual-band QHA formed by inverted-F monopoles [43]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.



**Figure 4.33** Layout of the feed network for the antenna [43]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.

distributions are mainly concentrated on the shorter loop when the antenna is operating at the L1 band. At the L2 band, the current distribution on the longer loop is stronger than that on the shorter one. The phenomenon indicates that the dual frequencies can be controlled by adjusting the lengths of the two loops.

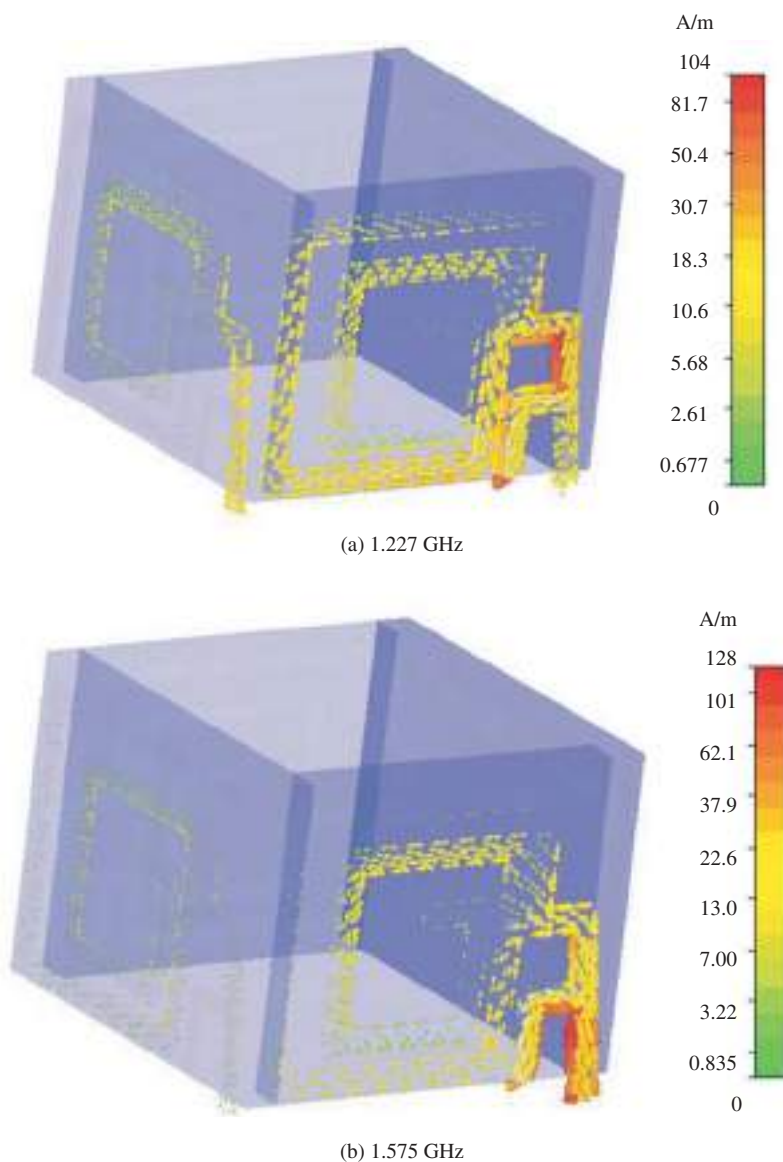
It is noted that the frequency range of the integrated antenna for  $|S_{11}| \leq -10$  dB is covering both the L1 and L2 bands. In terms of the radiation patterns, the cross-polarization levels are less than  $-10$  dB compared to the co-polarizations at both frequencies. The AR values at boresight are 2.8 dB at 1.227 GHz and 1.6 dB at 1.575 GHz. The antennas gains are found to be 1.1 dBi at 1.575 GHz and 0.6 dBi at 1.227 GHz, respectively.

In [44], the dual-band PQHA antenna comprises two PQAs concentrically arranged, but electrically isolated. Each PQA is formed by four inverted-F monopoles arranged orthogonally and excited in quadrature. The compact dual-band PQHA can operate at UHF RFID (902–908 MHz) and GPS (1575 MHz), thus it is suitable for applications in a portable terminal.

## 4.4 Multi-Band CP Slot Antennas

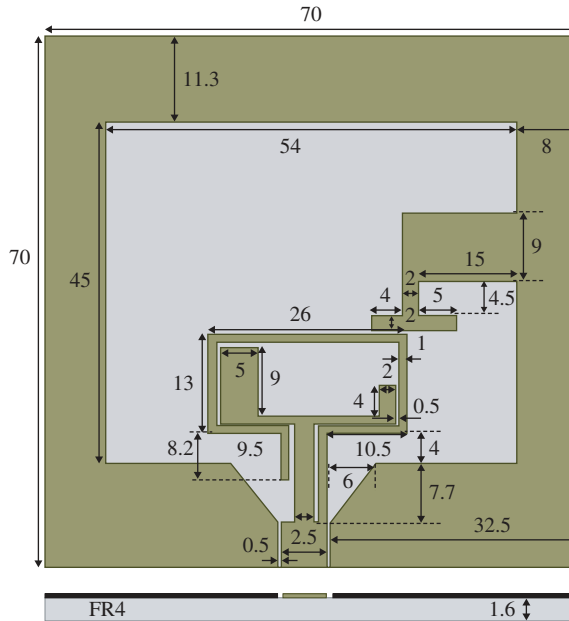
### 4.4.1 Dual-Monopole

In [45], a CPW-fed dual-band antenna with linearly polarized operation is obtained by utilizing two monopoles. The two operating modes of the antenna are associated with the



**Figure 4.34** Simulated current distributions of the antenna [43]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.

lengths of two monopoles, in which the lower operating frequency is determined by the longer monopole and the higher operating frequency depends on the shorter monopole. Based on this operating principle, multi-band CP slot antennas have been investigated in [46,47]. This type of antenna is formed by two monopoles and a slot antenna. Figure 4.35 presents the geometry of a dual-band CP slot antenna. The CPW-fed antenna consists of two deformed parallel monopoles in the feedline and a deformed ground plane. The longer



**Figure 4.35** Geometry of the dual-band CP slot antenna (Unit: mm) [46]. Reproduced with permission of © 2011 IEEE

curved monopole is surrounding the shorter fork-shaped monopole. The antenna with a size of  $70 \times 70 \text{ mm}^2$  is printed on an FR4 substrate with a thickness of 1.6 mm and a relative permittivity of 4.4. A crane-shaped strip is protruded from the slotted ground plane to generate CP fields. The optimized dimensions of the antenna are shown in Figure 4.35.

The measured impedance bandwidths ( $|S_{11}| \leq -10 \text{ dB}$ ) of the lower and upper bands are 17% (1.45–1.72 GHz) and 21% (1.86–2.29 GHz), respectively. The obtained 3-dB AR bandwidths are 9% (1.47–1.61 GHz) for the lower band and 11% (1.87–2.09 GHz) for the upper band, respectively. The obtained frequency ratio of the upper band to the lower band is 1.286.

The simulated and measured radiation patterns in [46] have shown that the cross-polarization levels are 15 dB lower than the co-polarizations in the broadside direction. Stable radiation patterns at four frequencies (1.52, 1.60, 1.92 and 2.04 GHz) have been obtained and the 3-dB beamwidths are around  $80^\circ$ . It is worthwhile to mention that it has bidirectional radiation and the radiation patterns in both sides have opposite polarizations and look similar. The antenna gain varies from 2.5 to 4 dBi within the dual bands.

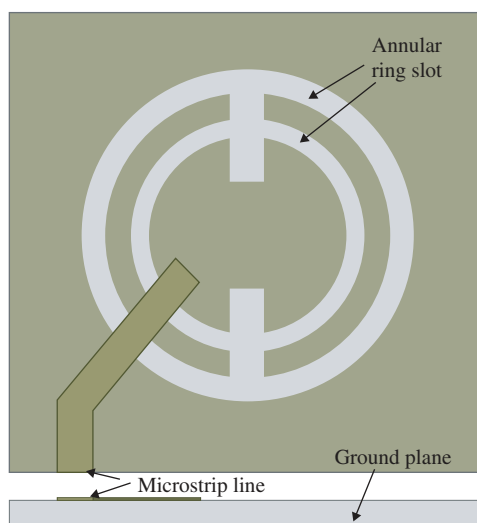
Moreover, a triple-band CP slot antenna with two monopoles has been studied in [47]. The CPW-fed square slot antenna consists of dual monopole feed, crooked F- and T-shaped strips. The design in [47] has the capability of operating over 1.96–3.26 GHz, 3.61–6.98 GHz and 7.87–11.24 GHz, while it can provide circular polarization for the second band.

### 4.4.2 L-Shaped Feed

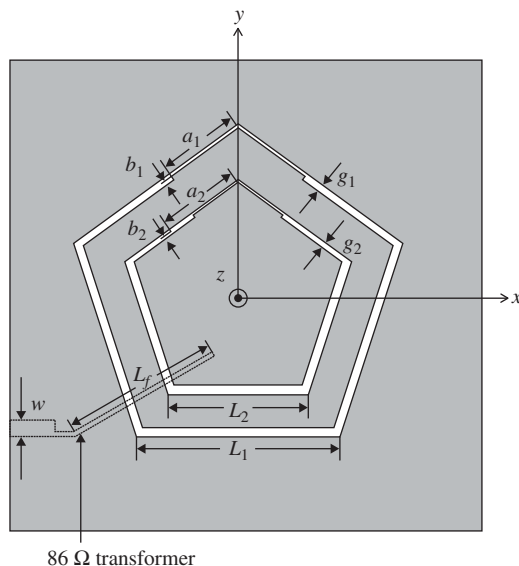
A microstrip-fed dual-band CP slot antenna with a broad bandwidth is presented in [48]. A thin microstrip line behaves as an impedance transformer between the tapered end of the L-shaped strip and the  $50\Omega$  microstrip-fed line. Similar to the discussions in Section 3.5.2, initially the broadband CP operation can be obtained by using an L-shaped feed. To generate an extra resonant mode, an L-shaped strip is connected to the L-shaped tuning stub. Besides, a narrow slit and an L-shaped slot are embedded in the L-shaped tuning stub to obtain better impedance matching and CP properties. The microstrip-fed slot antenna can achieve an impedance bandwidth ( $VSWR \leq 2$ ) of 21% and 3-dB AR bandwidth of 18.9% at 3.5 GHz in the lower band. For the upper band (5–6 GHz), the obtained impedance bandwidth ( $VSWR \leq 2$ ) and 3-dB AR bandwidth are 33% and 32.5%, respectively.

### 4.4.3 Concentric Ring Slots

Figure 4.36 presents the geometry of a dual-band annular-ring slot antenna which consists of two concentric annular-ring slots embedded at the centre of the ground plane. It is indicated that, when the distance between outer and inner ring slots increases, or if the frequency ratio  $f_H/f_L$  is larger than a certain value, the resonant frequency in the upper band will disappear. To overcome this shortcoming, a pair of rectangular notches are employed and located symmetrically between the two ring slots. Furthermore, circular polarization is obtained by adjusting the angle between the notches and the microstrip line. In this design, the angle between the notch and the feedline is selected as  $45^\circ$ . It is found that the lower band is centred



**Figure 4.36** Geometry of a dual-band annular-ring slot antenna [49]. Reproduced with permission of © 2009 John Wiley & Sons, Inc.

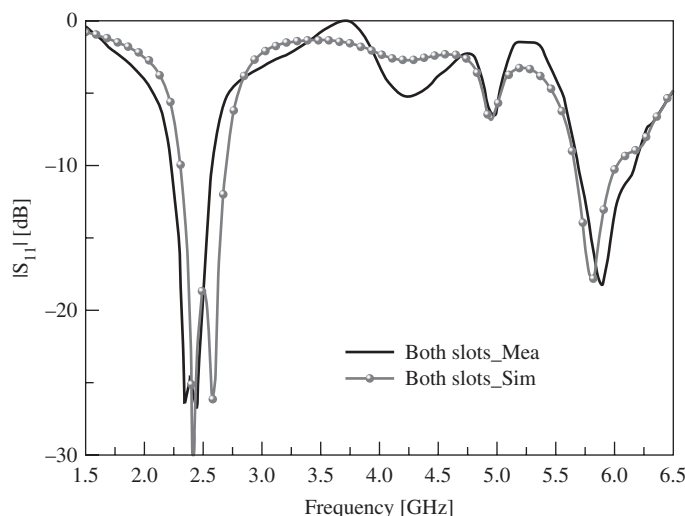


**Figure 4.37** Geometry of a dual-band pentagonal-ring slot antenna [50]. Reproduced with permission of © 2011 IEEE

at 2.45 GHz with the impedance bandwidth ( $|S_{11}| \leq -10$  dB) ranging from 2.3 to 2.6 GHz. The impedance bandwidth ( $|S_{11}| \leq -10$  dB) of the upper band is 5.1–6.1 GHz. Also, the CP bandwidth can cover the WLAN operation bandwidth (2.4–2.485 GHz, 5.15–5.35 GHz, and 5.725–5.825 GHz). The peak gain in the lower band is about 0.3 dBi with gain variation less than 0.15 dB. For the upper band, the peak gain reaches around 4.8 dBi with gain variation less than 2 dB.

Apart from annular-ring slots, slots with other shapes such as pentagonal- and square ring slots have also been utilized in [50,51] to achieve CP slot antennas. As depicted in Figure 4.37, the pentagonal-ring slot antenna is printed on an FR4 substrate ( $h = 1.6$  mm and  $\epsilon_r = 4.4$ ) with dotted line representing the 50- $\Omega$  microstrip line and white regions as the etched slots in the ground plane. The antenna is fed by a microstrip line through proximity coupling. The outer and ring slots are designed to operate around 2.4 and 5.8 GHz. The frequency ratio between the operating frequencies can be controlled by varying the dimension of two slots. The geometric dimensions of the antenna are as follows:  $L_1 = 17.4$  mm,  $L_2 = 16.7$  mm,  $a_1 = 10.9$  mm,  $a_2 = 2.4$  mm,  $b_1 = 0.1$  mm,  $b_2 = 0.1$  mm,  $g_1 = 0.1$  mm,  $g_2 = 0.1$  mm,  $L_f = 23.5$  mm and  $w = 3$  mm.

The simulated and measured results of reflection coefficient are shown in Figure 4.38. The measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) of the lower and upper bands are 14.3% and 8.1%, respectively. Also the obtained 3-dB AR bandwidths are 7.6% in the lower band and 4.1% in the upper band, as shown in Figure 4.39. Measured radiation patterns in two principal planes at two frequencies are plotted in Figure 4.40. As observed, the slot antenna presents bi-directional patterns with similar shape on both sides. The design arrangement radiates LHCP fields in the upper-half space and RHCP fields in the lower-half space.



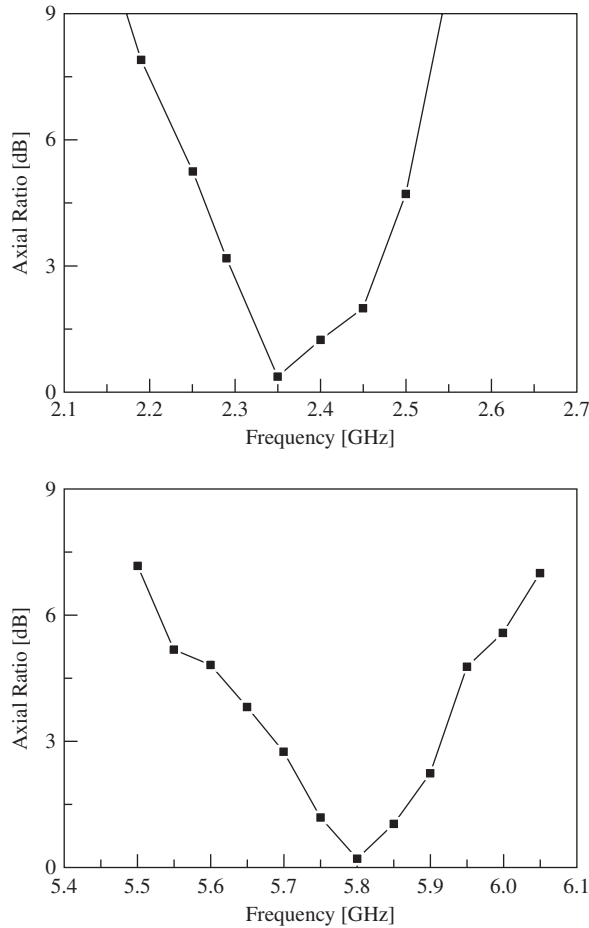
**Figure 4.38** Simulated and measured results of reflection coefficient [50]. Reproduced with permission of © 2011 IEEE

Opposite CP radiation can be obtained when the feedline is located at the right. Measured peak gains are 3.7 and 3.2 dBi within the lower and upper bands.

#### 4.4.4 Uni-directional CP Slot Antennas

In order to increase the antenna gain and reduce the backward radiation, Figure 4.41 presents the geometry of a cavity-backed CP annular slot antenna [51] operating at L1 and L2 bands. It consists of an F-shaped feedline on the top of a 1.6-mm thick FR4 substrate and a slotted ground plane with a side length of  $G$  on the other side of the substrate. The dual-band operations are obtained by inserting two concentric square annular slots in the ground. The lower operating frequency is determined by the outer slot whereas the higher resonant mode depends on the inner slot. A square cavity with a length of  $D$  and a height of  $H$  is attached to the ground to improve the front-to-back ratio. Circular polarization at two frequencies is achieved by connecting two stubs to the microstrip line with a width of  $w_f = 3$  mm. By adjusting the dimensions of the slots and the feedline, the antenna can support CP radiation in the dual bands. The optimized dimensions for achieving good CP performances in the dual bands are:  $D = 65.2$  mm,  $H = 20$  mm,  $G = 100$  mm,  $l_o = 55.2$  mm,  $l_i = 35$  mm,  $d_o = 4.6$  mm,  $d_i = 4$  mm,  $l_{s1} = 34.6$  mm,  $l_{s2} = 21.2$  mm,  $w_{s1} = 8$  mm,  $w_{s2} = 3.6$  mm and  $g = 8.4$  mm.

The current distributions of the antenna in the dual bands are illustrated in Figures 4.42 and 4.43, respectively. As observed, the current distribution at the L2 band is mainly concentrated on the middle annular ring surrounded by the two annular slots. At  $t = 0$ , the current flows at the upper left and lower right corners and their vector sum points from the upper right corner to the lower left corner. At  $t = T/4$ , the current flows at the upper right and the lower left corners and the vector sum points from the upper left corner to the lower right corner. The RHCP radiation is expected by following the movement of these vector sums with time.

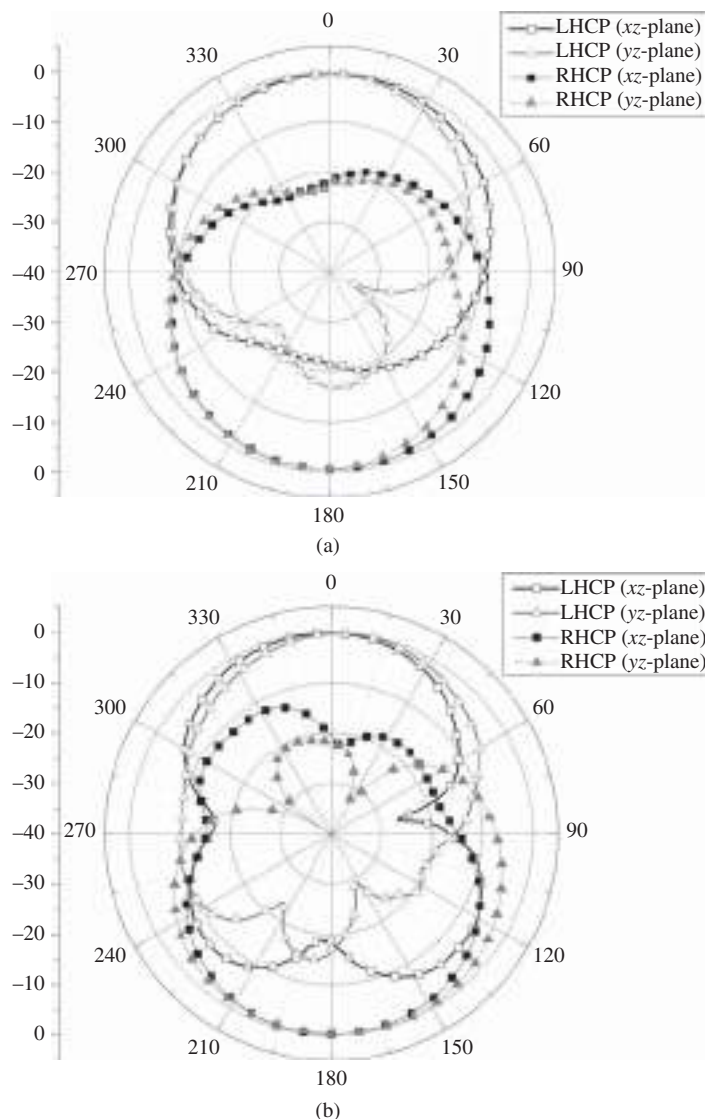


**Figure 4.39** Measured AR in the lower and upper bands [50]. Reproduced with permission of © 2011 IEEE

At the L1 band, the current distributions are mostly concentrated on the outer edges of the square patch and inner edges of the square ring. In other words, the current flows along the inner slot. The vector sum of current rotates counter-clockwise with time. Hence, the RHCP can be achieved at the L1 band.

The simulated and measured results of reflection coefficient and AR are shown in Figure 4.44. The measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) at L2 and L1 bands are 3.7% (1.19–1.235 GHz) and 1.2% (1.565–1.585 GHz), respectively. The obtained 3-dB AR bandwidths at L2 and L1 bands are 0.9% (1.220–1.231 GHz) and 0.6% (1.572–1.581 GHz), respectively. The measured AR values at 1.227 and 1.575 GHz are 1.39 and 2 dB, respectively.

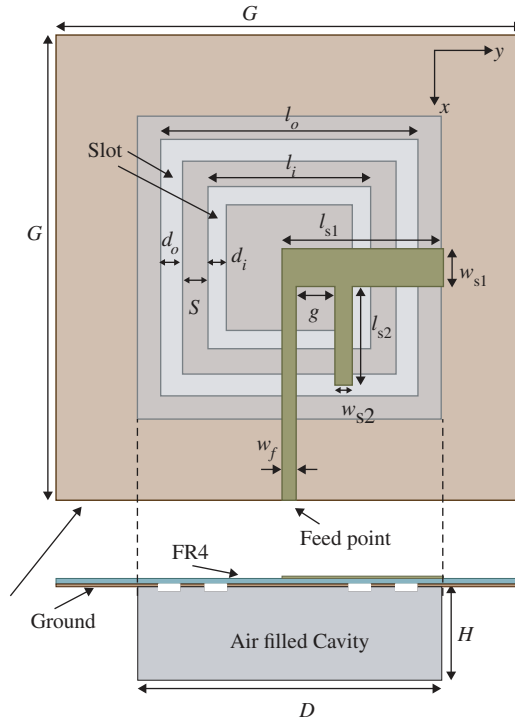
The simulated and measured radiation patterns at 1.227 and 1.575 GHz are illustrated in Figure 4.45. As observed, directional radiation patterns at both frequencies have been obtained. The radiation efficiencies are about 86.5% and 80.7% at L1 and L2 bands,



**Figure 4.40** Measured radiation patterns at (a) 2.4 and (b) 5.8 GHz [50]. Reproduced with permission of © 2011 IEEE

respectively. The 3-dB beamwidths are found to be around  $100^\circ$  in both bands. The measured antenna gains are over 1.45 dBic at the L1 band and 1.1 dBic at the L2 band.

Another dual-band CP dual-slot antenna [52] which can achieve directional radiation is covered by a protective dielectric, as shown in Figure 4.46. The lower operating frequency is realized by inserting a zonal slot of width  $W_z = 1$  mm onto a copper square cavity with an inner side length of  $A = 22$  mm and a thickness of 0.3 mm. The whole cavity is divided into the upper and lower parts due to the zonal slot. The impedance matching can be improved



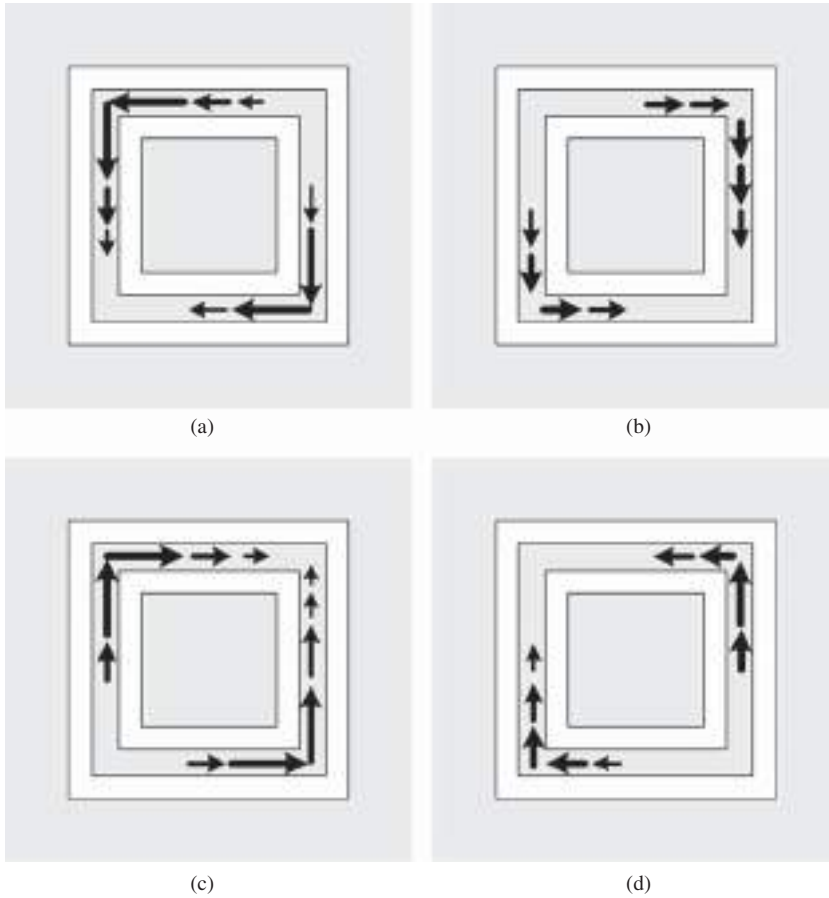
**Figure 4.41** Geometry of the dual-band CP cavity-backed annular-slot antenna [51]. Reproduced with permission of © 2012 IEEE

by the lower part. The upper and lower parts have the inner heights of  $h_1 = 15.5$  mm and  $h_2 = 5.5$  mm. A square ground plane is attached to the edge of the lower part, whereas two pieces of foam are inserted in the zonal slot to support the upper part. The ground plane has a side length of 150 mm and a thickness of 0.5 mm. An L-shaped probe with a radius of 0.5 mm is located diagonally inside the cavity with an offset of  $x_0 = 5.52$  mm to excite the zonal slot. The lengths of its horizontal and vertical arms are  $L_h = 14.5$  mm and  $L_v = 20$  mm, respectively. In order to generate CP fields, the zonal slot is perturbed by two symmetrical cuts placed at an offset of  $s = 3$  mm. It is a step aperture with the areas of the lower and upper parts given by  $2 \times 3$  mm<sup>2</sup> and  $1 \times 2$  mm<sup>2</sup>.

The upper operating frequency band is achieved by cutting an annular slot on the top face of the upper cavity. It has a radius of  $R = 5.45$  mm and a width of  $W_a = 0.7$  mm. Circular polarization is obtained by utilizing three asymmetrical meander-slot sections at  $\varphi = 0^\circ$ ,  $22.5^\circ$  and  $225^\circ$ , with the sections at  $0^\circ$  and  $22.5^\circ$  being identical. The parameters are given by  $L_a = 1.7$  mm,  $L_b = 2.2$  mm,  $L_c = 2.3$  mm,  $L_d = 2.8$  mm and  $W_b = 0.3$  mm. The widths of all the meander slots are 0.2 mm. The covering dielectric has a dielectric constant of  $\epsilon_r = 2.5$ , a loss tangent of  $\tan \delta = 0.0004$  and transverse dimensions of  $a = 36$  mm,  $d = 21$  mm.

The formulas for predicting the operating frequencies are given by [52]:

$$f_l \approx \frac{c}{4A\sqrt{\epsilon_e}} \quad (4.4)$$



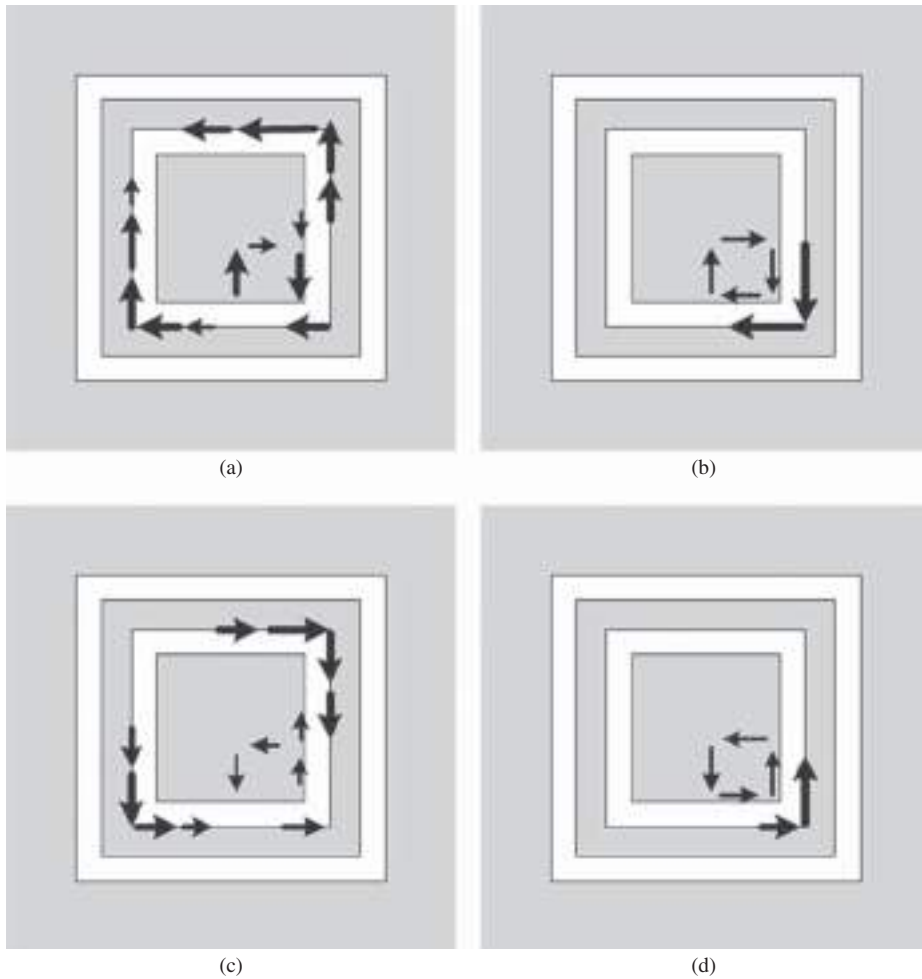
**Figure 4.42** Current distributions at L2 band with period  $T$ , (a)  $t = 0$ , (b)  $t = T/4$ , (c)  $t = T/2$  and (d)  $t = 3T/4$  [51]. Reproduced with permission of © 2012 IEEE

$$\varepsilon_e = \frac{2\varepsilon_r}{\varepsilon_r + 1} \quad (4.5)$$

$$f_u \approx \frac{c}{(2\pi R + 4L_a + 2L_c)\sqrt{\varepsilon_e}} \quad (4.6)$$

where  $\varepsilon_e$  is the effective dielectric constant,  $f_l$  and  $f_u$  are the lower and upper operating frequencies.

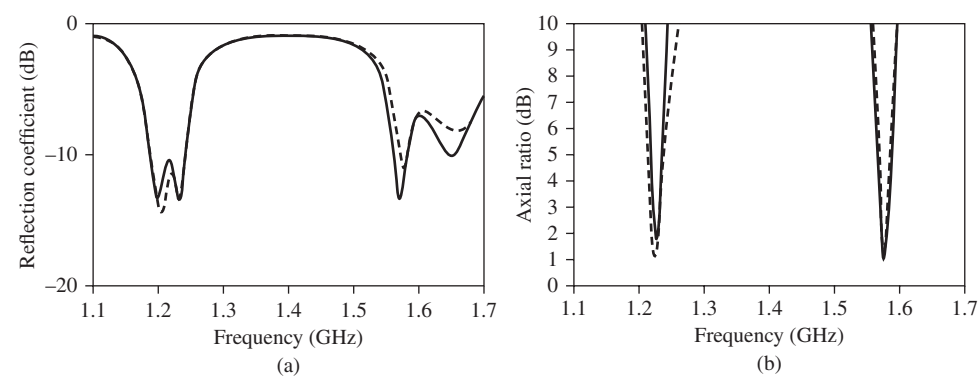
The simulated, measured and estimated resonant frequencies of the zonal- and annular-slot modes are summarized in Table 4.3. In the lower band, the measured resonant frequency with minimum  $|S_{11}|$  is 2.43 GHz which agrees well with the simulated value of 2.41 GHz (0.82% error). The estimated resonant frequency is 2.85 GHz (15.44% error) as the perturbation of the zonal slot has not been taken into consideration. In the upper band, the measured and simulated resonant frequencies are 5.79 and 5.84 GHz (0.86% error), respectively. The



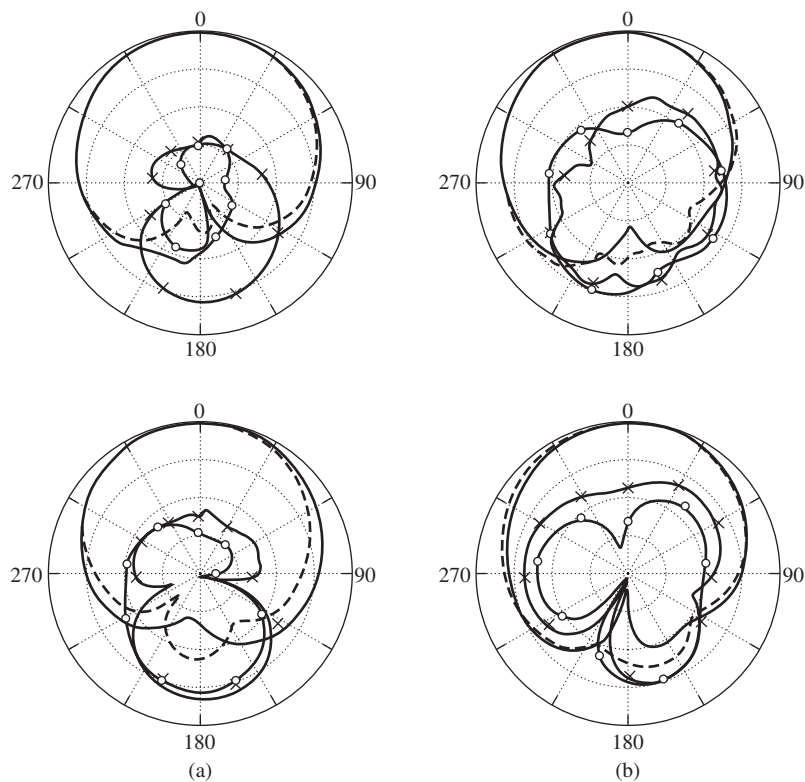
**Figure 4.43** Current distributions at L1 band with period  $T$ , (a)  $t = 0$ , (b)  $t = T/4$ , (c)  $t = T/2$  and (d)  $t = 3T/4$  [51]. Reproduced with permission of © 2012 IEEE

estimated resonant frequency is 5.5 GHz. A reasonable agreement between the estimation and simulation has been obtained.

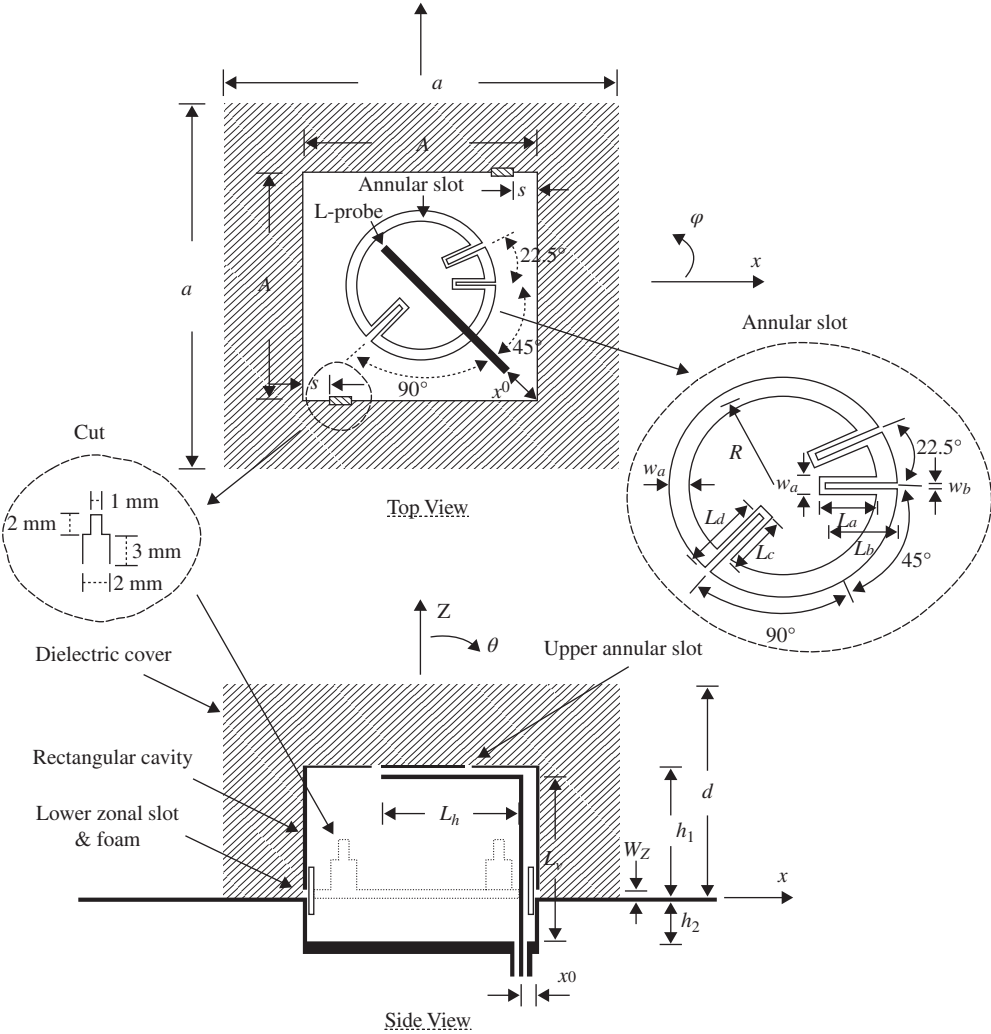
The measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) of the lower and upper bands are 15.83% (2.21–2.59 GHz) and 3.45% (5.7–5.9 GHz). The obtained 3-dB AR bandwidths are 2.86% (2.41–2.48 GHz) and 2.64% (5.6–5.75 GHz), respectively. For the patterns of zonal slot, the LHCP fields are less than  $-23$  dB compared to the RHCP fields in the broad-side direction. The front-to-back ratios of the x-z and y-z planes are around 22 and 18 dB, respectively. For the annular slot, the RHCP fields are stronger than the LHCP fields by more than 21 dB at boresight with the front-to-back ratios being about 17 dB for both planes. The maximum gains are 5.59 dBi at 2.47 GHz, and 3 dBi at 5.78 GHz, respectively.



**Figure 4.44** Simulated (— —) and measured (—) results of reflection coefficient and AR [51]. Reproduced with permission of © 2012 IEEE



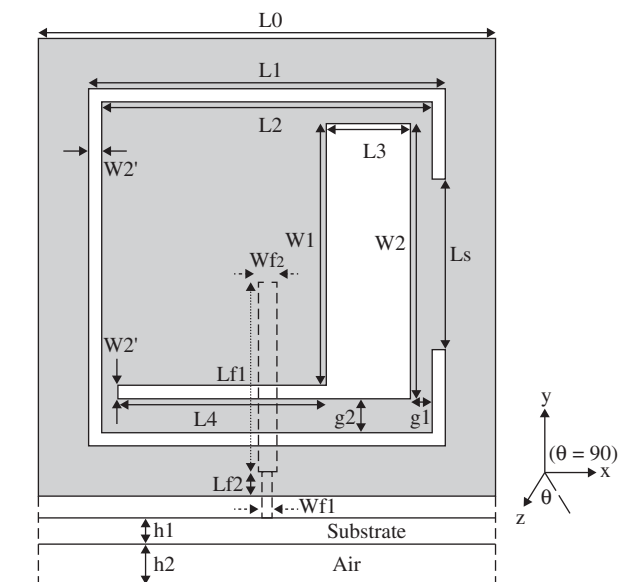
**Figure 4.45** Simulated and measured radiation patterns at (a) 1.227 and (b) 1.575 GHz. [51] (— : measured RHCP; — —: simulated RHCP; — o —: measured LHCP; — x —: simulated LHCP). Reproduced with permission of © 2012 IEEE



**Figure 4.46** Configuration of the CP dual-slot antenna. Two pieces of foam are inserted in the zonal slot to support the upper cavity [52]. Reproduced with permission of © 2009 IEEE

**Table 4.3** Comparisons among simulated, measured and estimated resonant frequencies [52]. Reproduced with permission of © 2009 IEEE

| Resonant mode | Simulation             | Measurement            |       | Estimation             |
|---------------|------------------------|------------------------|-------|------------------------|
|               | $f_{\text{sim}}$ (GHz) | $f_{\text{mea}}$ (GHz) | error | $f_{\text{est}}$ (GHz) |
| Zonal slot    | 2.41                   | 2.43                   | 0.82% | 2.85                   |
| Annular slot  | 5.84                   | 5.79                   | 0.86% | 5.5                    |



**Figure 4.47** Geometry of a dual-band CP antenna with C- and L-shaped slots [53]. Reproduced with permission of © 2011 John Wiley & Sons, Inc.

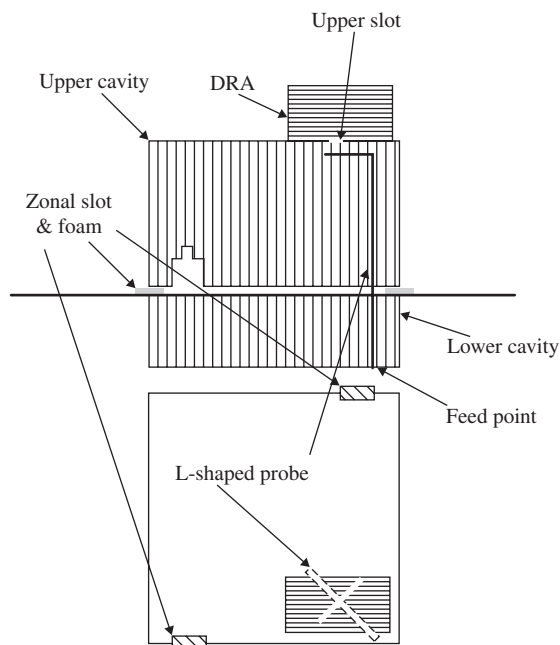
#### 4.4.5 Other Multi-Band CP Slot Antennas

In Figure 4.47, the dual-band slot antenna is realized by embedding an L-shaped slot and a C-shaped slot in the ground plane. The L-slot is inside the C-slot which has the same width and is different from a square ring slot in having a shortened section. The L-slot is formed by its horizontal and vertical portions with different widths. In order to improve the impedance matching, the microstrip line consists of two sections with different widths. The antenna in [53] is designed for operating in the 1.57 and 2.45 GHz bands with CP performance. The CP bandwidths of both bands are around 6%.

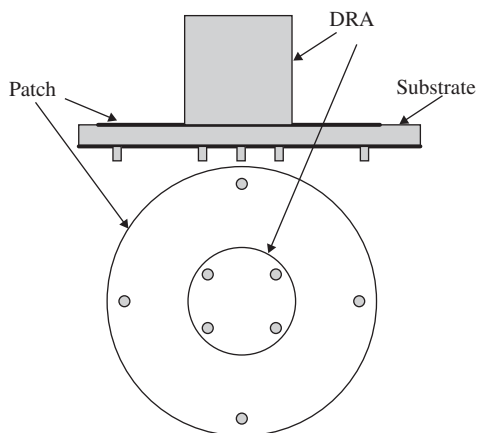
### 4.5 Multi-Band CP DRAs

#### 4.5.1 DRA/Slot or DRA/Patch Hybrid Antennas

In [54,55], the zonal slot antenna cut onto a cylindrical cavity is combined with a DRA to achieve a dual-band CP hybrid antenna. As shown in Figure 4.48, a zonal slot is cut onto a rectangular copper cavity, thus the cavity is divided into its upper and lower parts. Two pieces of foam are inserted in the slot to support the upper cavity. The zonal slot is perturbed by two symmetrical cuts to excite the CP fields for the lower band. A ground plane is connected to the lower cavity. The L-shaped probe consisting of a horizontal- and vertical arm is diagonally located in the cavity. The DRA is fabricated by a substrate with a relative permittivity of 10. A rectangular slot is embedded on the top of the cavity and inclined at  $45^\circ$  to generate CP fields for the upper band. The impedance bandwidths ( $|S_{11}| \leq -10$  dB) of both lower and upper bands are 12.8% (2.34–2.66 GHz) and 12.52% (5.69–6.45 GHz),



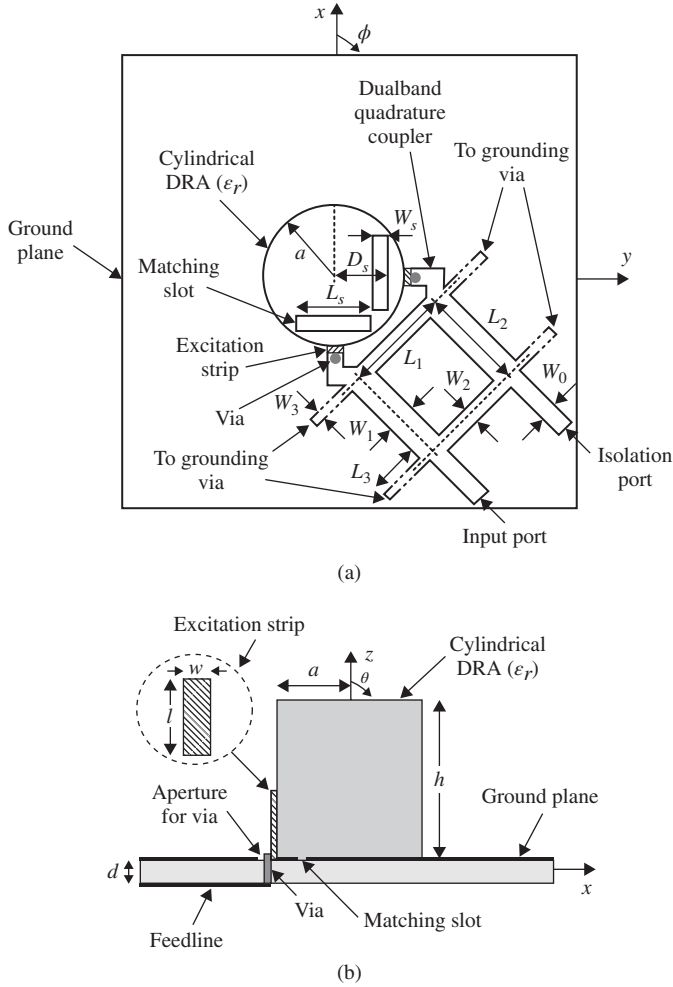
**Figure 4.48** Geometry of a dual-band CP DRA/slot hybrid antenna [55]. Reproduced with permission of © 2007 IEEE



**Figure 4.49** Geometry of a dual-band CP DRA/patch hybrid antenna [56]. Reproduced with permission of © 2003 IEEE

respectively. The obtained 3-dB AR bandwidths are 3.51% for the lower band and 4.87% for the upper band, respectively.

In addition, a dual-band CP hybrid antenna is realized by utilizing a cylindrical dielectric resonator and a microstrip patch, as studied in [56]. The antenna configuration is shown in Figure 4.49. A circular patch is printed on a 3.175-mm thick dielectric substrate with a



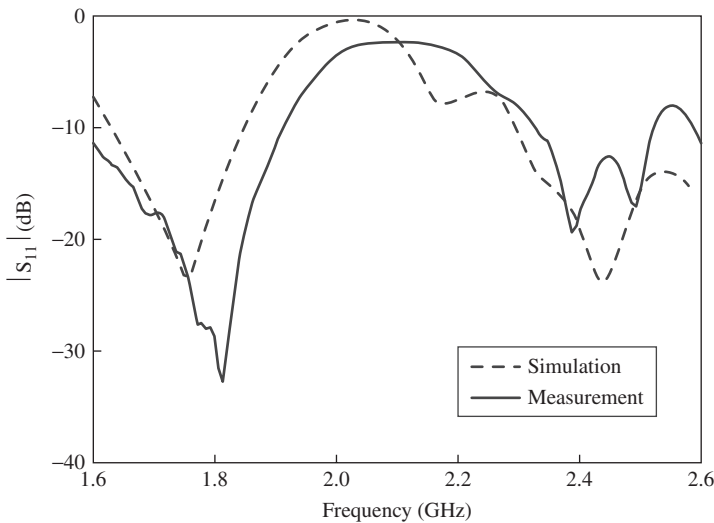
**Figure 4.50** Geometry of a dual-band CP dual-mode DRA [59]. Reproduced with permission of © 2012 IEEE

dielectric constant of 2.2. The circular patch is excited by four 50- $\Omega$  coaxes with sequential phase difference of  $90^\circ$  to achieve good circular polarization. The circular patch antenna is designed for operating in the lower band. The dual-band operation is obtained by stacking a cylindrical dielectric resonator on the top of the circular patch. The influence on the performance of the DRA by the patch is relatively small, as the diameter of the patch is relatively large compared to that of the DRA. Moreover, the circular patch can provide a ground plane for the DRA. The DRA is excited by four coaxes in phase quadrature and the feed position of the DRA is rotated by  $45^\circ$  with respect to that of the patch antenna to reduce the coupling. The simulated result demonstrates the hybrid antenna can resonate at 2.1 and 4.9 GHz.

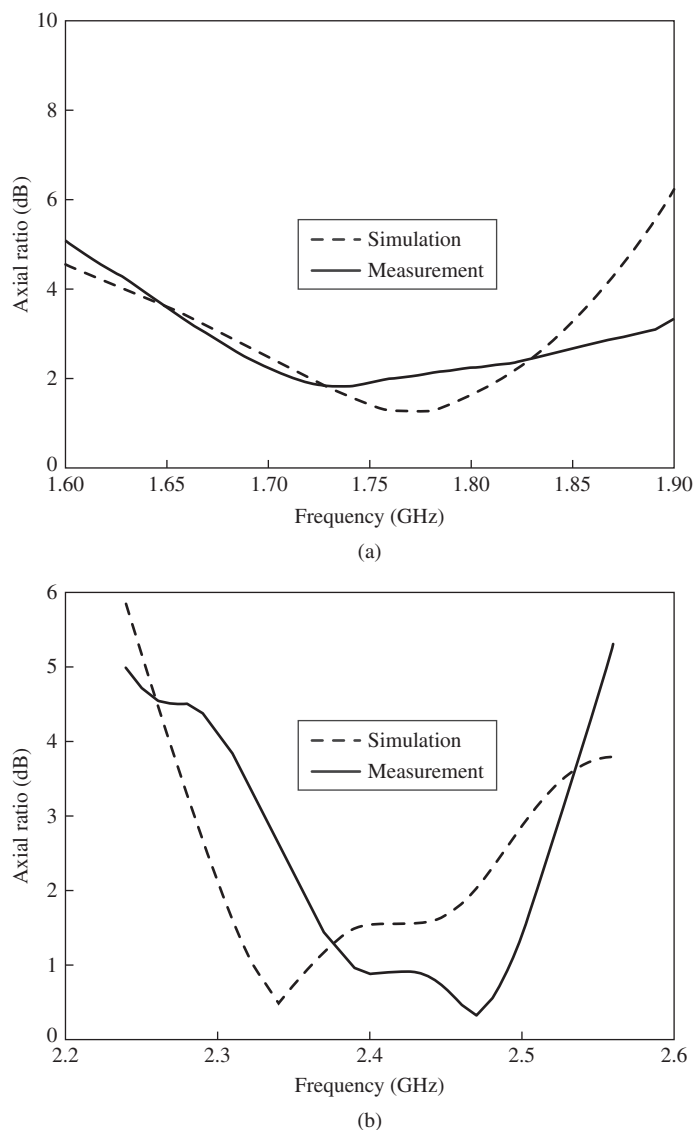
### 4.5.2 Excitation of Multiple Modes

It is found in [57,58] that, a higher-order DRA mode can be utilized to perform dual-band operation. It does not require a second resonating element to achieve dual-band DRA when using a higher-order mode. Figure 4.50 presents the geometry of a dual-band dual-mode cylindrical DRA. The cylindrical dielectric has a dielectric constant of  $\epsilon_r = 9.4$ , a radius of  $a = 18.7$  mm, and a height of  $h = 42.5$  mm. The dual-band operation is obtained by combining the  $\text{HEM}_{111}$  and  $\text{HEM}_{113}$  modes of the DRA. In order to achieve CP radiation, the quadrature strip-fed method is utilized to excite the antenna. Two feeding strips located angularly at  $90^\circ$  from each other are attached on the side wall of the DRA. Each strip has a length of  $l = 12.5$  mm and a width of  $w = 1$  mm. Two slots are inserted on the ground to improve the impedance matching. Each slot is located at a distance of  $D_s = 12.75$  mm, and has a length of  $L_s = 21$  mm and a width of  $W_s = 1.5$  mm. To provide two quadrature signals, a dual-band hybrid coupler is fabricated on the bottom side of a substrate which has a dielectric constant of  $\epsilon_{rs} = 2.33$ , a thickness of  $d = 1.57$  mm and a size of  $140 \times 140$  mm<sup>2</sup> whereas the ground is printed on the top side. The feeding strips are connected to the output lines of the hybrid coupler through vias. The geometrical parameters of the coupler are:  $L_1 = 26.9$  mm,  $L_2 = 26.5$  mm,  $L_3 = 56.65$  mm,  $W_0 = 4.66$  mm,  $W_1 = 7.3$  mm,  $W_2 = 4.44$  mm and  $W_3 = 0.46$  mm.

Figure 4.51 illustrates the simulated and measured reflection coefficient for the dual-band DRA. As observed, the measured impedance bandwidths ( $|S_{11}| \leq -10$  dB) of the lower and upper bands are 18.9% (1.58–1.91 GHz) and 7.8% (2.33–2.52 GHz), respectively. The simulated and measured AR values are depicted in Figure 4.52. It is found that, the measured 3-dB AR bandwidths of the lower and upper bands are 12.4% (1.67–1.89 GHz) and 7.4% (2.34–2.52 GHz), respectively. The peak antenna gains of the lower and upper bands are found to be 6.23 and 8.01 dBic, respectively.



**Figure 4.51** Simulated and measured reflection coefficient for the dual-band dual-mode DRA [59]. Reproduced with permission of © 2012 IEEE

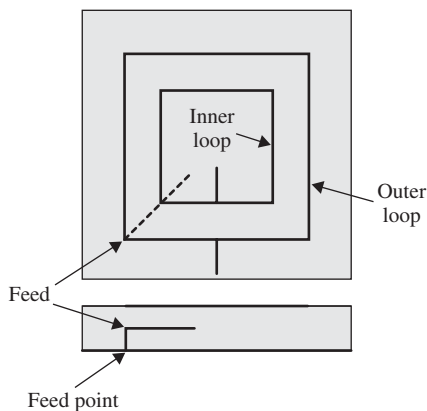


**Figure 4.52** Simulated and measured AR for the dual-band dual-mode DRA [59]. Reproduced with permission of © 2012 IEEE

## 4.6 Multi-Band CP Loop Antennas

### 4.6.1 Dual-Loop Antennas

Figure 4.53 illustrates the geometry of a dual-band CP dual-loop antenna. Two square loops are printed on a grounded dielectric substrate to achieve dual-band operations. Both loops have a perturbation segment for the CP radiation. One of the loops can radiate CP wave when

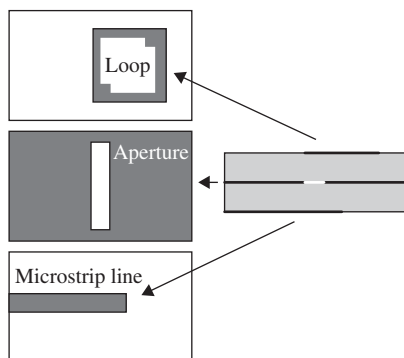


**Figure 4.53** Geometry of a dual-band dual-loop antenna [60]. Reproduced with permission of © 2009 IEEE

its circumference is approximately one wavelength. A proximity coupling feed technique with an L-shaped wire is utilized for the impedance matching with a  $50\text{-}\Omega$  coaxial line. The L-shaped wire is formed by its vertical and horizontal parts. It is noted that, the lower frequency band is determined by the outer loop while the inner loop is designed for operating over the upper frequency range. It has also been studied in [60] that compared to the discrete dual-loop antenna, a wide-band frequency response can be obtained when the corners of two loops are connected by wires.

#### 4.6.2 Excitation of Dual Modes

The investigation in [61] has shown that, when the mean circumference of a typical square loop is equal to an integral multiple of the guided wavelength, it can only create one resonant mode. However, the dual-mode square loop resonator can be realized when perturbations are employed at  $45^\circ$  or  $135^\circ$  in the loop, as shown in Figure 4.54. The loop antenna consists of a



**Figure 4.54** Geometry of a tri-band aperture-coupled loop antenna [61]. Reproduced with permission of © 2010 IEEE

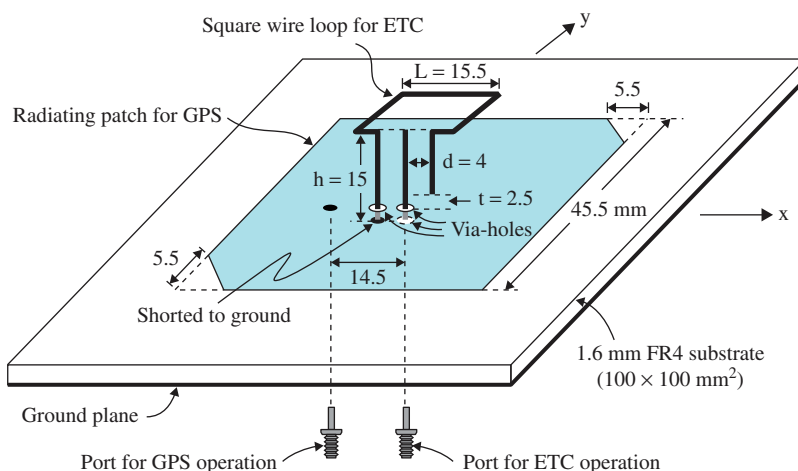
dual-mode square loop resonator on the upper layer, a slotted ground plane in the middle and a microstrip line on the lower layer. A wideband performance can be obtained by tuning the feedline. The measured 6-dB return loss response can cover three bands, 1.68–1.89 GHz, 3.66–3.86 GHz and 4.68–6.03 GHz. The minimum AR values for 1.87 GHz band, 3.75 GHz band and 5.32 GHz band are 0.25, 1.03 and 0.95 dB, respectively. The obtained 3-dB AR bandwidths in three frequency bands are 109, 90 and 150 MHz, respectively.

## 4.7 Other Multi-Band CP Antennas

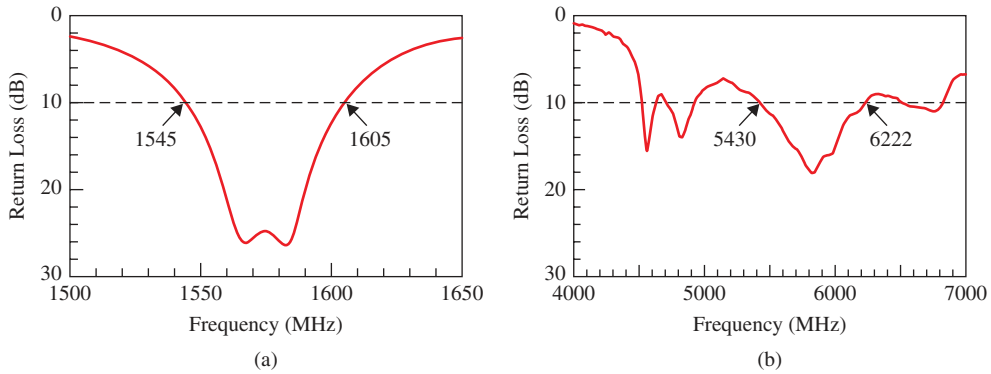
### 4.7.1 Hybrid Antennas

Several dual-band hybrid CP antennas have been studied in [62,63]. In [62], a spiral antenna-loaded a helix is investigated to provide circular polarization in dual bands. The total size of the antenna is unchanged when loading an addition helix. The lower band is determined by the helix whereas the upper frequency band can be tuned by adjusting the dimensions of the spiral. The obtained CP bandwidths of the lower and upper bands are 4% and 62.07%, respectively.

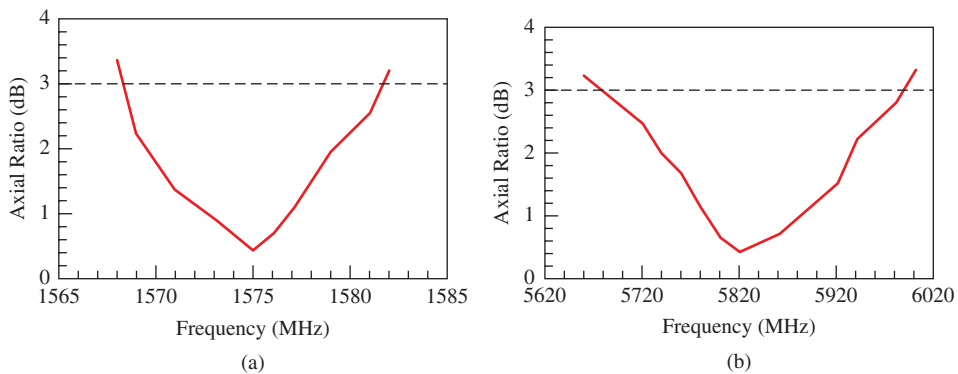
The CP antenna presented in Figure 4.55 can achieve a dual-band operation with two separate feeds. It consists of a patch antenna operating at 1.575 GHz and a square loop antenna designed for 5.8 GHz application. The circular polarization of the patch antenna is obtained by using a corner-truncated square patch. The square loop antenna is excited by a monopole feed and formed by three major portions: a horizontal square wire loop, a shorted vertical wire and an open vertical wire. The centre operating frequency of the loop antenna can be controlled by adjusting the length of the horizontal loop. The shorted vertical wire has the same length as the monopole feed and is short-circuited to the ground plane through a via hole in the square patch. The AR of the loop antenna depends on the length of the height of the loop or the length of the shorted vertical wire. It is noted that the monopole



**Figure 4.55** Geometry of a dual-band patch/loop hybrid antenna [63]. Reproduced with permission of © 2004 John Wiley & Sons, Inc.



**Figure 4.56** Measured return loss for the two bands [63]. Reproduced with permission of © 2004 John Wiley & Sons, Inc.



**Figure 4.57** Measured AR for the two bands [63]. Reproduced with permission of © 2004 John Wiley & Sons, Inc.

feed, shorted vertical wire and the open vertical wire are all in the same plane and separated with an equal distance. The excitation of the loop antenna is realized through capacitive coupling of the monopole feed to the open vertical wire.

The measured results of return loss and AR for the lower and upper bands are shown in Figures 4.56 and 4.57. As can be observed, the obtained 10-dB return loss bandwidths of the lower and upper bands are 60 MHz (1.545–1.605 GHz) and 792 MHz (5.43–6.222 GHz), respectively. The obtained 3-dB AR bandwidths are 14 MHz (0.9% at 1.575 GHz) in the lower band and 310 MHz (5.3% at 5.8 GHz) in the upper band. The measured antenna gains are 1.6 dBic at 1.575 GHz and 5.7 dBic at 5.8 GHz, respectively.

#### 4.7.2 Rectennas

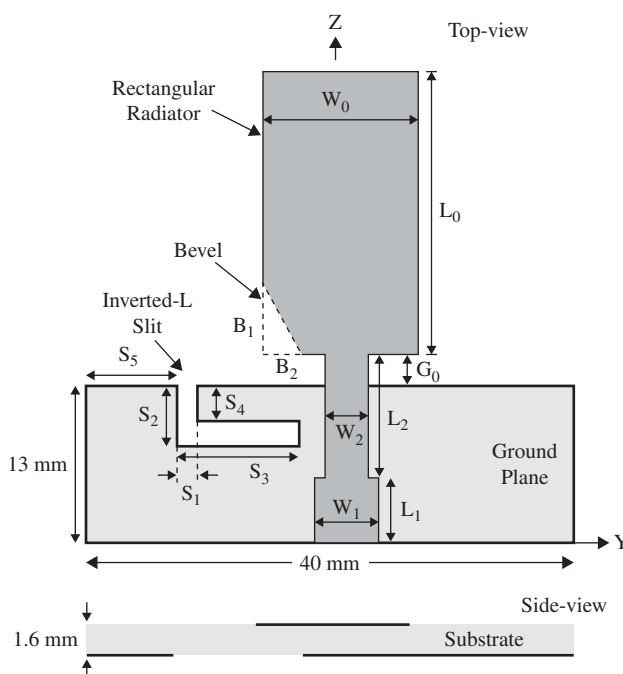
A rectenna is a rectifying antenna and utilized to convert microwave energy into direct current electricity. They are suitable for applications in wireless power transmission systems that transmit power by radio waves. Circular polarization enables the transmitting

or receiving antennas to be rotated without significant change in the output DC voltage. A dual-frequency CP rectenna for wireless power transmission at 2.45 and 5.8 GHz is investigated in [64]. The dual-band antenna consists of two nested microwave-fed shorted annular ring-slot antennas and two rectifier circuits. The design in [64] can achieve 1–2 V DC output voltage at a transmission distance of 2 m for relatively low power densities, that is, RFID applications.

#### 4.7.3 Dual-Band Dual-Sense CP Antennas

The dual-band dual-sense CP antennas can offer improved isolation between channels compared to the co-polarized systems. In addition, they are suitable for the applications where dual-band reception of both RHCP and LHCP signals are required. Recently, the research on the dual-band dual-sense CP antennas has attracted increasing attention. Several dual-band dual-sense CP antennas have been investigated in [65–73]. They include monopole [65], slot monopole [66,67], slotted patch [68,69], slot antenna [70,71], DRA [72] and loop antenna [73].

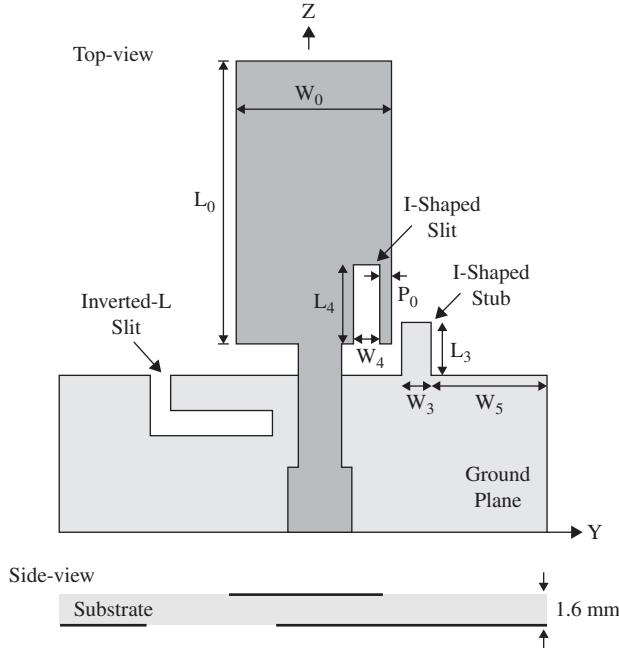
As shown in Figure 4.58, the microstrip-fed monopole antenna consists of a rectangular radiator and a ground plane. It is printed on an FR4 substrate with a thickness of 1.6 mm and a relative permittivity of 4.4. In general, the length of the monopole antenna is usually around a quarter-wavelength. An inverted L-shaped slit is embedded in the ground plane to excite two



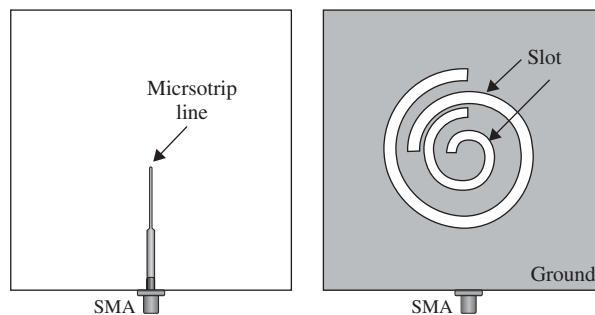
**Figure 4.58** Geometry of a dual-band dual-sense CP monopole antenna with an inverted-L slit [65]. Reproduced with permission of © 2009 IEEE

orthogonal fields with equal amplitudes and a  $90^\circ$  phase shift for radiating LHCP at 2.5 GHz and RHCP at 3.4 GHz. A bevel is cut in the rectangular radiator to enhance the impedance bandwidth. This design can achieve a broadband impedance bandwidth of 102.5% at the centre operating frequency of 4.35 GHz and dual-band CP operations of 6.0% LHCP at 2.485 GHz (2.41–2.56 GHz) and 6.7% RHCP at 3.425 GHz (3.31–3.54 GHz). To further increase the impedance and AR bandwidths, a modified monopole antenna with improved performance is also investigated in [65] and shown in Figure 4.59. An I-shaped slit and stub are added in the rectangular radiator and ground plane, respectively. The I-shaped slit in the rectangular radiator can excite RHCP fields in the upper frequency band. The AR bandwidth of the upper band can be increased by tuning the dimension of  $L_3$  and  $L_4$ . The impedance bandwidth can be increased to be 118.4%. The obtained 3-dB AR bandwidths are 5.6% LHCP in the lower band (2.41–2.55 GHz) and 23.1% RHCP in the upper band (3.45–4.35 GHz), respectively.

In addition, slot monopole antennas with dual-band dual-sense CP performance have been proposed in [66,67]. In [66], the CPW-fed slot monopole antenna has two spiral slots embedded in the ground plane which can result in different senses of circular polarization. The dual-band performance is obtained by applying the T-shaped strip. The measured 10-dB return loss bandwidths of the lower and upper bands are 8.7% and 23%, respectively. The obtained 3-dB AR bandwidths are 8.4% and 19.24% with respect to 1.6 GHz (RHCP) and 2.2 GHz (LHCP). The frequency ratio of two frequencies is 1.375. In [67], the dual-band dual-sense CP slot antenna is obtained by embedding the slots in two opposite corners for the lower band and adding a C-shaped grounded strip for the upper band.



**Figure 4.59** Geometry of a dual-band dual-sense CP monopole antenna with an inverted-L slit, an I-shaped slit and an I-shaped strip [65]. Reproduced with permission of © 2009 IEEE

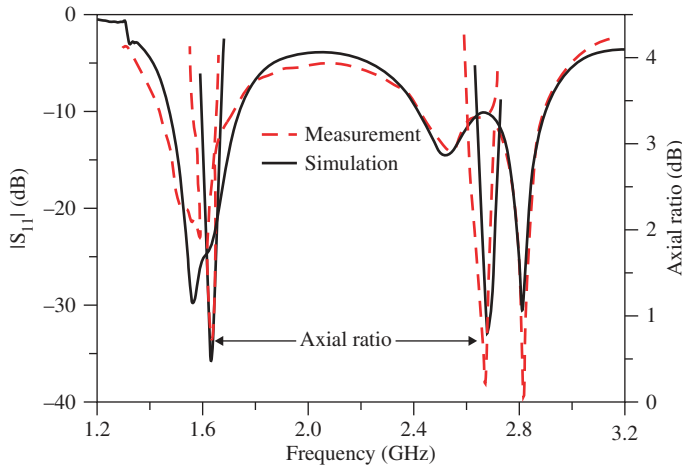


**Figure 4.60** Geometry of a dual-slot spiral slot antenna [70]. Reproduced with permission of © 2011 IEEE

A probe-fed slotted patch antenna is proposed and studied in [68]. It is formed by two corner-truncated patches which are linked together by four small strips. The dual-sense circular polarizations are obtained due to the corner-truncations on the patch. The antenna is designed for operating at 2.375 and 3.945 GHz. The obtained impedance bandwidths ( $|S_{11}| \leq -10$  dB) of the lower and upper bands are around 37 and 25 MHz, respectively. It is also found that, the 3-dB AR bandwidths are 10 MHz in the lower band (LHCP) and 23 MHz in the upper band (RHCP). In [69], the microstrip antenna has an annular-ring patch and is fed by an L-shaped microstrip line through the coupling of a ring slot in the ground plane. The annular-ring patch is printed on a 0.6-mm thick FR4 substrate. The L-shaped microstrip line is etched on the bottom side of a 1.6-mm thick FR4 substrate and the ground plane with an embedded ring-slot is on the top side. There is a foam material inserted between two FR4 substrates. The magnetic currents circulating along the ring slot at two frequencies have opposite flowing directions, thus causing the dual-sense circular polarization. The antenna in [69] has a 3-dB AR bandwidth of 2.2% in the 2.075 GHz band (LHCP) and 1.3% with respect to 2.735 GHz (RHCP). The measured peak gains are around 6.5 dBi for LHCP and 6 dBi for RHCP, respectively.

In [70], a dual-band dual-sense CP microstrip-fed antenna is realized by embedding two monofilar spiral slots in the ground plane, as shown in Figure 4.60. The ground plane is printed on one side of a substrate with a thickness of 1.57 mm and a relative permittivity of 3.5. It has a size of  $100 \times 100 \text{ mm}^2$  which is approximately a half of a free space wavelength at the lowest operating frequency. The microstrip line consisting of two stepped portions with different widths is printed on the other side for achieving good impedance matching. Two spiral slots are embedded in the ground plane to achieve dual-band dual-sense circular polarizations. The perimeter length of the inner slot is approximately  $\lambda_g$  ( $\lambda_g$  is the guided wavelength at the lowest frequency) for the lower frequency whereas it is around  $2\lambda_g$  for the higher frequency.

Dual-sense circular polarizations are obtained by realizing oppositely-directed current rotation for the two frequency bands. The investigations in [70] have shown that the current at the lower frequency travels in an anti-clockwise fashion when viewed from the +Z direction, thus leading to RHCP radiation. The upper frequency current travels in the opposite direction and yields the LHCP radiation in the +Z direction. In order to enhance the CP bandwidth in the lower band, a second spiral slot located inside the first

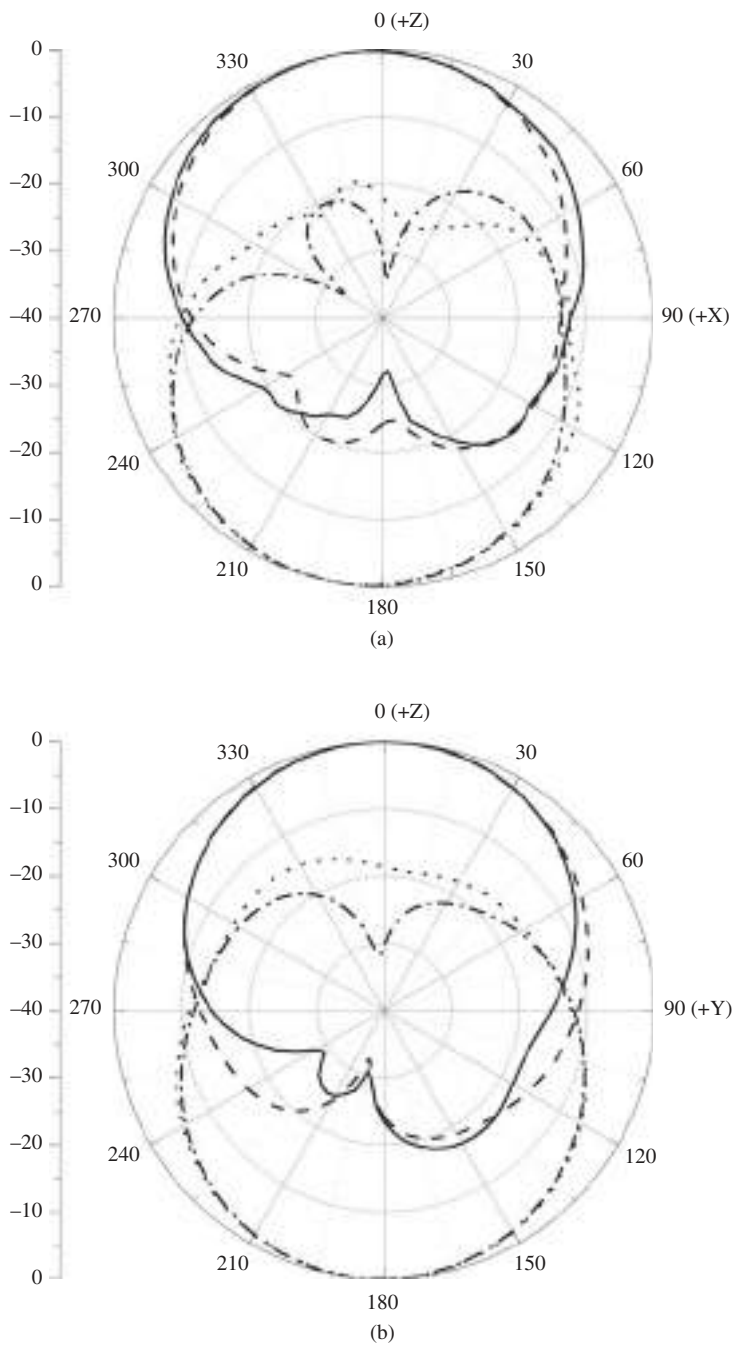


**Figure 4.61** Simulated and measured results of reflection coefficient and AR [70]. Reproduced with permission of © 2011 IEEE

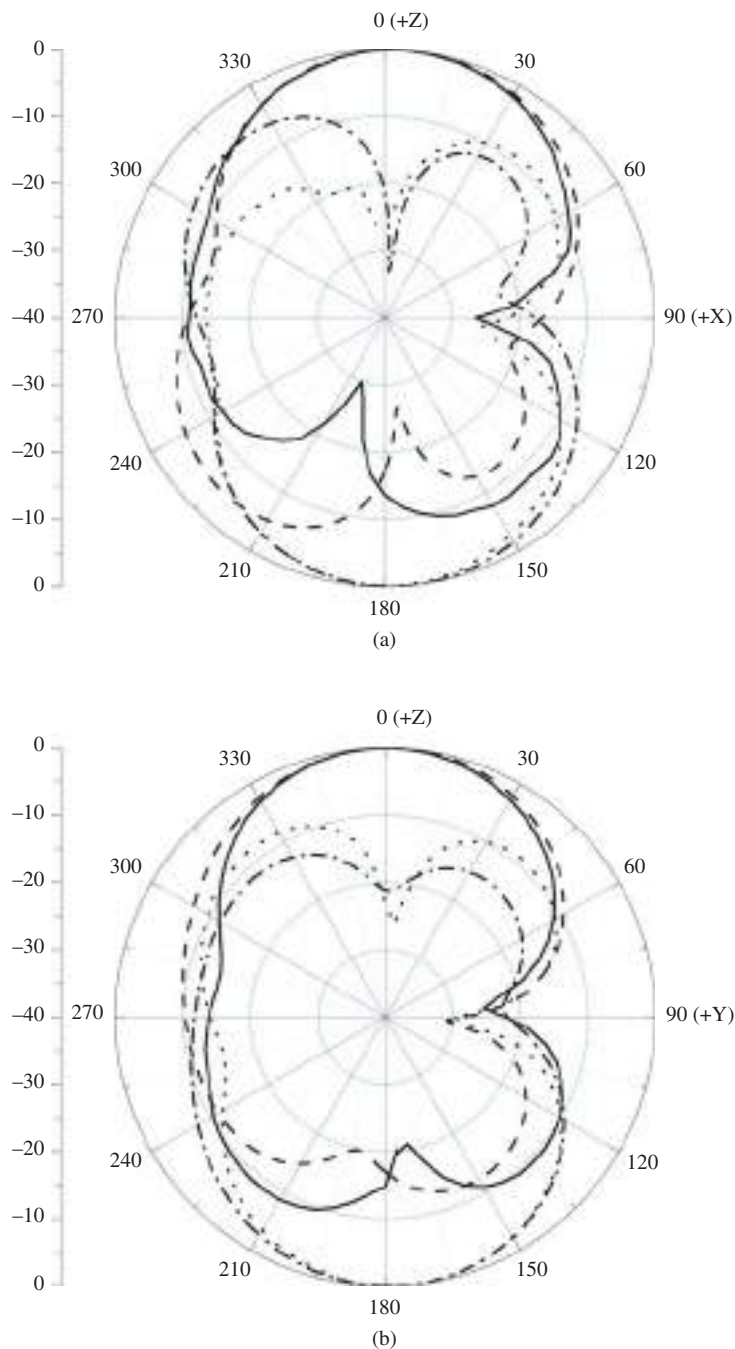
one is embedded in the ground plane. The smaller spiral slot is strongly coupled to the larger one for the lower frequency, thus significantly improving the impedance matching and AR. The perimeter length of the outer spiral slot is now approximately  $1.17 \lambda_g$  at the lower frequency, whereas it remains at approximately  $2 \lambda_g$  for the upper frequency. The simulated and measured results of reflection coefficient and AR are shown in Figure 4.61. The measured 10-dB return loss bandwidths of the lower and upper bands are 287 MHz (1.437–1.724 GHz) or 18.2% with respect to the centre frequency of 1.58 GHz, and 489 MHz (2.418–2.907 GHz) or 18.4% corresponding to the frequency of 2.663 GHz. The obtained 3-dB AR bandwidths of the lower (RHCP) and upper (LHCP) bands are 4.45% (1.58–1.652 GHz) and 3.5% (2.609–2.702 GHz), respectively.

The simulated and measured radiation patterns in two principal planes at two frequencies are shown in Figures 4.62 and 4.63. As observed, the cross-polarization levels at boresight are less than 20 dB compared to the co-polarization. At 1.64 GHz, the RHCP radiation pattern in the  $yo\bar{z}$  plane is symmetrical and has a 3-dB beamwidth of  $89^\circ$ . In the  $xo\bar{z}$  plane, the pattern is offset by  $10^\circ$  with a beamwidth of  $87^\circ$ . At 2.68 GHz, the LHCP pattern is symmetrical in the  $yo\bar{z}$  plane with a beamwidth of  $57^\circ$  and offset by  $5^\circ$  in the  $xo\bar{z}$  plane with a beamwidth of  $61^\circ$ . The measured antenna gain varies from 3.9 to 4.4 dBic in the lower band and from 2.8 to 3.8 dBic in the upper band.

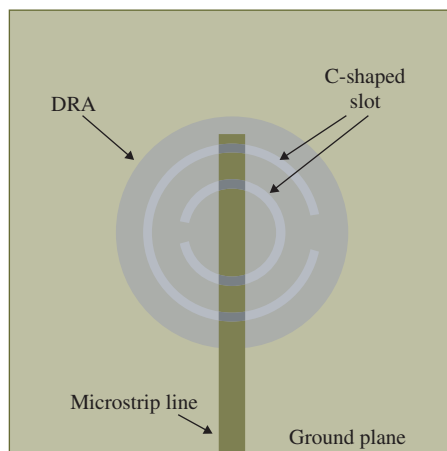
Another slot antenna with dual-band dual-sense circular polarizations is proposed in [71]. The embedded annular slot and cross slot can provide dual-band operations. The lower and upper bands are controlled by the annular slot and the cross slot respectively. Circular polarizations at the two frequencies are achieved by introducing asymmetry on both annular and cross slots. The microstrip feedline consists of a  $50\text{-}\Omega$  transmission line and an impedance transformer which can transform the slot impedance to the required input impedance. The obtained 3-dB AR bandwidths of the prototype using a substrate of  $\epsilon_r = 4.2$ ,  $h = 1.45$  mm are 5.92% at 2.5 GHz (RHCP) and 2.46% at 6.5 GHz (LHCP). The frequency ratio of two frequencies is 2.6. Another prototype fabricated on a substrate of  $\epsilon_r = 2.2$ ,  $h = 3$  mm can



**Figure 4.62** Measured and simulated radiation patterns at 1.64 GHz (a) xoz plane and (b) yoz plane. Measured RHCP \_\_\_\_ Simulated RHCP \_\_\_ Measured LHCP ..... Simulated LHCP \_.\_. Reproduced with permission of © 2011 IEEE



**Figure 4.63** Measured and simulated radiation patterns at 2.68 GHz (a) xoz plane and (b) yoz plane. Measured LHCP \_\_\_ Simulated LHCP \_\_\_ Measured RHCP ..... Simulated RHCP \_...\_. Reproduced with permission of © 2011 IEEE



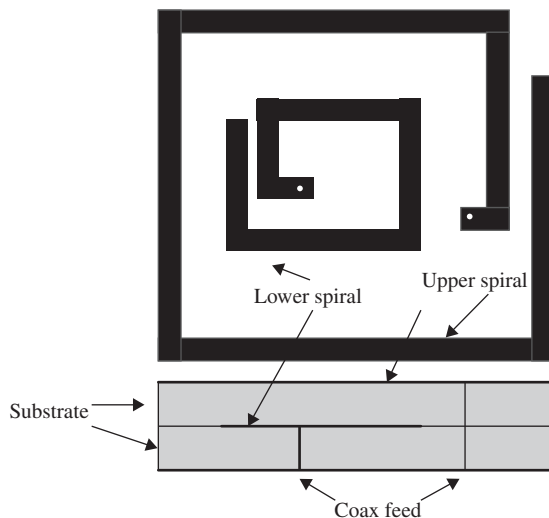
**Figure 4.64** Geometry of a dual-band dual-sense CP DRA antenna [72]. Reproduced with permission of © 2003 IEEE

achieve an improved CP bandwidth and a smaller frequency ratio. The measured 3-dB AR bandwidths for this prototype are 5.54% at 2.4 GHz (RHCP) and 4.29% at 4.9 GHz (LHCP). The obtained frequency ratio is 2.04.

The configuration of a dual-band CP DRA is illustrated in Figure 4.64, in which the dielectric resonator has a high relative permittivity of 79. The DRA is excited by two concentric annular slots in the ground plane through the coupling of the microstrip line. In order to obtain circular polarizations at both frequency bands, both annular slots are perturbed by a shorted section. Due to the opposite directions of the shorted sections for the annular slots, the dual-band dual-sense circular polarizations can be achieved. The outer C-shaped slot can provide LHCP for the lower frequency band, while the RHCP in the upper band can be controlled by the inner C-shaped slot.

Figure 4.65 presents the geometry of a dual-band CP loop antenna using dual-spiral. The monofilar spiral is wound to be approximately one guided wavelength long at the desired frequency and can behave as a one-wavelength electric current loop supporting CP radiation. As observed, the antenna has two layers and two radiating spirals are stacked on the top of each foam substrate. The outer spiral is printed on the upper substrate with a rotation sense yielding RHCP radiation at 3.0 GHz. The inner spiral is printed on the lower substrate and can generate LHCP radiation at 5.4 GHz. It is noted that each spiral is excited by a separate coax feed. The measurement demonstrates a 5.5% 3-dB AR bandwidth at 5.4 GHz and a 3.3% 3-dB AR bandwidth at 3.0 GHz instead of 7.1% and 5.8% for the numerical results. The study in [73] also shows that the AR bandwidth increases significantly (24% to 31%) when two antennas are combined in a sequentially rotated arrangement.

More recently, a dual-band aperture-coupled RFID reader antenna integrated with a meta-material branch line coupler to obtain dual-band dual-sense circular polarization is studied in [74]. The patch antenna is excited by the two output lines of the dual-band branch line coupler through two slots in the ground plane. The slots in the ground plane are located in a T-shape to increase the isolation. A cross slot is embedded at the centre of the patch to



**Figure 4.65** Geometry of a dual-band dual-spiral antenna [73]. Reproduced with permission of © 2003 IEEE

enhance isolation. When one port is excited, the LHCP is generated, whereas the RHCP is obtained when the other port is fed.

#### 4.7.4 Frequency Reconfigurable CP Antennas

Recently, CP antennas which have frequency reconfigurable capability have been investigated in [75–78]. By using RF-microelectro-mechanical system (MEMS) switches [77] or pin diodes [75,76], the antennas can radiate the same CP fields at the selected dual frequencies. The tuning frequency range can be enlarged by varying the capacitance values of the adopted capacitors, as studied in [78]. A probe-fed CP patch antenna is studied in [75]. The circular polarization is obtained by adding two rectangular stubs on the patch. Two diode controlled slots are embedded in the patch to achieve dual-band operations. The frequency ratio is determined by the slot length. As the slot length increases, the frequency ratio is also increased. The slot position also has an effect on the input impedance: the closer the distance between slots and feeding probe, the higher the impedance. It is noted that in the experiment, removable metal tapers are utilized to represent the on and off states of the diodes. When the diode is on, the current can flow directly through the diode and the antenna will resonate at the upper frequency. When the diode is off, the current will flow around the slots and a longer electrical path is introduced, thus the antenna will operate at the lower frequency.

## 4.8 Summary

This chapter provides a review of multi-band CP antennas. Different design techniques and a variety of multi-band CP antennas have been discussed, including multi-band CP patch antennas, multi-band CP QHAs and PQHAs, multi-band CP slot antennas, multi-band CP

**Table 4.4** Comparisons among different multi-band CP antennas

| Ant. Type     | Multi-band techniques                |                                 | Number of bands | Bandwidth | Height | Fabrication Complexity |
|---------------|--------------------------------------|---------------------------------|-----------------|-----------|--------|------------------------|
| Patch antenna | Single feed                          | Stacked patches [1–10]          | 2-4             | Narrow    | Medium | Medium                 |
|               |                                      | Slot loading [12–17]            | 2               | Narrow    | Low    | Low                    |
|               |                                      | Coplanar parasitic patches [18] | 2               | Narrow    | Low    | Low                    |
|               | Multi-feed                           | Stacked patches [19–22]         | 2               | Wide      | Medium | Medium                 |
|               |                                      | Coplanar parasitic patches [23] | 2               | Narrow    | Low    | Medium                 |
|               |                                      | Meta-material [24,25]           | 2               | Narrow    | Low    | Medium                 |
| Helix         | QHA                                  | Multiple QHAs [30–32]           | 2-3             | Narrow    | Large  | Medium                 |
|               |                                      | Hybrid coupler [33]             | 2               | Narrow    | Large  | Medium                 |
|               |                                      | Lumped element [34–36]          | 2               | Narrow    | Large  | Medium                 |
|               | PQHA                                 | Multiple fingers [37]           | 2               | Narrow    | Large  | Medium                 |
|               |                                      | Folded PQHA [38–41]             | 2-3             | Narrow    | Large  | Medium                 |
|               |                                      | Inverted-F monopoles [43,44]    | 2               | Narrow    | Medium | Medium                 |
| Slot antenna  | Dual monopole feed [46,47]           |                                 | 2-3             | Wide      | Low    | Low                    |
|               | L-shaped feed [48]                   |                                 | 2               | Wide      | Low    | Low                    |
|               | Concentric ring slots [49,50]        |                                 | 2               | Narrow    | Low    | Low                    |
|               | Uni-directional slot antenna [51,52] |                                 | 2               | Narrow    | Medium | Medium                 |
| DRA           | Hybrid antenna [54–56]               |                                 | 2               | Narrow    | Medium | Medium                 |
|               | Multiple modes [59]                  |                                 | 2               | Narrow    | Medium | Medium                 |
| Loop antenna  | Dual-loop [60]                       |                                 | 2               | Narrow    | Medium | Low                    |
|               | Dual-mode [61]                       |                                 | 3               | Narrow    | Medium | Low                    |

DRAs, multi-band loop antennas, and so on. To give a general idea of multi-band CP antennas, Table 4.4 summarizes the performance of some typical multi-band CP antennas. As shown, the multi-feed stacked patches in [19–22] can achieve multi-band CP operation with a broadband performance within each band, at the expense of complexity in antenna designs. Single-feed slot-loaded patches [7] and the dual-loop antenna in [60] are attractive for dual-band CP operations due to their advantages of a simple structure and easy fabrication. As the multi-band CP antenna is a hot topic and there are a vast amount of antennas available in the literature, only a limited number of antennas are included in Table 4.4. Thus the table provides an initial idea only but not the whole picture.

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# 5

## Circularly Polarized Arrays

### 5.1 Introduction

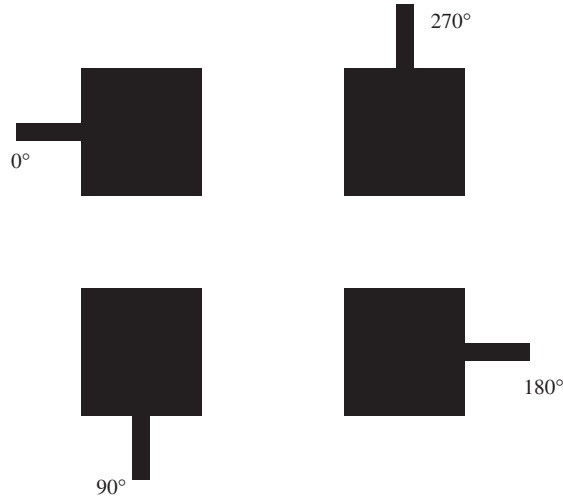
Previous chapters have already discussed the basics of CP antennas as well as various techniques for designing small CP antennas, broadband CP antennas and multi-band CP antennas. Array antennas are useful for applications where a high gain, low sidelobe beam reconfigurability are required, such as satellite communications, DBS, WLAN and space communications. CP arrays are important for applications that need a high gain to overcome the free-space loss due to the long distance between radio transmitter and receiver. This chapter will review different techniques for designing CP array antennas, including CP patch arrays, CP DRA arrays, CP printed slot arrays, CP waveguide slot arrays, CP reflectarrays, CP active integrated arrays and CP arrays with beam reconfigurability. Basic principles of designing CP arrays will be illustrated together with examples. Performance of some CP arrays is compared at the end of this chapter.

### 5.2 CP Patch Antenna Arrays

#### 5.2.1 Sequential Rotation

Microstrip patch arrays are widely used in wireless communications due to their advantages of low profile, light weight, easy fabrication, low cost and conformability to curved structures [1]. CP arrays can be formed by using a number of CP or LP radiating elements. Each antenna element can be single-feed or dual-feed in phase quadrature. Better polarization purity can be achieved if four feeds are employed for each patch element as the undesired modes in the patch can be suppressed. However, the four-feed patch requires a more complicated feed network to provide four ports with  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  phase differences. For large CP arrays, these feed networks will occupy lots of space, leading to increased complexity and high cost.

A technique that uses single-feed LP patches to design CP arrays is proposed in [2]. Figure 5.1 shows a CP array formed by using four linearly polarized elements. As shown, four elements are sequentially rotated and have a phase distribution of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$



**Figure 5.1** Configuration of the conventional CP patch antenna array using sequential rotated LP patch elements [2]

and  $270^\circ$ . This technique enables a significant reduction in the size, complexity, weight and RF loss of the array feed networks, and is particularly attractive for applications in large arrays.

Using this technique to design wide bandwidth CP microstrip patch arrays is also studied in [3], where the radiation performance of a CP microstrip patch array with an antenna element of either linear polarization or circular polarization is investigated. In [3], the radiation performance of the sequentially rotated array is analysed by using its corresponding transmission line equivalent circuit model. In this analysis model, the equivalent circuit model of a LP patch antenna is described as a parallel circuit consisting of one inductor, one capacitor and one resistor. For the CP patch antenna, it is modelled as a series connection of two LP patches. According to [3], the input admittance ( $Y_p$ ) of a LP patch can be calculated by:

$$Y_p = G \left\{ 1 + jQ \left( \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega} \right) \right\} \quad (5.1)$$

where  $G$  is the patch resonance conductance,  $Q$  is quality factor,  $\omega$  is the patch operating frequency and  $\omega_r$  is the resonant frequency of the patch. For a CP patch, its input admittance can be calculated by:

$$Y_p = \frac{2G^2 - B^2 + 2jGB}{2G + jB} \quad (5.2)$$

where

$$B = GQ \left( \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega} \right) \quad (5.3)$$

Details of CP arrays analysis using sequentially rotated LP elements and CP elements are presented in [3]. It is shown that clusters of three sequentially-rotated CP patch elements, in which case a triangular lattice configuration is employed, achieve broader bandwidth compared to that of clusters of two or four elements. Also the gain of sequentially-rotated

arrays can benefit from using the CP elements. This implies that a wideband CP antenna array can be achieved by having a low-Q antenna as the radiating element and with a sequential rotation technique.

One concern of using a LP antenna to design CP antenna arrays is the increased cross-polarization level. After investigating a  $2 \times 2$  sequential rotated CP microstrip antenna, the author in [4] points out that even using the LP antenna as the radiation element for a CP antenna array, low cross-polarization within the bandwidth can still be reached if there is a high-performing feed network that can provide accurate phase and amplitude for the array. Meanwhile, the mutual coupling between antenna elements also plays an important role in increasing the cross-polarization of the array. In the case where the antenna element is spaced with  $0.6\text{--}0.7\lambda$ , the contribution from the radiators to the cross-polarization is at least 10–12 dB [4].

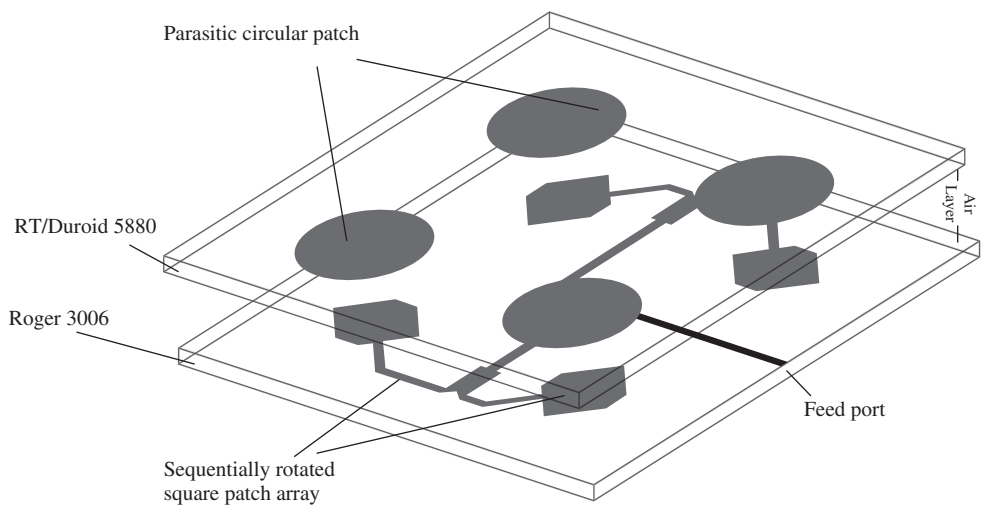
### 5.2.2 Broadband CP Patch Antenna Arrays

It is always desirable to have an antenna array of broad bandwidth, including both the impedance and AR bandwidth. In this section, some typical techniques that can be employed to increase the bandwidth of the CP antenna array using either single or dual-feed patches will be presented.

#### 5.2.2.1 Single-Feed Patch

As discussed in Section 5.2.1, the sequentially rotated technique is an effective method to improve the bandwidth of the CP antenna array. If there is a need to further increase both the impedance and AR bandwidth of the CP antenna array, it can be realized by introducing the parasitic radiators to a sequentially rotated antenna array, like the one proposed in [5]. The layout of this array is presented in Figure 5.2. This CP array is designed to resonate at 10.2 GHz. Circular patches, which are printed on RT/Duriod 5880 ( $\epsilon_r = 2.2$ ) with a thickness of 0.79 mm, are placed at a height of  $0.11\lambda_{10.2\text{ GHz}}$  above the quasi-elliptical patches as the parasitic radiator. The four sequentially rotated square patch antennas are printed on 0.63-mm thick RO3006 ( $\epsilon_r = 6.15$ ) and are fed with  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  phase differences by employing a microstrip feed network, as shown in Figure 5.2. Two corners of each square patch are truncated to generate the required CP operation and the 50- $\Omega$  microstrip line is located  $45^\circ$  clockwise relative to the major axis of the square patch, which results a RHCP radiation. To reach the LHCP, the feeding line can be changed to  $45^\circ$  counterclockwise relative to the major axis of the square patch.

The measurement results of the single antenna element (square patch with the parasitic element) and a  $2 \times 2$  subarray with sequential rotation is summarized in Table 5.1. As can be seen from this table, the single antenna element exhibits 10-dB return loss bandwidth of 18.5% and 3-B axial ratio bandwidth of 7.5% with central frequency of 10.2 GHz. Compared to the conventional microstrip antennas, which normally only have a bandwidth of only a few percent [6], there is a significant improvement on the antenna operation bandwidth by using the parasitic radiator. After employing the sequential rotation techniques to the  $2 \times 2$  subarray, the 10-dB return loss bandwidth further increases to 27.7% whereas the axial ratio bandwidth improves to 23.5%. The measurement results also show that the gain of the CP array is 12.5 dBi at 10.2 GHz.



**Figure 5.2** Geometry of the wideband stacked CP antenna array [5]

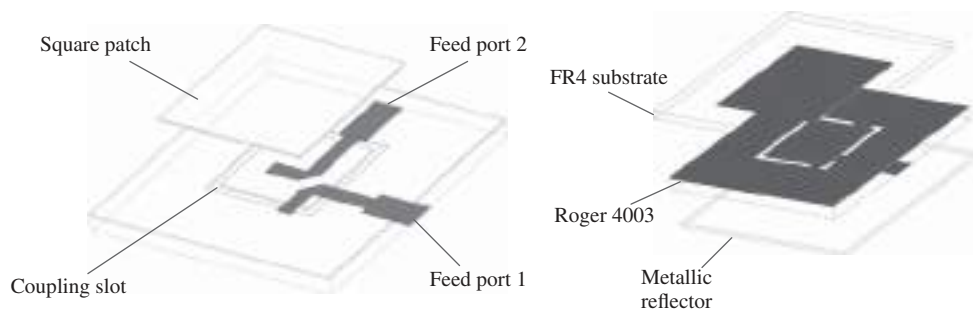
**Table 5.1** Comparison of the radiation performance between single antenna element and its array [5]

|                                                                     | Single antenna element<br>(square patch with a<br>parasitic element) | 2 × 2 subarray (with<br>sequential rotation and<br>parasitic elements) |
|---------------------------------------------------------------------|----------------------------------------------------------------------|------------------------------------------------------------------------|
| 10-dB Return Loss Bandwidth (with<br>central frequency at 10.2 GHz) | 18.5%                                                                | 27.7%                                                                  |
| AR bandwidth(with central frequency<br>at 10.2 GHz)                 | 7.5%                                                                 | 23.5%                                                                  |
| Gain                                                                | 7.5 dBi                                                              | 12.5 dBi                                                               |

From the measurement results given in Table 5.1, it can be concluded that after employing the sequential rotation technique and introducing parasitic radiators, the CP patch array can have both 10-dB return loss and 3-dB AR bandwidth larger than 20%, which is several times larger than a conventional microstrip patch array.

**5.2.2.2 Dual-Feed Patch**

As discussed in Chapters 1 and 2, CP patch antennas can be designed using a single-feed or multi-feed technique. Using a single-feed technique to design CP patch antenna leads to a simpler feed network; however, without using any bandwidth enhancement techniques, it has the inherent disadvantages of narrow impedance and AR bandwidth. The use of a dual-feed antenna can solve this issue at the expense of increased system complexity, especially when a large number of antenna elements are used. Investigations have been done in order to apply the dual-feed technique to the CP antenna array design with reduced system complexity. One



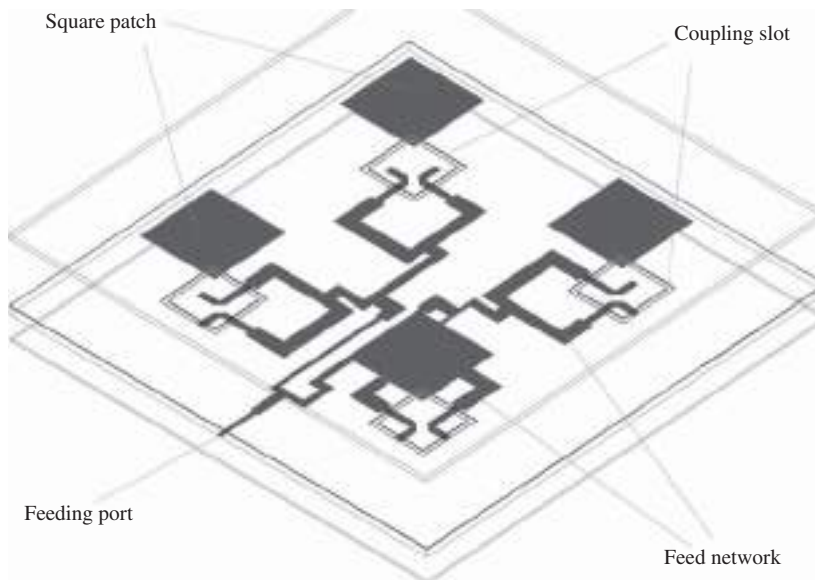
**Figure 5.3** Configuration of the dual-feed slot-coupling CP patch. The figure in the left shows the detailed view of the patch with two feed lines and the figure on the right presents the overall structure of this CP patch [7]

such design is proposed in [7], where the dual-feed slot-coupling feeding technique is used in the design of a wideband CP patch array.

Figure 5.3 shows the configuration of the antenna element proposed in [7]. This antenna array is designed for the operation at WiMax frequency band (3.3–3.8 GHz). A square patch is printed on the bottom side of a 1.58-mm thick FR4 ( $\epsilon_r = 4.4$ ) substrate. One slot ring and two orthogonal microstrip feed lines are printed on each side of a 1.53-mm thick Roger 4003 ( $\epsilon_r = 3.55$ ) substrate. There is an 11 mm air gap between the FR4 and Roger 4003 substrate. One metallic ground plane is added below the feed line to reduce the back radiation. By using the slot coupled feeding technique, the feed line can be printed on different layers and in this way more space for the feed network can be allocated compared to the microstrip-fed antenna arrays that have antennas and the feed network printed on the same layer.

To further increase the bandwidth of the array, a sequential rotation technique is used. The sequential rotation technique is a well-known method that can be used to improve the CP purity, radiation pattern symmetry and impedance as well as AR bandwidth. Therefore, combining both techniques, dual-feed slot coupling and sequential rotation feeding, wideband operation of the CP antenna array can be achieved. Figure 5.4 shows the feed network of the resulting  $2 \times 2$  array. This antenna is designed to have an RHCP radiation and the measurement results show that this antenna array has 3-dB AR bandwidth of 30% with central frequency of 3.55 GHz and a 10-dB return loss bandwidth of 50% (from 2.6–4.4 GHz), which is much larger compared to the conventional CP patch antenna arrays and even larger than the sequentially rotated antenna array with parasitic radiators as presented in Section 5.2.2.1. The measurement results also indicate that the cross polarization is 20 dB less than the co-polarization within the HPBW. The maximum measured gain is about 12.8 dBi.

As can be seen from this design, the use of dual-feed or multi-feed radiating elements in sequentially rotated CP arrays can increase the bandwidth of CP arrays at the expense of increased system complexity, size and losses of the array feed networks. Such arrays are rather difficult to implement on a single layer due to the space problem. The space problem can become more serious if RF and microwave active circuits are required to be integrated with antenna elements and feed networks. It is possible to alleviate this space problem of CP arrays by using a slot-coupled multi-layer structure that is, the patch antenna and feed network are placed at different layers, such as the one shown in Figure 5.4.



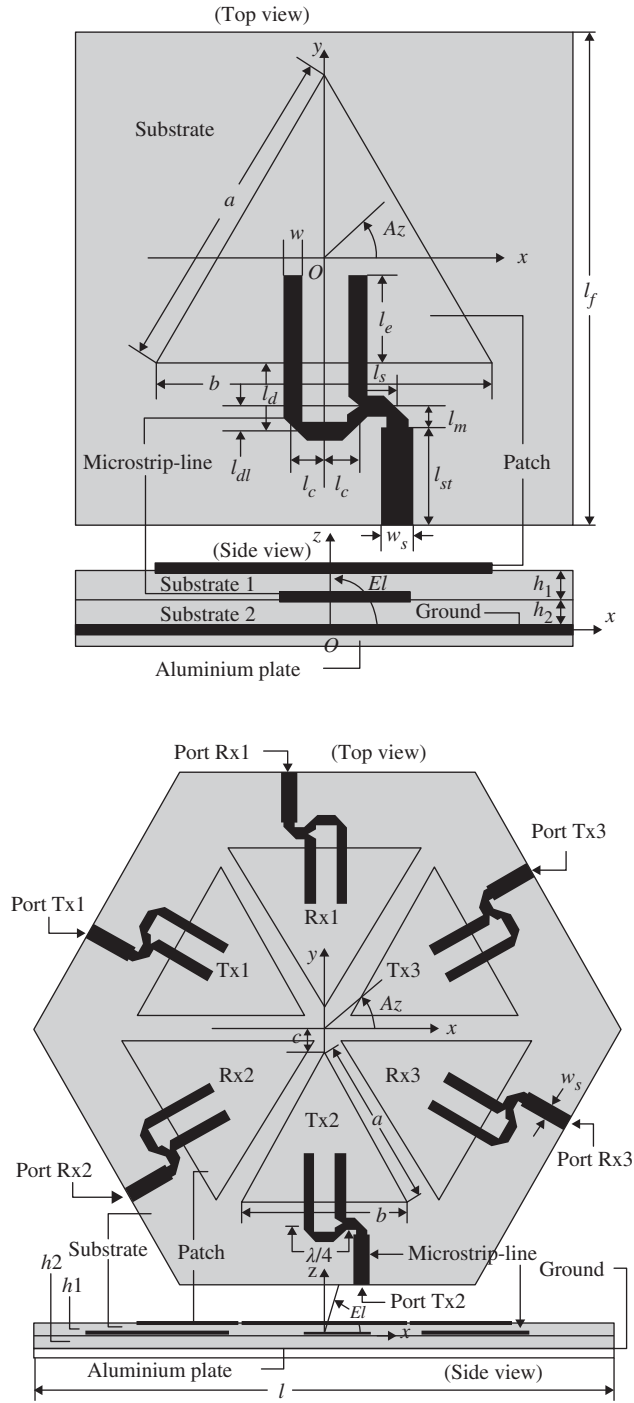
**Figure 5.4** The layout of the  $2 \times 2$  sequentially rotated CP array [7]

### 5.2.3 Multi-Band CP Patch Antenna Arrays

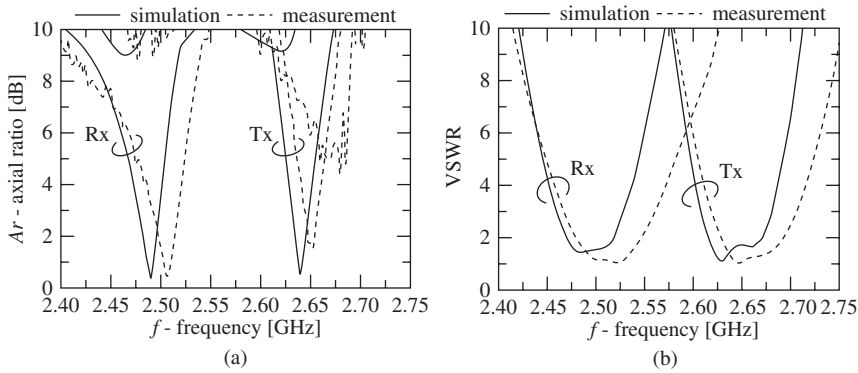
The multi-band CP antenna array gains its application in the field of satellite communications, where the transmitting and receiving bands are allocated with different frequencies [8,9], and RFID applications, where one high gain RFID reader is required to operate at different RFID bands [10]. The difficulty in designing a multi-band CP patch array lies in the fact that at different resonant frequencies, the corresponding mode must have two orthogonal components with equal amplitude and phase quadrature, conditions which are not easy to meet simultaneously.

One approach to design a multi-band CP patch antenna array is to place different single band CP antennas in the same array lattice and thus the overall antenna array can operate at different frequency bands. The dual-band CP equilateral triangular patch array reported in [8] is given as one example. Figure 5.5 shows the configuration of the triangular patch antenna and the layout of the dual-band antenna array. This antenna array is built for satellite tracking for ground applications in Japan. The requirement is to have a CP antenna array that works at 2.5 GHz for reception and 2.65 GHz for transmission. As can be seen from Figure 5.5, the triangular patch is fed by using proximity dual-feed, one of which is  $\frac{1}{4}$  longer than the other to create a  $90^\circ$  phase delay. With the feed configuration shown in this figure, the triangular patch can radiate LHCP waves. Both the patch and the feed line are printed on a 0.8-mm thick substrate with a relative permittivity of 2.17.

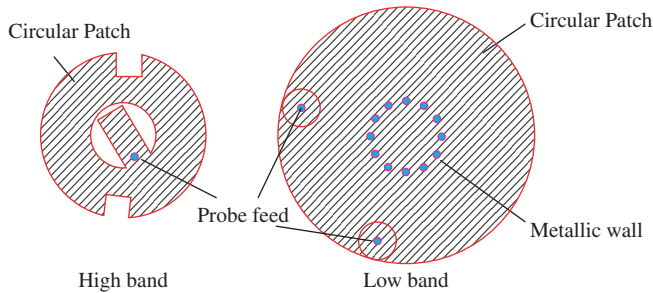
To make the triangular patch resonate at 2.5 GHz, the side length of the triangular patch ( $a$ ) is chosen to be 52.5 mm whilst by letting this length to be 49.4 mm, the central frequency of this triangular patch can shift to 26.5 GHz. As presented in Figure 5.5, both the Tx and Rx sections are composed of three patch antennas and the array elements are arranged to minimize the total size of the antenna array. Figure 5.6 shows the measured and simulated VSWR,



**Figure 5.5** Configuration of the single triangular patch antenna and the resulting dual-band antenna array [8]. Reproduced with permission of © 2005 IEEE



**Figure 5.6** Measured and simulated VSWR, axial ratio of the dual-band CP triangular antenna array [8]. Reproduced with permission of © 2005 IEEE



**Figure 5.7** The plain view of the radiation elements for higher (left) and lower (right) bands developed by EADS CASA ESPACIO for the Navigation antenna of a Galileo constellation [9]. Reproduced with permission of © 2007 EurAAP

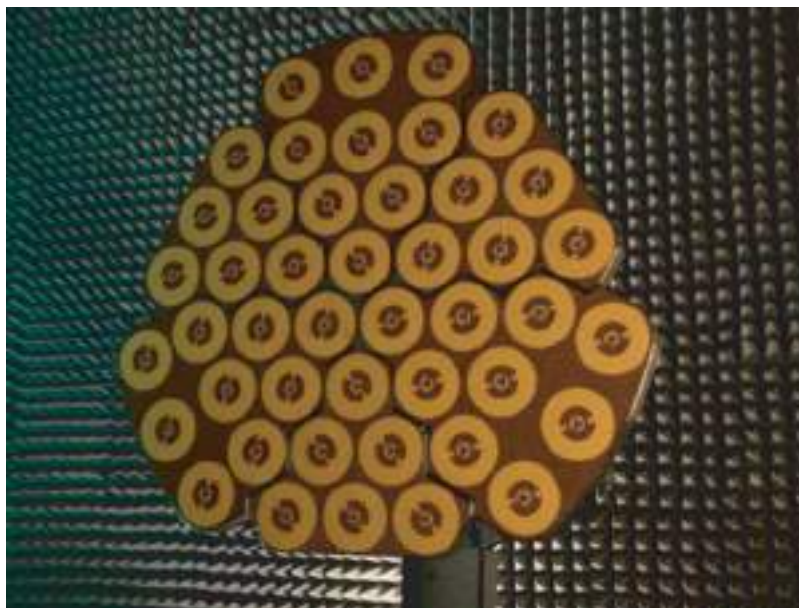
axial ratio of the CP antenna array. The measurement results indicate that the antenna array can operate at 2.5 and 2.65 GHz with a good axial ratio. For both the Tx and Rx antenna, the measured gain is about 6 dBi.

Another technique that can be employed to design a multi-band CP antenna array is to use multi-band antennas as the radiating elements. One of the options is to use antennas with a stacked structure and patch antennas of different resonant frequencies that can be printed at different layers and fed independently. One such design is reported in [9], which aims to design a dual-band CP antenna array for Galileo system navigation at the L-band (1.15–1.6 GHz). For this system, it is required to have an antenna array operate at 1145–1299 MHz and 1555–1595 MHz with RHCP. Figure 5.7 presents the layout of the stacked circular patch antenna for lower and higher bands.

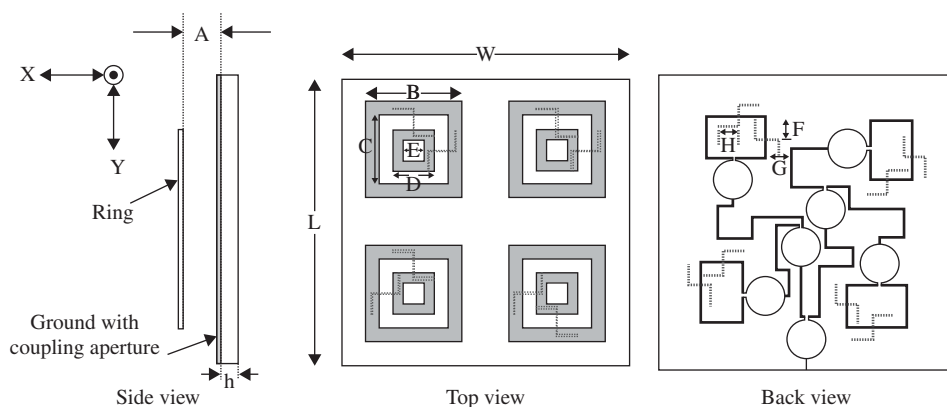
The top circular patch is fed by a single coaxial pin and two notches are embedded on the patch to create the RHCP radiation. The patch for lower band operation is placed below the higher band patch and fed with two coaxial pin with phase quadrature to obtain CP radiation. The circular patch for the higher band is short-circuited at the inner periphery with the purpose of keeping minimum interference between both circular patches. Besides this, a circular metallic wall is used to surround the patches in order to shield the radiators in the array and thus have good isolation between adjacent elements. The measurement results

show that this dual-band CP array has gain of 15 dBi at both bands and within the required frequency bands, the return loss is always better than 15 dB. Figure 5.8 shows a photo of this stacked patch array under measurement.

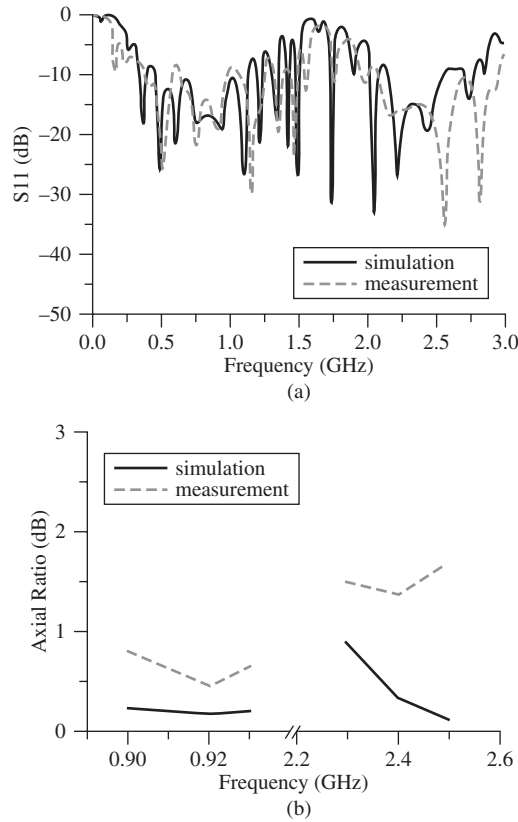
A high-gain CP dual-band antenna array for RFID reader applications is recently proposed by [10]. Different from the two approaches presented previously, a single layer dual-band dual-ring radiator is employed as the array element. Figure 5.9 shows the side, top and back view of this dual band CP antenna array. The antenna element consists of two rings, each of which can resonate at different frequencies: 900 MHz and 2.4 GHz. The



**Figure 5.8** A photo of the stacked multi-band CP array for Galileo system navigation developed by EADS CASA ESPACIO under measurement [9]. Reproduced with permission of © 2007 EurAAP



**Figure 5.9** Side, top and back view of the dual band CP antenna array for RFID reader application [10]. Reproduced with permission of © 2012 IEEE



**Figure 5.10** Measured and simulated S11, axial ratio of the dual band CP antenna array for RFID reader application [10]. Reproduced with permission of © 2012 IEEE

dual-band antenna is excited with  $90^\circ$  phase offset through apertures. Wilkinson power dividers are used to distribute the excitation from the input to the individual elements. Since the frequency ratio for the dual band antenna is close to 3:1, it is possible to maintain the phase offset at both band to the desired value. To improve the CP performance, the antenna elements are sequentially rotated.

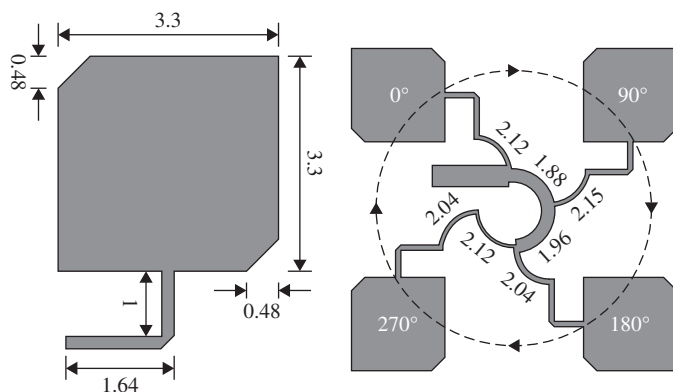
This antenna is fabricated on a 0.508-mm thick FR4 substrate. Figure 5.10 presents the measured and simulated S11, axial ratio of this dual band CP antenna array. There is a good impedance matching at both bands and the axial ratio at the desired frequency bands is always below 2 dB. At the lower band, 8 dBi peak gain is reached while at the higher band, the peak gain is about 10 dBi.

#### 5.2.4 High-Efficiency CP Patch Arrays at the Ku Band and Above

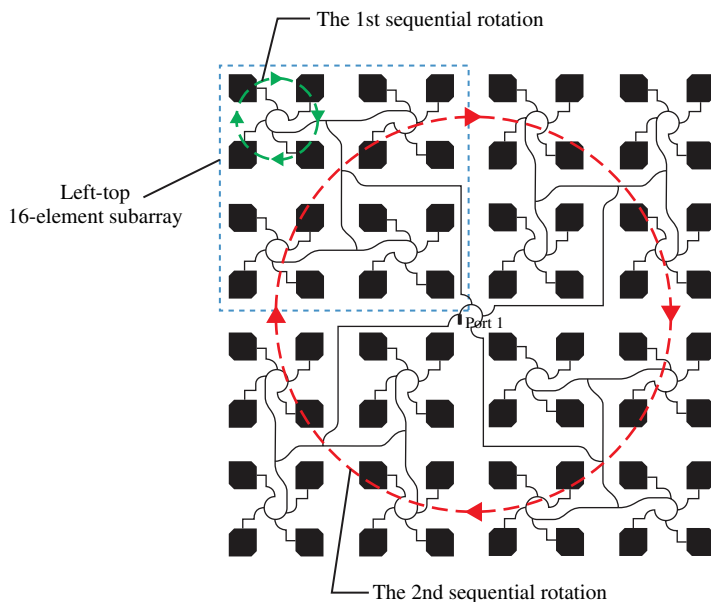
When designing a microstrip antenna array consisting of a large number of elements with the microstrip feeding lines printed on the same layer, the loss and undesired radiation from the feed network are problematic, especially for the application at millimetre-wave frequencies. To reduce these unwanted effects from the feed lines, besides using the aperture coupled

feeding like the one already presented, several other approaches have also been proposed. One effective method is the double use of sequential rotation technique in large arrays. A 64-element wideband CP microstrip antenna array that can operate in the frequency band of 27–31 GHz is presented in [11]. The antenna element is a square-shaped patch with two truncated corners and the four-element subarray is arranged with sequential rotation to achieve better CP bandwidth, as shown in Figure 5.11.

Then, the subarray is used to form a full array containing a total number of 64 elements, which is shown in Figure 5.12. As can be seen from this figure, different to the conventional

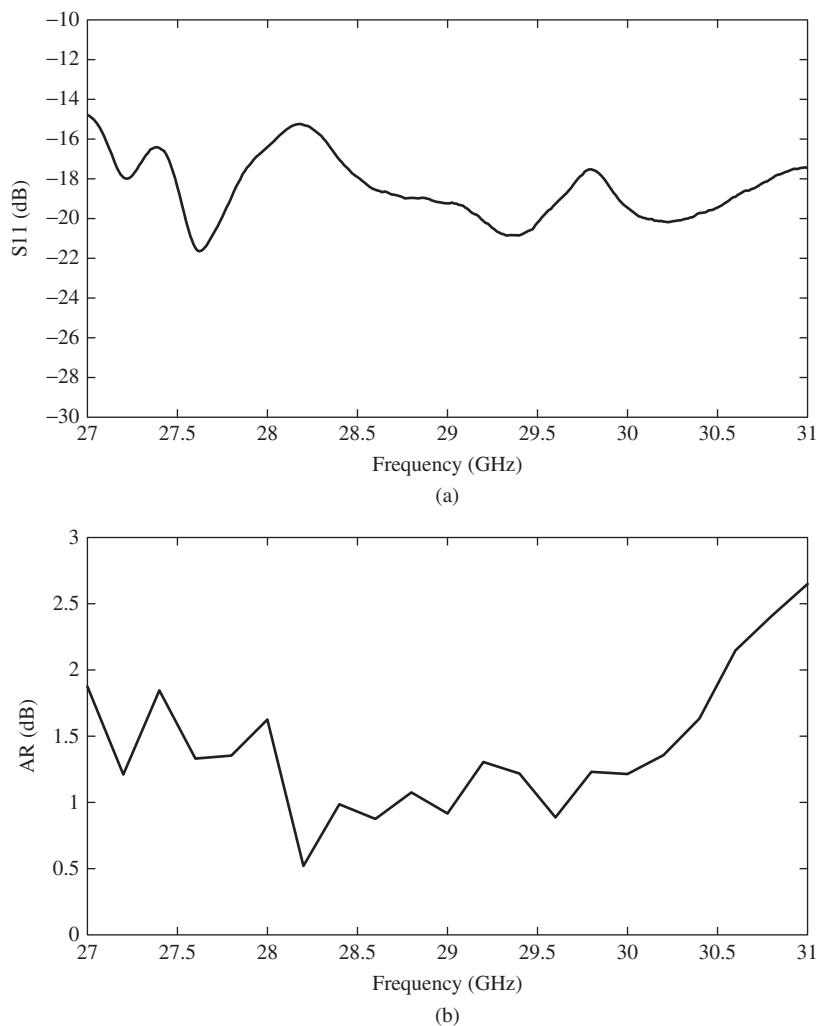


**Figure 5.11** Geometry of the single radiation element and configuration of the subarray [11]. Reproduced with permission of © 2011 IEEE

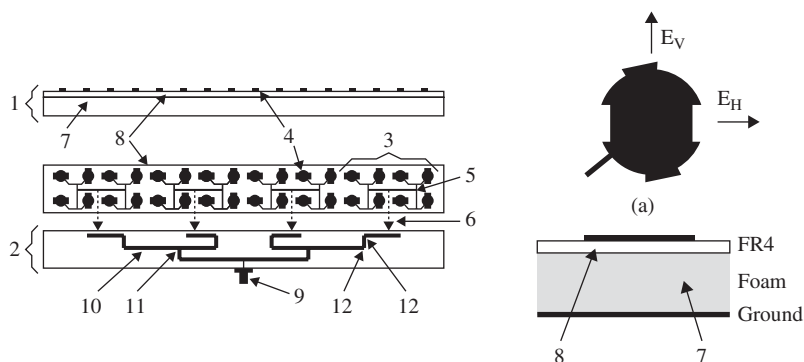


**Figure 5.12** Configuration of the full array using the double sequential rotation technique [11]. Reproduced with permission of © 2011 IEEE

array configuration where the subarrays are arranged symmetrically, the  $4 \times 4$  subarray is arranged in a clockwise rotation sequence. According to [11], this double sequential rotation can cancel the undesired radiation from the feed network and thus contributes to the polarization purity and radiation pattern symmetry. This microstrip antenna array is printed on a 0.254-mm thick RT/Duriod 5880 ( $\epsilon_r = 2.2$ ) and the distance between each antenna element is  $0.77\lambda_{29\text{ GHz}}$ . Low permittivity dielectric material is chosen to obtain good AR and impedance bandwidth. Figure 5.13 shows the measured S11 and axial ratio of this 64-element antenna array. The measurement results show that both the 10-dB return loss and 3-dB axial ratio bandwidth cover the frequency band of 27–31 GHz, which corresponds to a wide bandwidth of more than 13.8% with central frequency of 29 GHz. Moreover, the measured radiation patterns indicate that this antenna array exhibits a high polarization



**Figure 5.13** Measured S11 and AR bandwidth of the array using the double sequential rotation technique [11]. Reproduced with permission of © 2011 IEEE



**Figure 5.14** The configuration of the waveguide-fed CP microstrip array and the structure of the single antenna element [12]. Reproduced with permission of © 2005 IEEE

purity and good radiation performance, which means that the unwanted radiation from the feed lines has been largely reduced. The peak gain of this array is around 20 dBi and the efficiency is about 70%.

Besides the methods presented here a high efficiency CP microstrip array can also be obtained by employing waveguide feeding techniques. In this case, instead of using any microstrip line, the antenna elements are excited by waveguides and thus radiation as well as losses from the microstrip line can be avoided. A low-cost, high-efficiency quasi-planar CP array fed by waveguides is reported in [12]. In this work, a  $2 \times 16$  CP array is designed for the Ku-band application. Figure 5.14 shows the configuration of the waveguide-fed CP microstrip array and the structure of the single antenna element. This antenna array is composed of two parts. The first part is the antenna array elements. There are 32 elements in total for the array and the structure of the single antenna element is the one presented in Figure 5.14. The radiating element is a circular patch with two stubs and two notches. It is designed on a 0.1 mm thick FR4 and mounted 1.5 mm above the ground plane. When a microstrip line feeds the element at an angle of  $45^\circ$ , two perpendicular modes can be generated with phase quadrature, thus leading to CP radiation. The second part of the array is the waveguide feed network. To feed the antenna, an additional waveguide-to-microstrip transition is required, which is fully described in [12]. This waveguide-to-microstrip transition will add about 0.4 dB loss in power transfer from the input waveguide to the microstrip feed network. A photo of the fabricated feed network is shown in Figure 5.15.

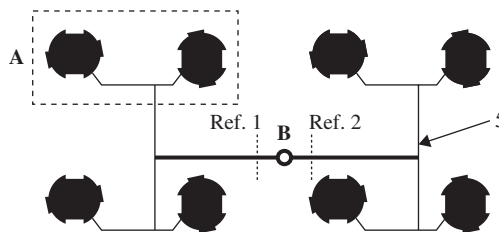
Sequential rotating is employed to improve the CP performance of this array. Figure 5.16 presents the layout of the  $2 \times 4$  subarray. The measurement results indicate that within the required frequency band (12.2–12.7 GHz), it has CP gain of 23 dBi and AR less than 1.1 dB. Moreover, the aperture efficiency has reached 63%, because of the low loss from the feed network.

### 5.3 CP Dielectric Resonator Antenna Arrays

DRA has the advantages of light weight, low cost, compact size and high radiation efficiency at millimetre frequencies. Therefore, using the DRA to design a CP antenna array has also



**Figure 5.15** A photo of the fabricated feed network for the waveguide-fed CP microstrip array [12]. Reproduced with permission of © 2005 IEEE



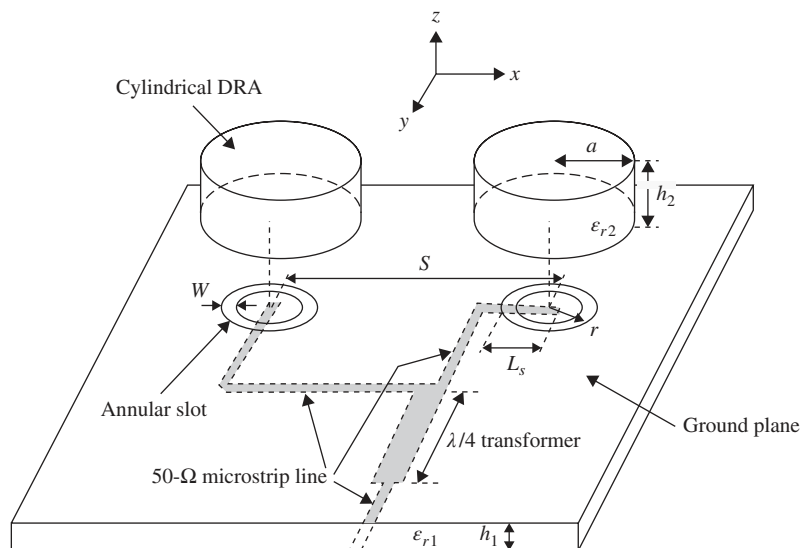
**Figure 5.16** Layout of the  $2 \times 4$  subarray with sequential rotating [12]. Reproduced with permission of © 2005 IEEE

attracted much research interest. Various single-feed CP DRA antennas have been discussed in Chapter 2. Similar to the CP patch arrays, DRA CP arrays can also be realized by using either LP or CP DRA elements. Compared to CP arrays using multi-feed DRA elements, it is easier to employ LP DRA or single-feed CP DRA elements which can reduce the complexity, size and cost of array antennas.

Compared to CP patch arrays, DRA based CP arrays have the potential to achieve higher radiation efficiency at millimetre-wave frequencies and broader bandwidths. Some typical examples will be illustrated in the following subsections.

### 5.3.1 LP Element

Similar to the microstrip patch antenna arrays, CP DRA array can be designed by using LP elements fed with phase quadrature. A CP dielectric resonator antenna array consisted of



**Figure 5.17** Configuration of the CP DRA array consists of two LP DRAs [13]. Reproduced with permission of © 1999 John Wiley & Sons, Inc.

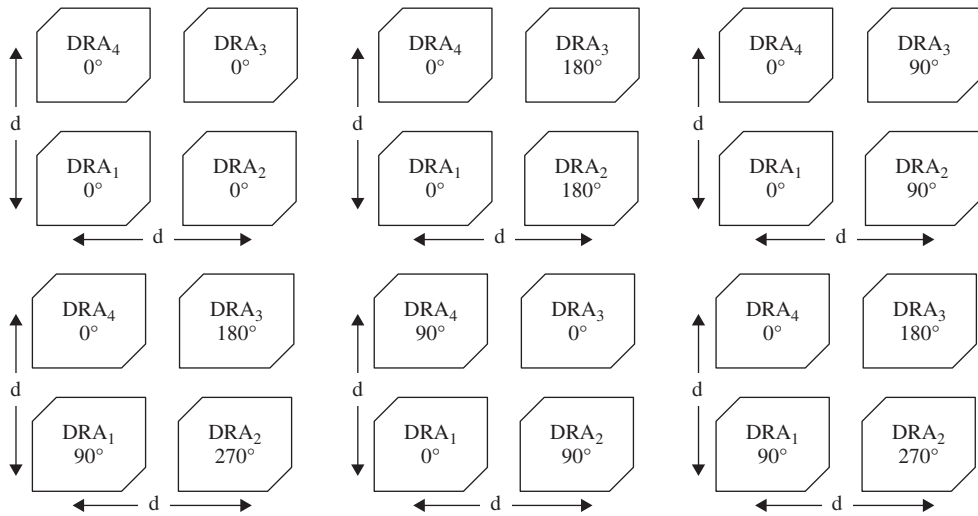
two LP cylindrical DRAs is demonstrated in [13]. The dielectric resonator has a cylindrical shape and is made of dielectric material with relative permittivity of 79. The use of such high permittivity material can contribute to the size reduction of the antenna. The DRA array is designed to operate at 1.5 GHz and the configuration of this two-DRA array is shown in Figure 5.17.

The two dielectric resonators are excited by using two annular slots and the microstrip lines are printed on the bottom side of the substrate, a 1.6-mm thick FR4. A quarter-wavelength impedance transformer is used for the purpose of the impedance matching. The two feeding lines for the DRA elements are designed to have different length, to create the desired 90° phase difference. The feeding configuration shown in Figure 5.17 can create a RHCP radiation; by swapping the two feeding lines, LHCP can be obtained. Measurement results show that this two-DRA array has a 10-dB return loss bandwidth of 5.9% and 3-dB AR bandwidth of 2% with the central frequency at 1.564 GHz. This radiation performance is similar to the conventional microstrip patch antenna arrays.

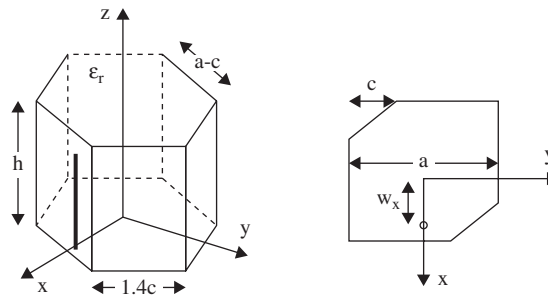
### 5.3.2 CP Radiation Element

Generally speaking, the single-feed CP DRA has a relatively narrow CP bandwidth. This is similar to the single-feed microstrip antenna. Therefore, studies have been carried out to identify the best configuration to improve the bandwidth of the CP DRA array. In [14], the effect of different feeding configurations on the radiation performance of a CP DRA array, namely impedance matching bandwidth, AR bandwidth and radiation patterns, are investigated by studying six different array configurations, as shown in Figure 5.18. The structure of the single DRA is given in Figure 5.19, which is a square-shaped dielectric with two corners removed, for the purpose of creating CP radiation, and is fed by a single coaxial probe feed.

After studying these six different configurations of CP DRA arrays, it is concluded that sequential rotation feeding provides the best impedance bandwidth and the AR bandwidth



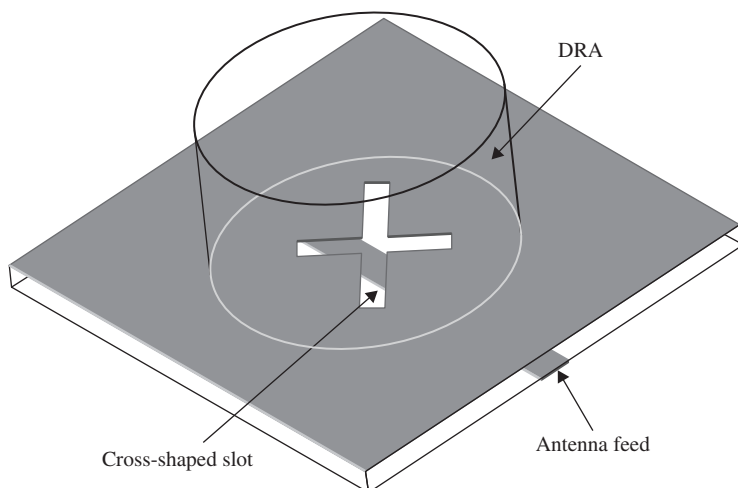
**Figure 5.18** Different configuration for the four elements CP DRA array [14]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.



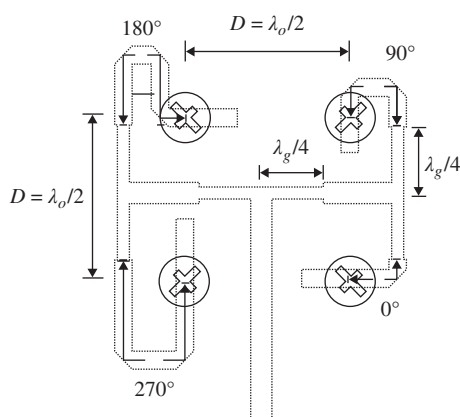
**Figure 5.19** Structure of the single DRA used for the array design [14]. Reproduced with permission of © 2003 John Wiley & Sons, Inc.

[14]. Meanwhile, the broadside CP radiation is frequency independent. This conclusion agrees with the findings for the CP microstrip patch array design. Thus, to improve the CP performance of the DRA array, it is an effective method to employ the sequential rotation technique. A CP dielectric resonator antenna subarray using cylinder-shape dielectric fed by cross-slot aperture is presented in [15]. This DRA array is designed to operate at 4.4 GHz. To achieve wider CP bandwidth, a sequential rotation technique is employed. Figure 5.20 shows the structure of the single CP DRA element and Figure 5.21 presents the configuration of this  $2 \times 2$  array.

The cylindrical dielectric resonator has a relative dielectric constant of 16 with a radius of 5.96 mm and height of 9.82 mm. As depicted in Figure 5.20, a cross slot of unequal length is etched on the copper layer of the substrate ( $\epsilon_r = 2.33$ , thickness = 1.57 mm) beneath the cylindrical dielectric resonator to generate the CP radiation. On the bottom side the substrate,

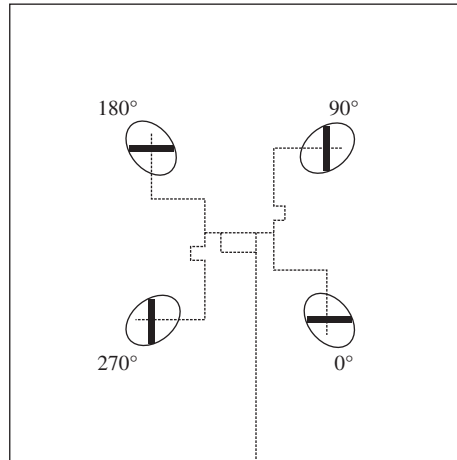


**Figure 5.20** Geometry of the CP DRA element [15]



**Figure 5.21** The layout of the  $2 \times 2$  CP DRA array [15]. Reproduced with permission of © 2000 John Wiley & Sons, Inc.

it is a microstrip line and by varying its length, impedance matching can be optimized. In the CP subarray configuration as shown in Figure 5.21, the DRA elements are sequentially rotated with relative phases of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$ . The distance between each DRA element is 33.5 mm, which is around  $0.5\lambda_{4.4 \text{ GHz}}$ . The measurement results show that this CP DRA subarray has 10-dB impedance bandwidth of 19% and 3-dB AR bandwidth of more than 16% with central frequency of 4.4 GHz, which is several times larger than the bandwidth of single CP element (it is shown in [15] that the single CP element only has 9% impedance bandwidth and 5.6% AR bandwidth). This is contributed by the using of sequentially rotated technique. It is also found that the maximum gain of the CP-element subarray is about 12 dBi.

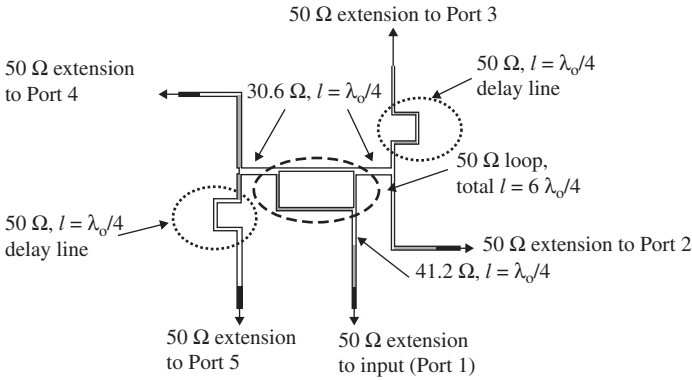


**Figure 5.22** Configuration of aperture coupled  $2 \times 2$  CP DRA subarray with hybrid ring feed network [17]. Reproduced with permission of © 2006 IEEE

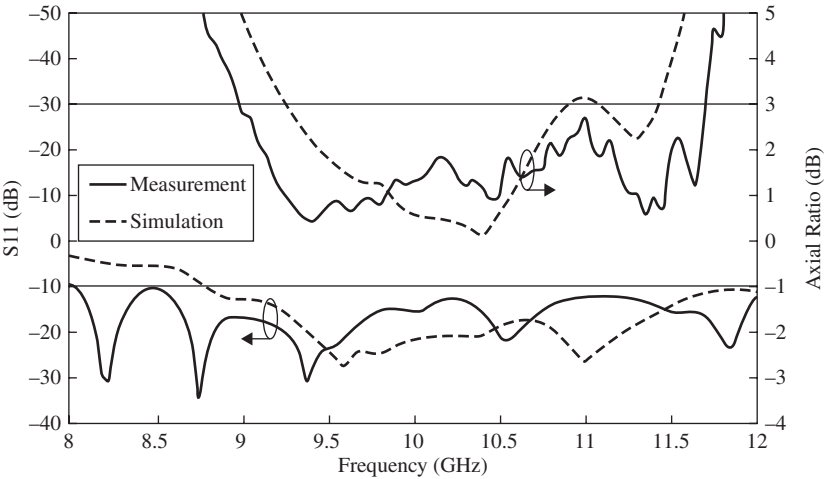
Based on the studies presented in [14], theoretically the AR bandwidth of the sequential feed subarray is independent of frequency when ideal sources are used. However, the feed network usually limits the bandwidth of the overall system, due to the phase variations within the frequency band. As a result, employing different feed networks to improve the AR bandwidth of the antenna array has been investigated. By using the hybrid-ring feed network [16], the bandwidth of the CP DRA array can be improved. A wideband CP elliptical DRA subarray fed by a hybrid-ring feed network is introduced in [17,18]. The hybrid-feed network is well-known for its advantage of exhibiting linear phase difference across a wide impedance bandwidth. Therefore, the use of such hybrid ring feed network in sequential feed CP DRA subarray is studied. Figure 5.22 shows the configuration of an aperture coupled  $2 \times 2$  CP DRA subarray with a hybrid ring feed network.

The antenna used in this subarray is an elliptical dielectric resonator with a relative dielectric constant of 10.2. The DRA is rotated anti-clockwise by  $45^\circ$  to excite two orthogonal modes for RHCP radiation. If the DRA is rotated clockwise by  $45^\circ$ , LHCP radiation can be reached. Under the DRA, there is a 0.254-mm thick substrate with a relative dielectric constant of 10.2. A slot is etched on the ground plane as the aperture to couple the power from the microstrip line that is printed on the other side of the substrate. The details and layout of the hybrid ring feed network are given in Figure 5.23. The feed network includes a hybrid ring and two T-junction power dividers. The output of the hybrid ring is connected to a  $90^\circ$  T-junction power divider to produce sequential rotation phase shifts. For more theoretical explanation of this hybrid feed network, the reader can refer to [16].

The measurement results show that this CP DRA sub-array exhibits a wide bandwidth including both impedance and AR bandwidth: 43.9% 10-dB return loss bandwidth and 26.1% 3-dB AR bandwidth with a central frequency of 10.25 GHz. This result is presented in Figure 5.24. Moreover, it has stable radiation patterns across the AR bandwidth and the cross polarization level is 17 dB lower than the co-polarization at the beam peak. Within the 3-dB AR band, this array has a gain of about 12 dBi. Table 5.2 compares the radiation performance of the two  $2 \times 2$  DRA subarrays proposed in [15] and [17]. As can be seen from the table, a good feed network is critical in improving the overall performance of the CP DRA array.



**Figure 5.23** Layout of the hybrid ring feed network and its corresponding parameters [17]. Reproduced with permission of © 2006 IEEE



**Figure 5.24** Measured and simulated return loss and axial ratio bandwidth of the CP elliptical DRA subarray with hybrid-ring feed network [17]. Reproduced with permission of © 2006 IEEE

**Table 5.2** Comparison of the two DRA subarray

|                             | CP DRA array<br>proposed in [15] | CP DRA array<br>proposed in [17] |
|-----------------------------|----------------------------------|----------------------------------|
| DRA material $\epsilon_r$   | 16                               | 10.2                             |
| 10-dB return loss bandwidth | 19% ( $f_0 = 4.4$ GHz)           | 43.9% ( $f_0 = 10.25$ GHz)       |
| 3-dB AR bandwidth           | 16% ( $f_0 = 4.4$ GHz)           | 26.1% ( $f_0 = 10.25$ GHz)       |
| Polarization                | LHCP                             | RHCP                             |
| Cross-polarization level    | > 15 dB                          | > 17 dB                          |

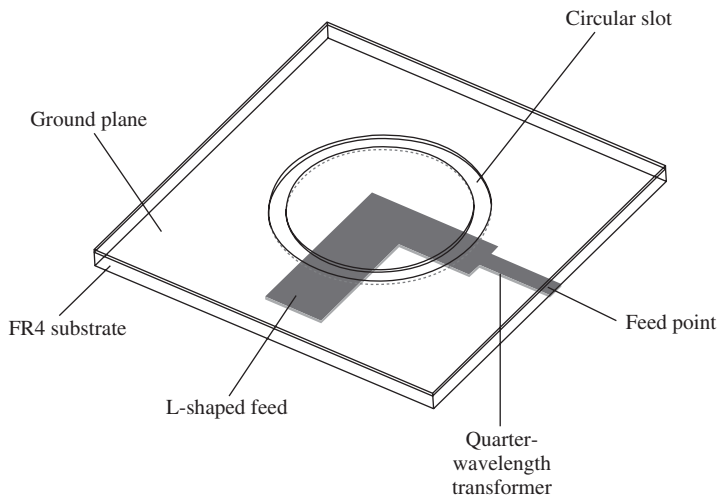
## 5.4 CP Slot Array Antenna

### 5.4.1 Printed Ring Slot CP Arrays

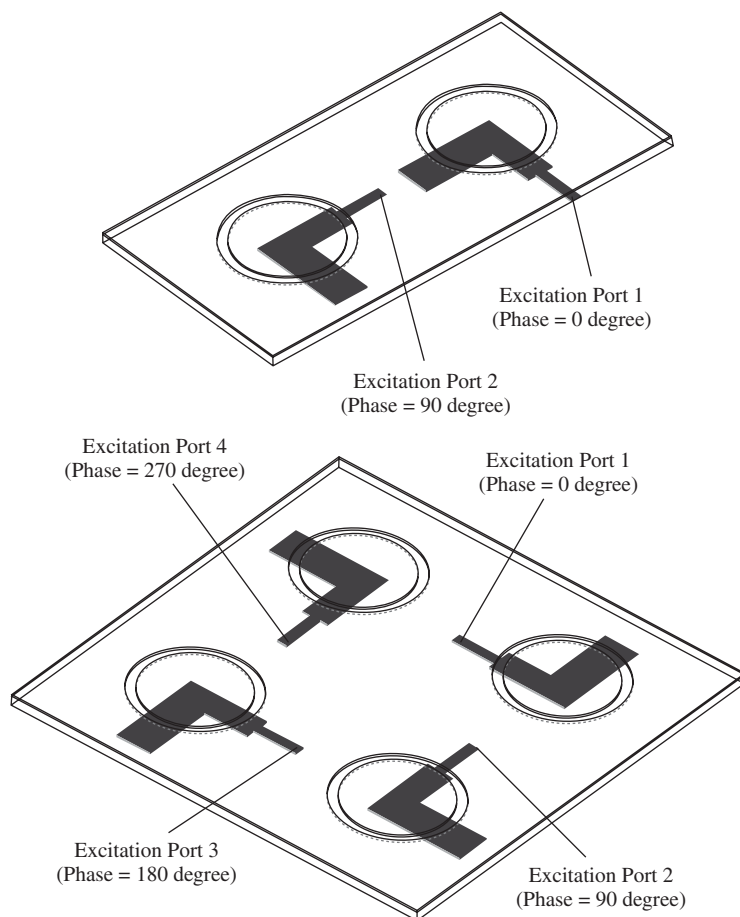
Printed slot antenna arrays have advantages of low profile, simple structure, broadband performance and easy fabrication. Chapters 1–4 have already discussed the basic principles of CP slot antenna designs as well as small slot antennas, broadband slot antennas and multi-band slot antennas. For example, a CP slot antenna can be realized by using a printed square ring slot fed by an L-shaped microstrip feed line. When comes to the printed slot antenna array design, it is important to take the overall system complexity into consideration. Particularly, the feed network should be as simple as possible when large number of array element would be used. Therefore, a printed slot array antenna using a single series microstrip line feed is a preferred configuration.

One broadband printed CP ring-slot array with a simple configuration is proposed in [19]. Figure 5.25 shows the geometry of this ring-slot antenna. This slot antenna is printed on a 1.6 mm thick FR4 substrate and is designed to operate at 2.4 GHz. By letting the circumference of the ring slot be about  $0.7\lambda_{2.4\text{ GHz}}$  long, the slot antenna can resonate at the desired frequency. The slot antenna is fed by a L-shaped microstrip and by optimizing the dimensions of the feed line, two orthogonal modes of the ring-slot antenna with equal amplitude and  $90^\circ$  phase difference can be excited. To get good impedance matching, a quarter-wavelength impedance transformer is employed.

Figure 5.26 presents the layout of the  $1 \times 2$  and  $2 \times 2$  slot antenna array using the slot antennas presented previously as the array elements. Sequential rotation technique is employed to achieve better AR bandwidth. The spacing ( $d$ ) between two adjacent elements is  $0.7\lambda_{2.4\text{ GHz}}$ . The measurement results show that the  $1 \times 2$  slot antenna array has a 10-dB return loss bandwidth of 17.7% and 3-dB AR bandwidth of 8.7% with central frequency of 2.4 GHz. For the  $2 \times 2$  slot antenna array, the 10-dB return loss bandwidth reaches 51%



**Figure 5.25** Geometry of the ring-slot antenna element with single feed [19]

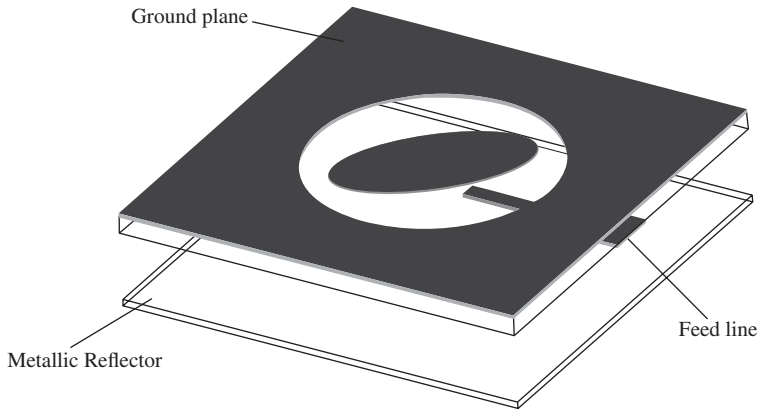


**Figure 5.26** The layout of the  $1 \times 2$  and  $2 \times 2$  slot antenna array with sequential rotated feed network [19]

whilst the 3-dB AR bandwidth increases to 15%. The peak gains of these two arrays are 6.6 dBi and 9 dBi, respectively.

#### 5.4.2 CP Slot Array with Metallic Reflector

The slot antenna has an omni-directional radiation pattern. In some applications where it is required to limit the radiation to a hemispherical space then a reflecting plate needs to be employed. Although that using the metallic reflector can increase the overall height of the slot antenna, on the other hand the feed network can be shielded from the radiation aperture, thus reducing the spurious radiation from the feed. Figure 5.27 shows the configuration of a CP ellipse-loaded circular slot antenna. This slot antenna is designed for millimetre-wave WPAN applications at 60 GHz [20]. This slot antenna consists of a circular slot and an elliptical patch centred within a square ground plane. The antenna element is printed on

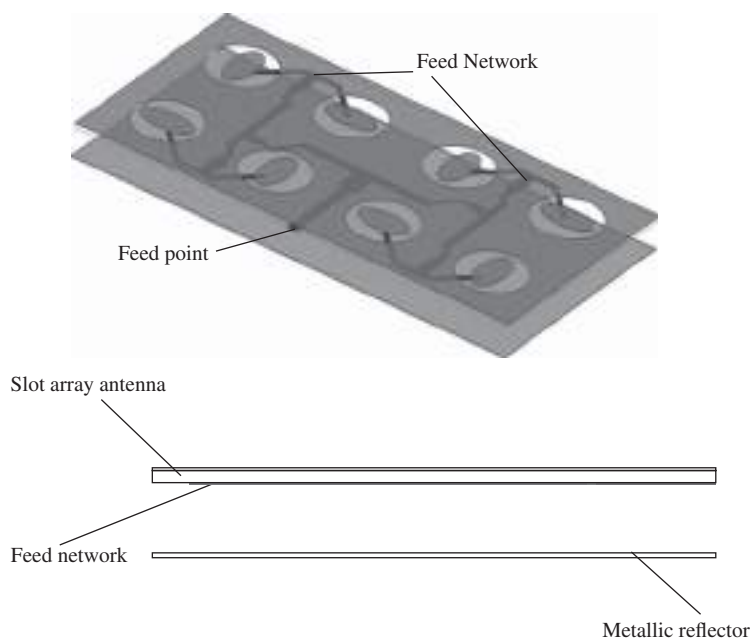


**Figure 5.27** Exploded view of the circularly polarized ellipse-loaded circular slot antenna with a metallic reflector [20]

a substrate with thickness of 0.1 mm and relative permittivity 3.2. A ground plane is located at quarter wavelength below the slot antenna as the metallic reflector.

For CP radiation, the major axis of the ellipse patch is oriented  $45^\circ$  to the microstrip feed line, in order to excite two orthogonal resonant modes with  $90^\circ$  phase shift. The minor-to-major axes ratio of the ellipse is a key parameter in determining the axial ratio of the slot antenna. By varying the length and width of the ellipse, the resonant frequency and coupling between the orthogonal modes can be optimized to obtain good circular polarization. Figure 5.28 presents the configuration of the  $4 \times 2$  ellipse-loaded slot antenna array. This array consists of two identical  $2 \times 2$  sub-arrays. Sequential rotation in both phase and spatial orientation is used to improve the axial ratio bandwidth of this slot antenna array. The feed network consists of T-junction power dividers and a quarter-wavelength transformer. The measurement results show that this antenna array has 3.3 dB AR bandwidth of 34.6% with central frequency of 60 GHz and impedance bandwidth from 50.25–74.5 GHz, which corresponds to a bandwidth of 38.9%. Within the bandwidth, it has a maximum gain of 15.6 dBi.

It is also possible to employ some broadband techniques to further improve the bandwidth (include both the impedance and AR bandwidth) of the CP slot array antenna. For example, adding a parasitic patch above a wide-slot antenna can increase the bandwidth of a slot antenna. One example of such a design is reported in [21]. This slot antenna is designed to operate at X band. The wide circular slot located at the bottom layer and is excited by using the proximity coupling an L-shaped microstrip line. A metallic reflector is located 10 mm below the slot antenna. A circular parasitic patch is printed on the top layer, which can serve as a director to enhance the gain and bandwidth of the slot antenna. A  $4 \times 4$  slot array antenna has been designed using this stacked slot antenna and a corporate feed network is used. The measurement results show that it has a 3-dB AR bandwidth of 21% and the 10-dB return loss bandwidth of 40% with central frequency at 10.05 GHz. The average gain of this array is 18 dBi within the 3-dB AR bandwidth. Although the bandwidth of this CP slot array with parasitic radiator is smaller than the one presented in Figure 5.28, the wideband operation of this array is achieved without using the sequential rotation technique.



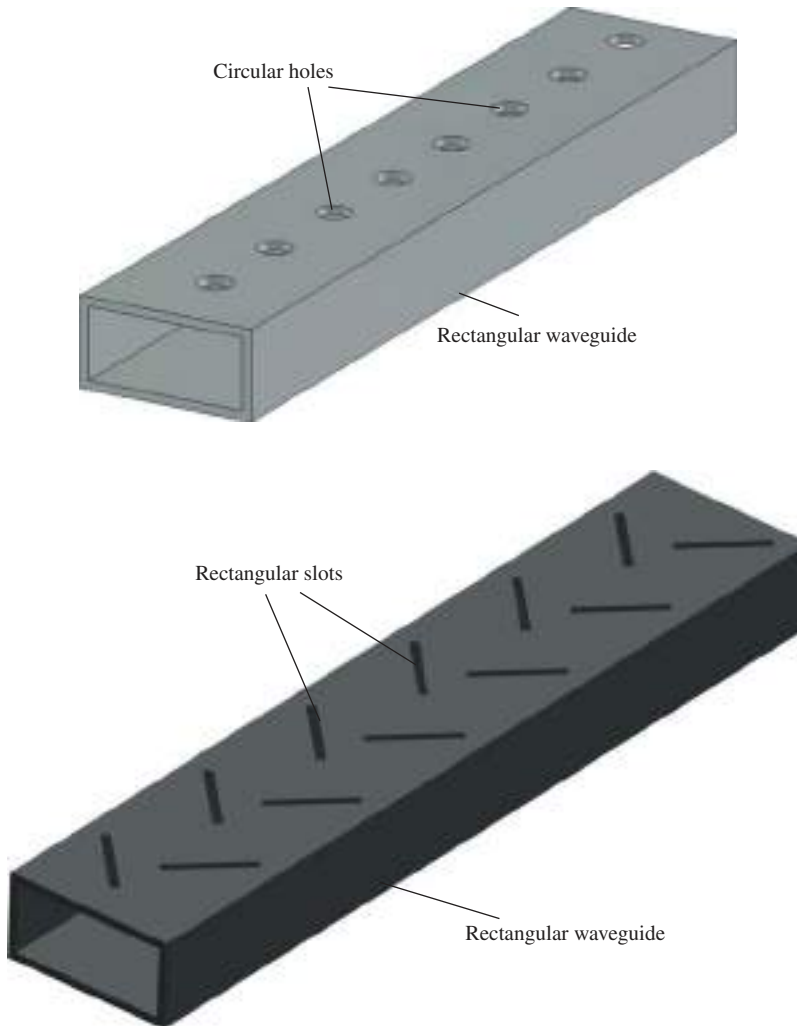
**Figure 5.28** Configuration of the  $4 \times 2$  ellipse-loaded slot antenna [20]

### 5.4.3 CP Waveguide Slot Antenna Arrays

The CP waveguide slot antenna array usually consists of circular holes, a pair of crossed slots cut or multiple slots cut on the broad wall of a rectangular waveguide, as presented in Figure 5.29. The circular polarization can be generated by cutting either the holes or the slots at the points where the transverse magnetic-field and the longitudinal magnetic field are equal in magnitude and in phase quadrature, thus inducing circularly polarized current and radiating CP waves. One interesting property of this kind of radiator is that it can radiate different polarized waves forwards and backwards. This directive property can be utilized to implement polarization detectors to separate different CP waves (RHCP and LHCP) from the incoming wave [22].

For a CP waveguide slot antenna array, the slots cut on the waveguide have to be reflectionless, which means that the width of the slots needs to be much smaller than their length. This type of antenna is inherently matched and when the slots are resonant, about 75% of the incident waves are radiated (with a VSWR of 1.12) [22]. When designing a slot array, separating each slot antenna by a distance of one guided-wavelength is required, to ensure that each antenna element can resonate with in-phase condition. Meanwhile, it is also important to make sure that this distance is no bigger than half of the free space wavelength to avoid the grating lobes. Therefore, it is necessary to load the waveguide to reduce the effective guided wavelength.

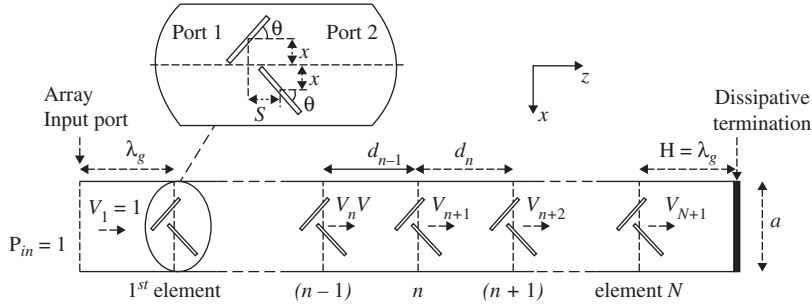
A design procedure for CP waveguide slot linear arrays is described in [23] and is briefly summarized next. Figure 5.30 shows the layout of the slot antenna element and the configuration of the entire array. This array is designed to have the minimum AR in the broadside



**Figure 5.29** Conventional configurations of the circularly polarized waveguide slot antenna array

direction. To design such a CP slot antenna array, firstly the slot antenna element needs to be able to radiate CP waves. The radiation properties of the slot antenna element can be optimized by adjusting the values of the slot's space  $s$  and the slot's length.

Since the waveguide is a two-port device, the amplitude and the phase of  $S_{21}$ , and the phase variation of the electric field-peak on the first and second slot ( $E^{1st}$  and  $E^{2nd}$ ) of the radiating elements needs to be studied by varying the slot length and slot space. When the slot element radiates CP waves, the phase variation of  $E^{1st}$  and  $E^{2nd}$  is the same. Because each antenna element radiates a fraction of the incident power, the complex amplitude of the  $TE_{10}$  mode travelling wave through the waveguide is tapered. Assume that  $V_n$  is the complex amplitude



**Figure 5.30** The geometry of the slot antenna element and the configuration of the waveguide slot antenna array [23]. Reproduced with permission of © 2006 IEEE

of the  $TE_{10}$  mode travelling wave that impinges on the  $n^{th}$  array element, it is defined that:

$$T_n = \frac{V_{n+1}}{V_n} \quad (5.4)$$

If  $P_n^{irr}$  is the normalized power radiated by the  $n^{th}$  element and  $P_n^{inc}$  is the power incident on the  $n^{th}$  element, then the following equation can be obtained:

$$P_n^{irr} = P_n^{inc}(1 - |T_n|^2) \quad (5.5)$$

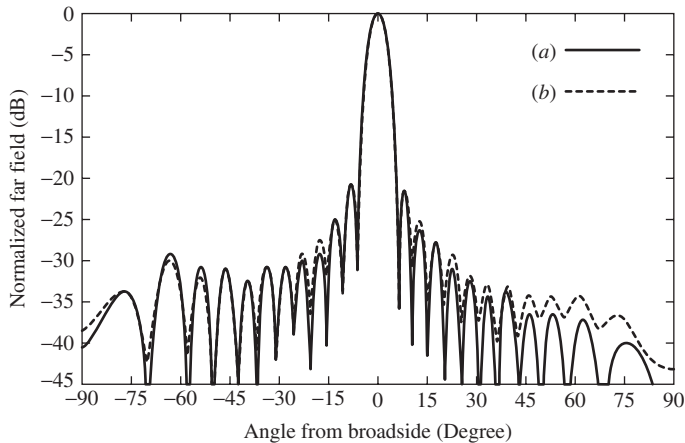
$$\sum_{n=1}^N P_n^{irr} = 1 \quad (5.6)$$

The following recursive equations can be also derived:

$$|T_1| = \sqrt{1 - P_1^{irr}} \quad (5.7)$$

$$|T_n| = \sqrt{1 - \frac{P_n^{irr}}{P_{n-1}^{inc} \times |T_{n-1}|^2}} \text{ for } N = 2 \dots N \quad (5.8)$$

Then, using these formulas, the  $|T_n|$  can be computed. Neglecting the mutual coupling between the slot antenna elements, the length of the slot, the space between the pair of slots can be determined. The distance between each slot antenna needs to be chosen in the way that each element has equal phase excitation. A 15-element slot antenna array with Taylor aperture with operation frequency at 7.5 GHz is presented in [23]. Teflon ( $\epsilon_r = 2.1$ ,  $\tan(\delta) = 0.001$ ), which has a high dielectric strength, is used to load the waveguide to reduce the effective guided wavelength. Figure 5.31 shows the measured radiation pattern of the CP waveguide slot linear array either with an absorber at the array termination or with a short circuit termination. In either case, this antenna array shows high directivity and low sidelobes. Measurement results also indicate that at the operating frequency, the value of AR in the broadside direction is less than 0.6 dB and below 1 dB within the whole 3-dB beamwidth.



**Figure 5.31** Radiation pattern of the CP waveguide slot linear array with: (a) an absorber at the array termination and (b) with a short circuit termination [23]. Reproduced with permission of © 2006 IEEE

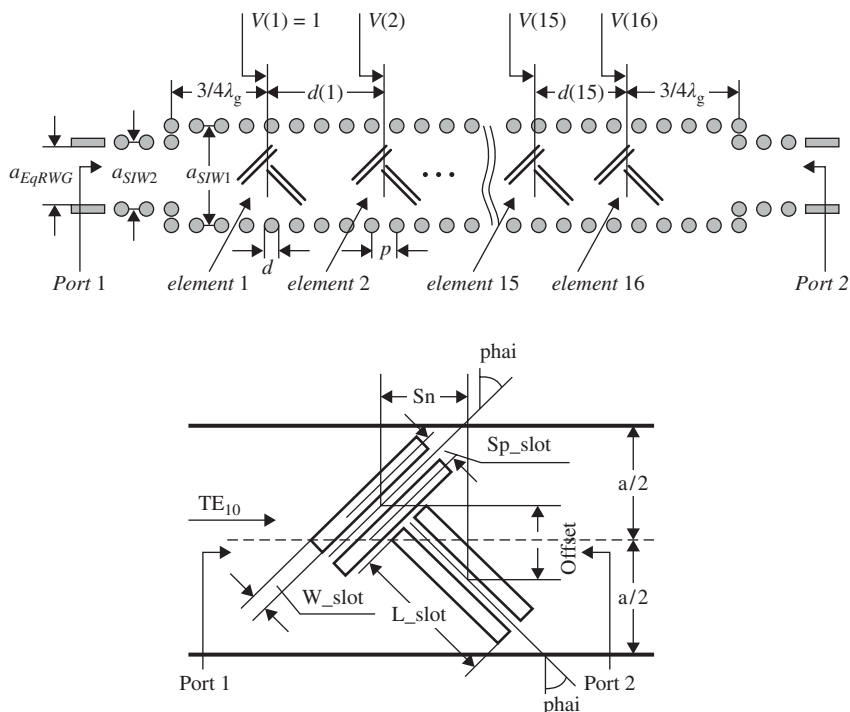
**Table 5.3** Radiation performance of the slot array within 100 MHz bandwidth [23]. Reproduced with permission of © 2006 IEEE

| Frequency (GHz) | Gain (dBi) | Efficiency (%) | SSL (dB) | AR (dB) |
|-----------------|------------|----------------|----------|---------|
| 7.45            | 18.04      | 89             | -19.3    | 1.4     |
| 7.50            | 18.06      | 92             | -20.9    | 0.6     |
| 7.55            | 18         | 92             | -21.4    | 1.3     |

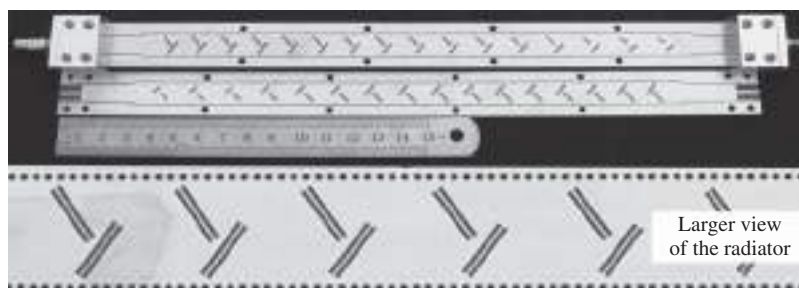
Table 5.3 summarizes the radiation performance of this waveguide slot array in a 100 MHz bandwidth around the operation frequency (7.5 GHz). As can be seen from this table, within the operation frequency band, the slot linear array exhibits high radiation efficiency and low axial ratio.

The Substrate Integrated Waveguide (SIW) technique has attracted much research interests as its advantages of compact size, low loss (especially for mm-wave circuits) and cost-effective solution for integrating active circuits, passive components and radiating elements on the same substrate. As a result, it is also possible to design a CP waveguide slot linear array on a substrate integrated waveguide. In this case, an array can be fabricated by using PCB technology and the slots with arbitrary shapes can be etched with high accuracy and low cost. One such design is presented in [24]. Figure 5.32 shows the layout of this SIW slot array and the geometry of the single radiation element.

Different from the slot antenna element presented in [23], four slots are used in one radiating element. This type of configuration is found to be useful when the SIW is used, as the high width-to-height ratio (WHR) can influence the return loss and AR of a conventional two-compound slots CP radiating element. The array is designed following the procedure proposed in [23] and the antenna prototype is fabricated on a 1.5-mm thick Rogers 5880 ( $\epsilon_r = 2.2$ ) substrate by using standard PCB process. This array has 16 elements in total and Figure 5.33 shows a photo of the fabricated prototype.



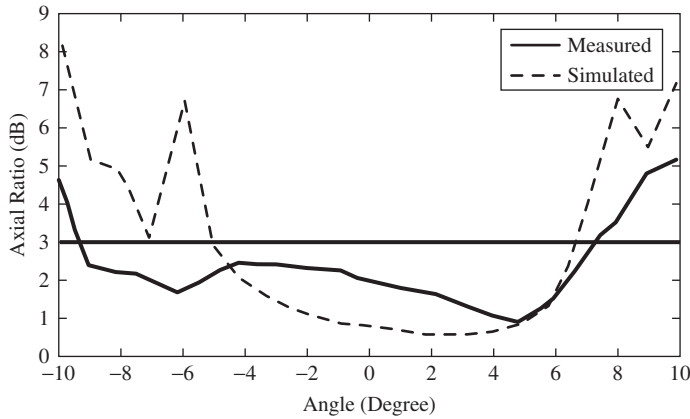
**Figure 5.32** Layout of this SIW slot array and the geometry of the single radiation element [24]. Reproduced with permission of © 2009 IEEE



**Figure 5.33** A photo of the fabricated SIW slot array [24]. Reproduced with permission of © 2009 IEEE

Measurement results show that this SIW slot array has RHCP radiation from 15.8–16.2 GHz and has a low sidelobe level of -23dB. At the broadside direction, the axial ratio is 1.95 dB, as given in Figure 5.34. The disagreement between the measured and simulated results is mainly due to the fabrication accuracy. The gain of it is about 18.9 dBi and it is estimated that 93.3% radiation efficiency at the designed frequency can be reached.

As can be seen from these two design examples, although the size of the overall array is not as compact as the printed microstrip antenna arrays due to the length of the waveguide, the waveguide slot antenna array exhibits high radiation efficiency and good AR at broadside.



**Figure 5.34** Measured AR compared with the simulated one at 16 GHz [24]. Reproduced with permission of © 2009 IEEE

#### 5.4.4 Radial Line Slot Antenna

Satellite communication systems, such as Direct Broadcast from a Satellite (DBS) usually needs to have powerful satellite transponders. This means that a high directivity CP antenna array is needed. To achieve a high gain, an antenna array with large aperture sizes is required. When the aperture size of the antenna array increases, the conductor and dielectric losses become notable. With the same aperture, the gain and efficiency of the slot array antenna is usually higher than the microstrip patch array. Therefore, the radial line slot antenna (RLSA) has gained wide application in the field of satellite communications.

RLSA belongs to a class of slotted waveguide antennas and it has a circular aperture. Figure 5.35 shows the structure of the conventional RLSA [25]. The slots are arrayed on the aperture in a way that they can be excited by the radial currents flowing over the aperture and produce circular polarization. In this design, a bended waveguide is used to excite the slot array antenna.

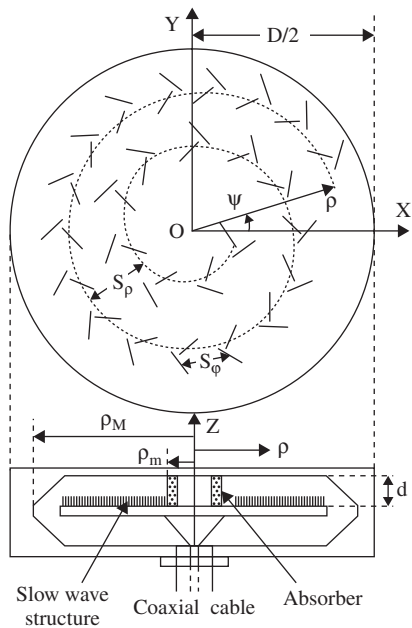
As can be seen from Figure 5.35, the slots are placed in sequence along a designed spiral. In order to suppress the grating lobes of the antenna array, the distance between each pair of slots ( $S_\rho$  and  $S_\phi$  as indicated in Figure 5.35) must be smaller than free space wavelength  $\lambda_0$ . The rules of designing a RLSA of radially inward travelling-wave mode is explained in [25]. There are two steps involved in determining the arrangement of the slots. The first step is to decide the position of the first slot pair. Figure 5.36 shows the coordinate of the slots.

The coordinates of the first pair of the ( $P1$  and  $P2$ ) can be calculated by using the following formulas with a giving initial value for  $\rho_1$ :

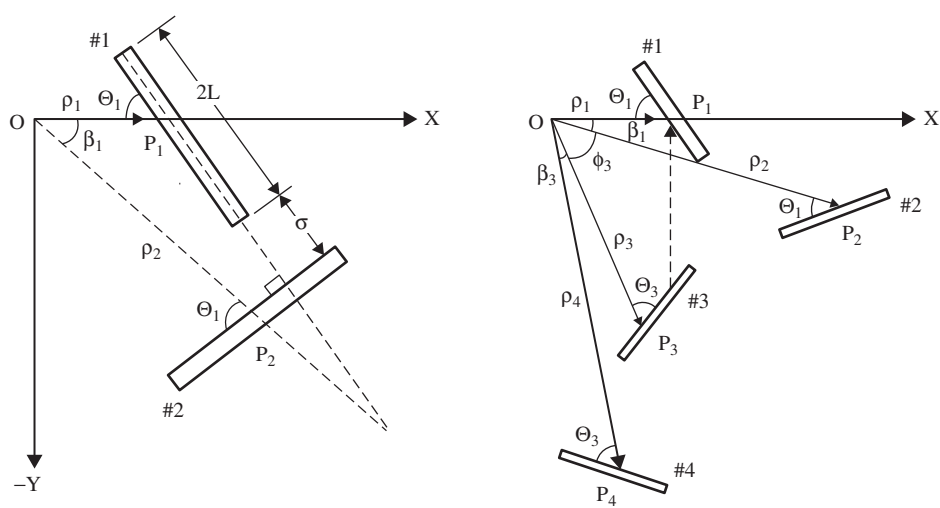
$$\arg(H_1^{(1)}(k\rho_2)) - \arg(H_1^{(1)}(k\rho_1)) = \frac{\pi}{2} \quad (5.9)$$

$$\rho_2 \sin \Theta_1 - \rho_1 \sin \Theta_1 = L + \sigma \quad (5.10)$$

$$\beta_1 = 2\Theta_1 - \frac{\pi}{2} \quad (5.11)$$



**Figure 5.35** The structure of the RLSA [25]. Reproduced with permission of © 1985 IEEE



**Figure 5.36** Coordinates and parameters of the slots of a RLSA [25]. Reproduced with permission of © 1985 IEEE

where  $H_1^{(1)}(z)$  is the Hanke function of the first kind of order,  $\rho_n$  is the distance from origin  $O$  to the centre  $P_n$  of the slot  $n$ ,  $\beta_{2n-1}$  is the angle  $P_{2n-1}OP_{2n}$ ,  $\sigma$  is the distance between two slots,  $k$  is the wavenumber in the waveguide and  $\arg(x)$  represents the phase of  $x$ .

After getting the position of the first slot pair, the next slot pair can be calculated by:

$$\phi_3 = \arg(H_1^{(1)}(k\rho_3)) - \arg(H_1^{(1)}(k\rho_1)) \quad (5.12)$$

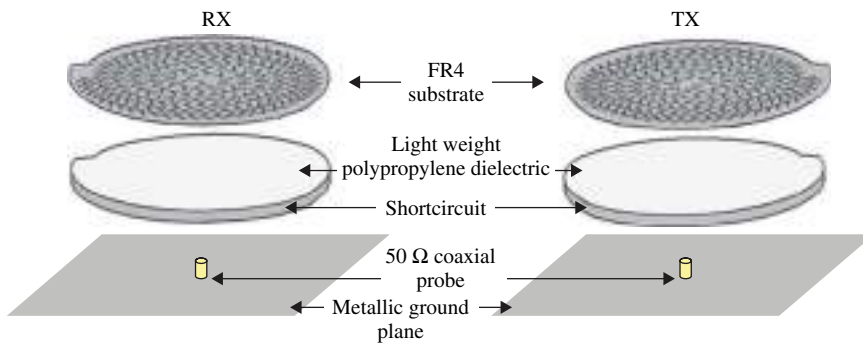
$$S_\phi^2 = \rho_1^2 + \rho_3^2 - 2\rho_1\rho_3 \cos\{\arg(H_1^{(1)}(k\rho_3)) - \arg(H_1^{(1)}(k\rho_1))\} \quad (5.13)$$

Then the successive slots can be determined by following the same procedures.

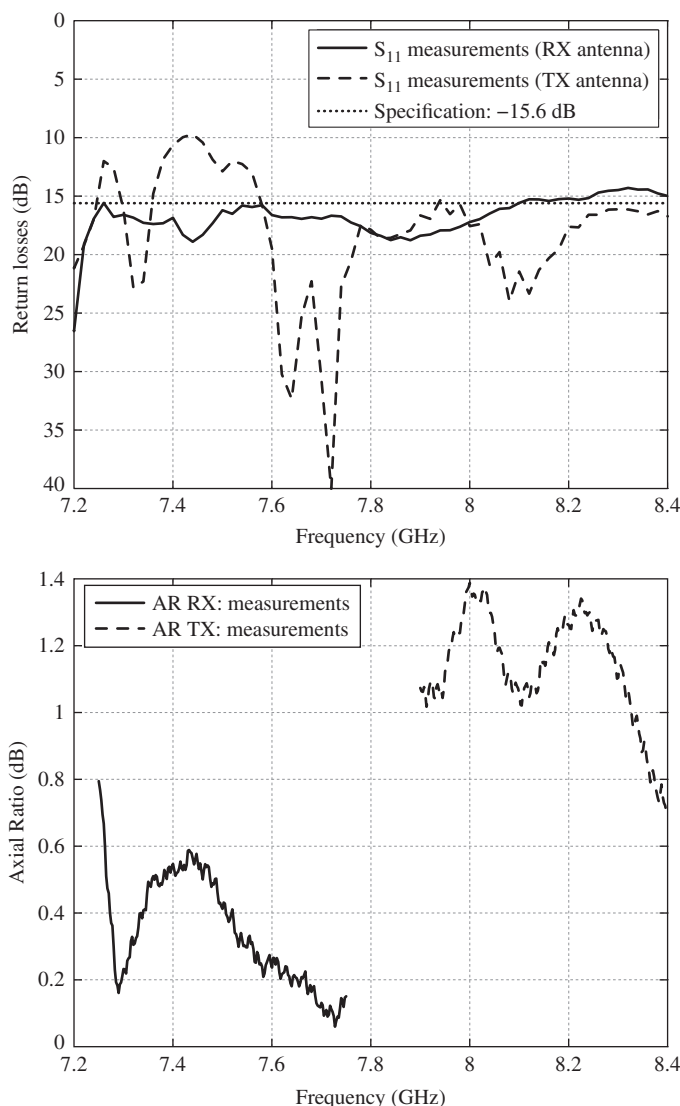
Conventional RLSAs are built with a waveguide structure, which is inherently heavy for large apertures. Significant weight reduction can be achieved if the RLSA is constructed by using PCBs. In [26], two lightweight portable planar RLSAs for satellite communications in X-band are reported. One of the RLSAs is designed to operate at 7.25–7.75 GHz as an Rx antenna with LHCP whilst the other one is designed to operate at 7.9–8.4 GHz as an Rx antenna with LHCP. Both of these two RLSAs have a two-layer dielectric structure. Figure 5.37 shows the exploded view of both antennas.

The slot elements are printed on a 0.127-mm thick FR4 substrate and a light weight polypropylene substrate of relative dielectric constant 1.8 and thickness of 10 mm is placed between the slot antenna and the metallic ground plane. The slot positions are calculated by using the equations provided in [25]. A 50  $\Omega$  SMA connector is used as the feed as illustrated in Figure 5.37. The SMA connector is centrally located at the phase centre of the slot array antenna in order to generate a radially outward travelling-wave mode. The slots are arranged spirally and tilted at 90° to one to other at their centre to generate CP radiation. RHCP can be obtained if the slot pairs are excited with a relative phase shift of  $-90^\circ$  and LHCP can be obtained if the relative phase shift changes to  $90^\circ$ . In this design, the distance between two consecutive slot pairs is chosen to be  $0.74\text{--}0.78\lambda_0$  to avoid the grating lobes and overlapping of the adjacent slots.

Figure 5.38 shows the measured return loss and axial ratio of the Rx and Tx RLSAs. It can be seen that within the required operation frequency bands, the measured return loss is always better than 15.6 dB and the AR is smaller than 1.4 dB. The co-polarization level is 25 dB higher than the cross-polarization in the broadside direction and the gain is higher



**Figure 5.37** Details of the light weight RLSA [26]. Reproduced with permission of © 2011 IEEE



**Figure 5.38** Measured return loss and axial ratio of the light weight Rx and Tx RLSAs [26]. Reproduced with permission of © 2011 IEEE

than 25 dBi for both Rx and Tx array. The measurement results also show that this slot array antenna has a radiation efficiency better than 70%.

To obtain a CP radial line slot antennas with either pencil beam or shaped beam patterns, an automatic design rule is explained in [27]. It takes into consideration of the side lobe levels and the influence of the spurious higher modes in the parallel plate waveguide (PPW) when designing the layout of the RLSA slot. After that, an optimization scheme is introduced to optimize the radiation pattern of the RLSA. Readers can refer to [27] for more details of the array design procedure and how to define the fitness function for the array optimization.

## 5.5 CP Printed Reflectarrays

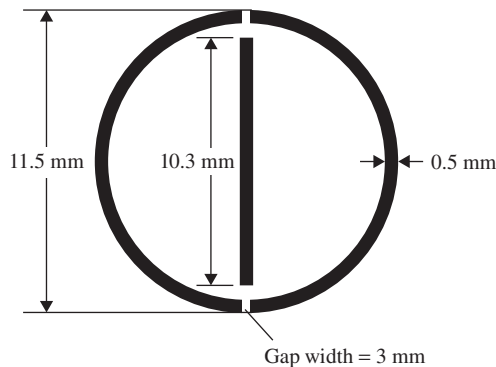
A reflectarray is an antenna array that consists of many radiating elements, each of which has a preadjusted phase, to form a focused beam when it is illuminated by a feed. At the central frequency, the reflectarray can operate similar to a parabolic antenna. The printed reflectarray has the advantages of easy fabrication, light weight and offers the possibility of beam steering as phased arrays. Compared to the conventional microstrip antenna array, the printed reflectarray eliminates the complexity and losses of the feed network. However, due to the limitation of the microstrip antennas, the bandwidth of the printed reflectarray is normally narrow.

### 5.5.1 CP Feed

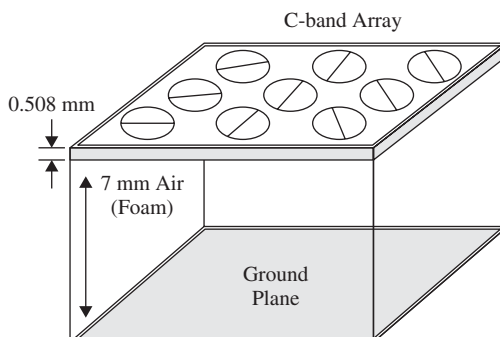
A straightforward approach to design a CP printed reflectarray is to introduce printed CP antennas as the unit cells of the reflectarray and feed the reflectarray using a CP feed. Therefore, to design such a reflectarray, it is important to have an antenna unit cell that can radiate CP waves and provide the desired phase compensation to form a focused beam as described in [28].

One circularly polarized reflectarray with microstrip ring elements having variable rotation angles is presented in [29]. The microstrip ring element is used in this design because of its broadband performance and easy adaptation to the multilayer structure, as it has little blockage to other layers. Figure 5.39 presents the geometry of the single antenna element and Figure 5.40 shows the structure of the reflectarray. As can be seen from Figure 5.39, there are two gaps in the microstrip ring. These gaps are created for the purpose of keeping the same kind of CP polarization between the incident and reflected waves. The microstrip bar in the middle of the ring can help to give more suppression between the RHCP and LHCP: the simulation results provided in [29] show that when the RHCP plane wave is incident, the reflected LHCP can be suppressed by more than 20 dB compared to the reflected RHCP.

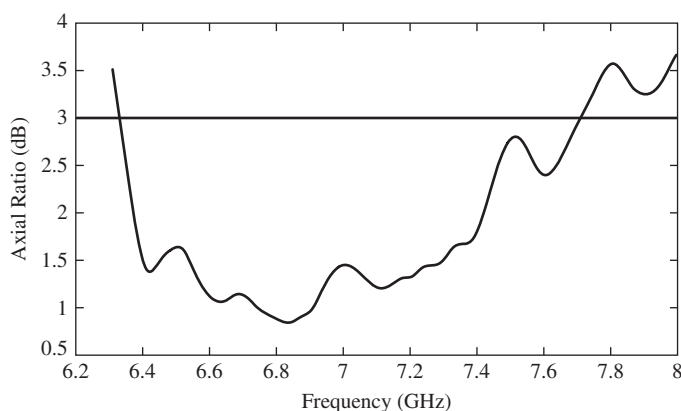
By changing the rotation angle of the ring element, the relative phase of the reflected wave can be adjusted from  $0^\circ$  up to  $330^\circ$  when it is rotated counter-clockwise from  $0^\circ$  to  $170^\circ$ . This reflectarray is designed to operate at 7.1 GHz and the ring antenna elements are



**Figure 5.39** Layout of the single microstrip ring antenna element [29]. Reproduced with permission of © 2004 IEEE



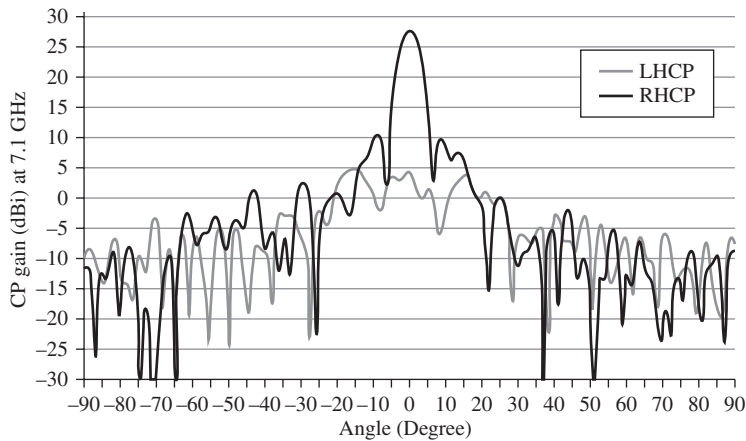
**Figure 5.40** Structure of the reflectarray using microstrip ring with different rotation angle [29]. Reproduced with permission of © 2004 IEEE



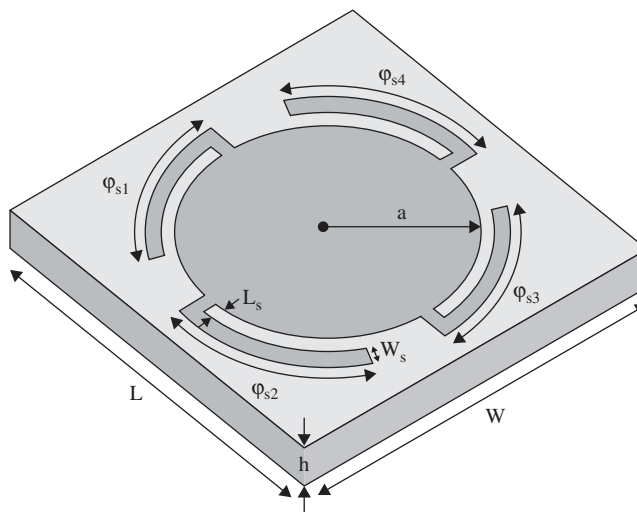
**Figure 5.41** Measured AR against frequency of the CP reflectarray with rotated ring element [29]. Reproduced with permission of © 2004 IEEE

separated by  $0.5\lambda_{7.1\text{ GHz}}$ . The reflectarray is printed on a 0.508-mm thick Roger Duroid 5780 substrate ( $\epsilon_r = 2.33$ ) and this layer is placed 7 mm above the ground plane. The parameters of the antenna element are given in Figure 5.39. The overall reflectarray has a diameter of 0.5 m. The measurement result (Figure 5.41) shows that the reflectarray has a 3-dB axial ratio of 18% with central frequency of 7.1 GHz. Figure 5.42 shows the measured radiation pattern of this CP reflectarray. It has a peak gain of 27.6 dBi while the cross polarization is 23 dB less than the co-polarization at the beam peak. The theoretical efficiency of this reflectarray is estimated to be 59.6%

The antenna unit cell proposed in [29] can provide up to  $330^\circ$  phase compensation when it is rotated. When comes to the design of large size reflectarray, the difference in spatial phase delays between the feed and elements in the array can limit the bandwidth (e.g. 1-dB gain bandwidth) of the reflectarray. Therefore, it is always desirable to have one antenna element that is capable of providing a large-range linear phase response [30]. To address this issue, one microstrip circular patch with four attached variable-length phase delay lines is proposed in [31]. The structure of this antenna element is shown in Figure 5.43.

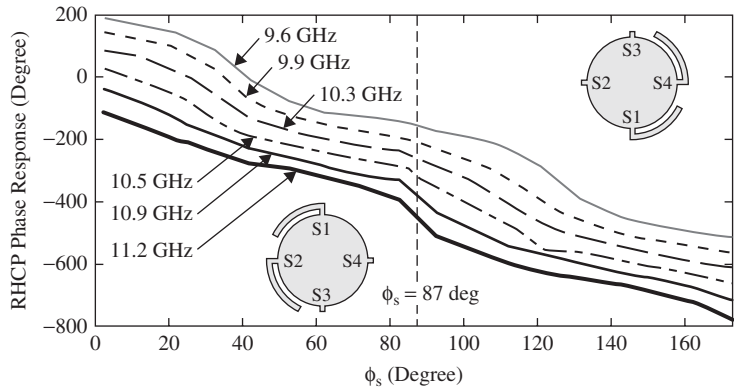


**Figure 5.42** Measured radiation pattern of the CP reflectarray with rotated ring element [29]. Reproduced with permission of © 2004 IEEE



**Figure 5.43** The structure of the microstrip circular patch with four attached variable-length phase delay lines for the CP reflectarray design [31]. Reproduced with permission of © 2012 IEEE

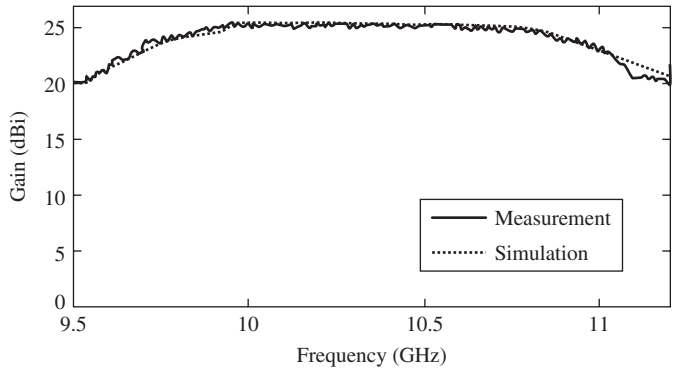
This antenna unit cell is printed on a Rogers 4003 substrate ( $\epsilon_r = 3.55$ ) with thickness of 0.813 mm. The radius of the circular patch is chosen to be 4.2 mm in order to resonate at 10.3 GHz. Simulation results show that by varying the length of the corresponding attached phase delay line ( $\varphi_{s(i)}$ ), it is possible to reach a continuous phase delay of more than  $650^\circ$ . Figure 5.44 shows the phase response of the CP antenna element excited at different frequencies with a RHCP incident plane wave. It can be seen that at different frequencies in the range of 9.6–11.2 GHz, the antenna unit cell has a parallel phase response, which implies that it can be useful for a broadband reflectarray design. This centre-fed reflectarray



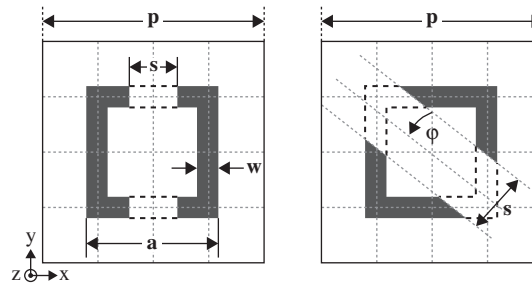
**Figure 5.44** Phase response of the CP antenna element excited at different frequencies [31]. Reproduced with permission of © 2012 IEEE

antenna has a size of  $23 \times 23$  cm. The measurement results show that the resulted CP reflectarray with a central feed has a 3-dB AR bandwidth of 14.6% with central frequency of 10.3 GHz. Meanwhile, it exhibits 10% 1-dB gain bandwidth and 15.5% 3-dB gain bandwidth with maximum gain of 25.3 dBi at 10.3 GHz, as shown in Figure 5.45. The measured radiation efficiency of this reflectarray is 45% and it can be improved by using low loss dielectric material as the substrates.

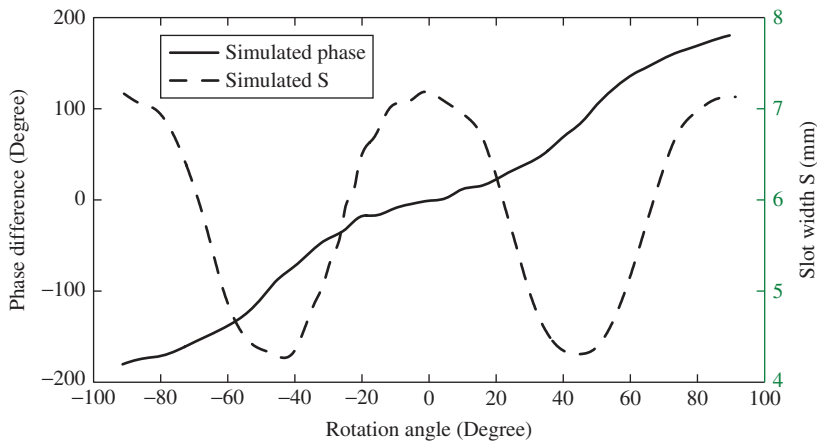
Instead of rotating the antenna elements, phase delay can also be created by changing the location of the slot on the microstrip ring. In this way, the antenna element does not need to be physically rotated. It can be advantageous to use such technique to design a multilayer reflectarray because the lattice of the array does not need to be changed. This also gives more flexibility in designing a multi-band structure with different elements inserted into the same substrate layer. In [32], an offset-fed X-band reflectarray antenna using this modified element rotation technique is presented. The rotation of the slots is demonstrated in Figure 5.46.



**Figure 5.45** Measured RHCP gain against frequency [31]. Reproduced with permission of © 2012 IEEE



**Figure 5.46** Demonstration of the modified element rotation technique [32]. Reproduced with permission of © 2012 IEEE



**Figure 5.47** The reflected phase against element rotation angle with varying slot width at 8.4 GHz [32]. Reproduced with permission of © 2012 IEEE

The microstrip square ring is printed on a 1.57 mm thick grounded substrate with relative dielectric constant  $\epsilon_r = 2.33$ . Figure 5.47 shows the reflected phase of the antenna elements and its corresponding slot width at a different rotation angle. The slot width needs to be adjusted at a different rotation angle in order to optimize the magnitude of the co-polarization component at the centre frequency. In this design, the central frequency is set to be 8.4 GHz.

In order to reduce the blockage from the feeding horn, an offset feeding is used. In total, 556 elements are used in this reflectarray. The measurement results indicate that this CP reflectarray can operate at 8.4 GHz with a gain of 30.4 dBi. The aperture efficiency is calculated to be 53%. Moreover, both the 3-dB axial ratio bandwidth and the 3-dB gain bandwidth are around 6%.

### 5.5.2 LP Feed

Another approach to designing a CP reflectarray is to use the linearly polarized feed with its electric field vector aligned in the diagonal direction ( $45^\circ$ ) of the array lattice. Then, the vector of the incident wave can be decomposed into one vertical and one horizontal

component. In order to have a CP reflected wave, the antenna element needs to be designed in the way that these two perpendicular components will be reflected back with a  $90^\circ$  phase difference and equal amplitude, which results in a CP wave. The phase condition can be summarized as:

$$Ph_x = \phi \quad (5.14)$$

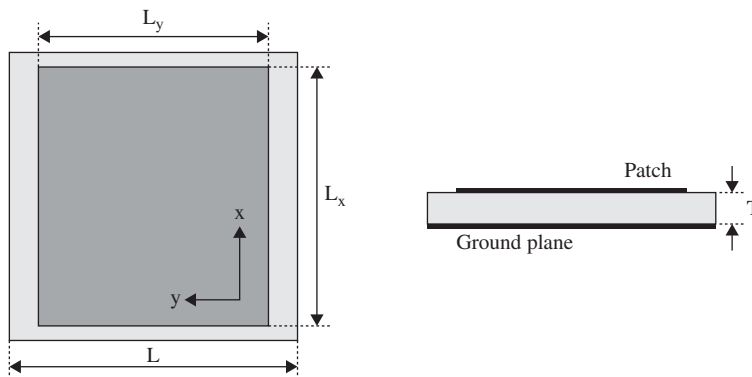
$$Ph_y = \phi \pm \frac{\pi}{2} \quad (5.15)$$

where  $Ph_x$  and  $Ph_y$  represents the phase of the reflected wave in the X- and Y-direction.

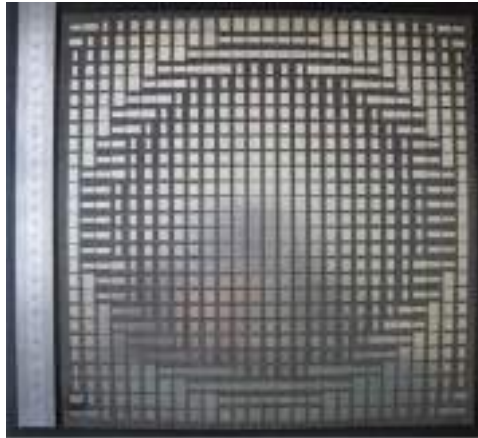
To demonstrate this technique, one good design example is presented in [33], where a sub-wavelength element for broadband CP reflectarrays is proposed. The geometry of the antenna element is shown in Figure 5.48. The rectangular patch is printed on a 2-mm thick substrate with relative dielectric constant of 2.65. The distance between each antenna element is kept as  $\lambda_{10 \text{ GHz}}/3$  instead of the commonly used  $\lambda_{10 \text{ GHz}}/2$ , which has been proved that through using this configuration, better radiation performance can be achieved in terms of the gain and AR bandwidth [33].

The polarization of the LP feed is set to parallel with the diagonal line of the square shape substrate. The lengths of the rectangular patch,  $L_x$  and  $L_y$ , are decided in the way that there is always a phase difference of  $90^\circ$  between the two reflected wave components (in horizontal and vertical direction). A photo of the fabricated reflectarray is shown in Figure 5.49 and it can be observed that the antenna unit cells have a rectangular shape in a square lattice in order to induce the CP reflected waves. This reflectarray consists of  $27 \times 27$  elements. Measurement results (Figure 5.50) show that it has 1-dB gain bandwidth of 17% and a 3-dB axial ratio bandwidth of 11%. Its peak gain is 25.8 dBi and the radiation efficiency is about 39% at 9.8 GHz.

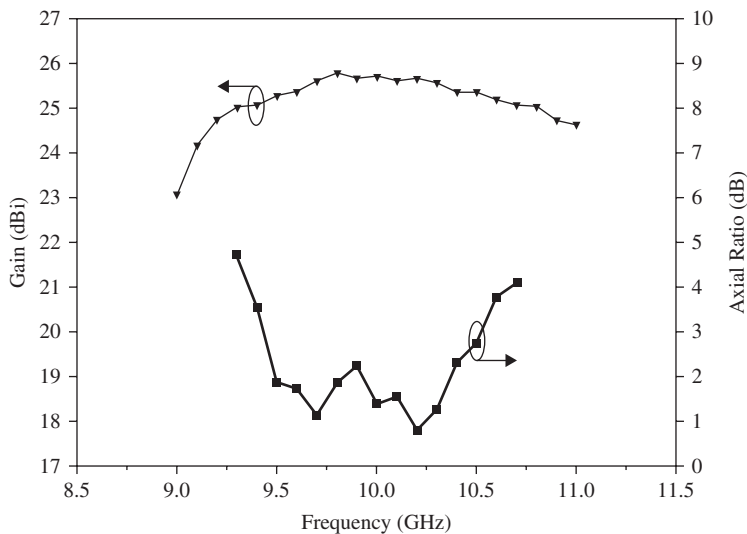
It is known that a microstrip patch antenna with stacked structure can improve its impedance bandwidth. This also stands when such a technique is applied to the design of the CP reflectarray. With a LP feed, one dual-layer broadband CP reflectarray consisting of T-shaped elements [34] and one CP reflectarray consisting of rectangular patches [35] have been reported. The structures of these two designs are illustrated in Figure 5.51 and



**Figure 5.48** Top and side view of the antenna element with rectangular patch [33]. Reproduced with permission of © 2010 IEEE



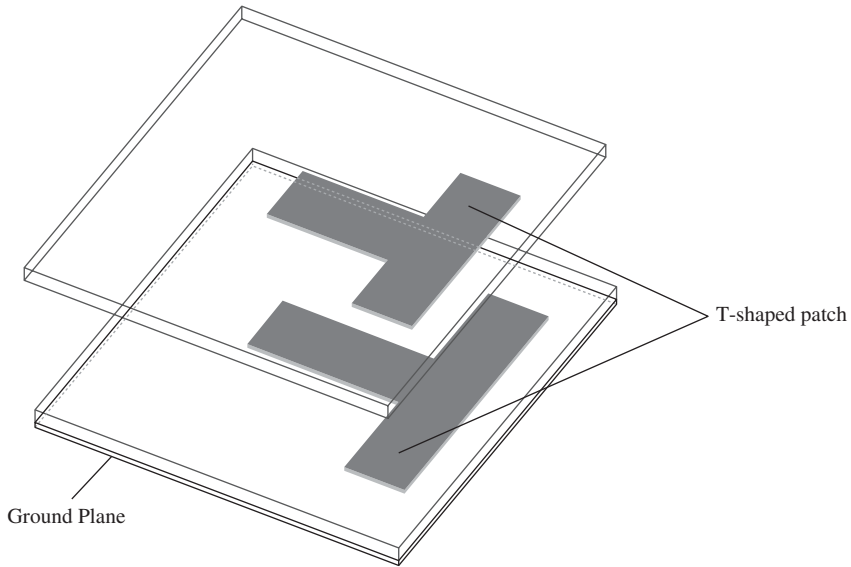
**Figure 5.49** A photo of the fabricated CP reflectarray using a LP feed [33]. Reproduced with permission of © 2010 IEEE



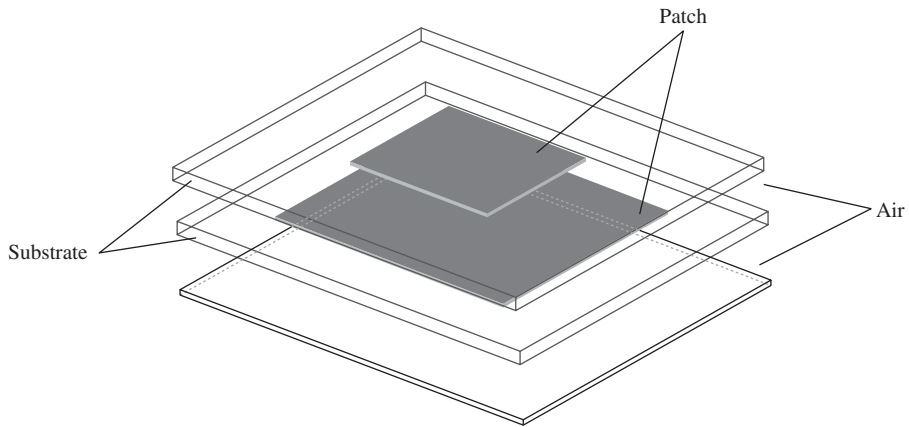
**Figure 5.50** Measured 1-dB gain bandwidth and a 3-dB axial ratio bandwidth of the CP reflectarray with sub-wavelength element [33]. Reproduced with permission of © 2010 IEEE

Figure 5.52. To employ a multilayer structure to design a broadband reflectarray, it is required that the patch on the top layer is smaller compared to the one in the bottom layer, for the purpose of reducing the blockage to the antenna at the lower layer. Besides the  $90^\circ$  phase difference requirement for the reflected wave in X- and Y-direction, the dimensions of the patches on both layers need to keep the same ratio. In the case of the T-shaped antenna element proposed in [34], the following condition needs to be satisfied:

$$\frac{L_{v1}}{L_{v2}} = \frac{L_{h1}}{L_{h2}} = \text{constant} \quad (5.16)$$



**Figure 5.51** Exploded view of the dual-layer reflectarray element using two T-shaped patches [34]



**Figure 5.52** Exploded view of the multilayer reflectarray element using two rectangular patch [35]

where  $L_{v1}$  and  $L_{h1}$  is the length of the vertical and horizontal section of the T-shaped element on the top layer,  $L_{v2}$  and  $L_{h2}$  is the length of the vertical and horizontal section of the T-shaped element on the bottom layer, respectively. While in the case of the multilayer reflectarray element using two rectangular patches as the one proposed in [35], it will be:

$$\frac{b_x}{a_x} = \frac{b_y}{a_y} = \text{constant} \quad (5.17)$$

where  $a_x$  and  $a_y$  is the length and width of the patch on the bottom layer,  $b_x$  and  $b_y$  is the length and width of the patch on the top layer, respectively.

**Table 5.4** The radiation performance of the dual-layer reflectarray proposed by [34] and [35]

|                             | Measurement result<br>of design[34] | Measurement result<br>of design[35] |
|-----------------------------|-------------------------------------|-------------------------------------|
| No. of Elements             | 81                                  | 37                                  |
| 3-dB AR bandwidth           | 28% ( $f_0 = 10$ GHz)               | –                                   |
| Minimal AR                  | 0.62 dB at $f = 9.5$ GHz            | –                                   |
| 1 dB Gain bandwidth         | 20% ( $f_0 = 10$ GHz)               | –                                   |
| Maximum Gain or Directivity | 19.4 dBi at $f = 9.5$ GHz (Gain)    | 17.5 dBi (Directivity)              |
| Antenna efficiency          | 44%                                 | 33.6%                               |

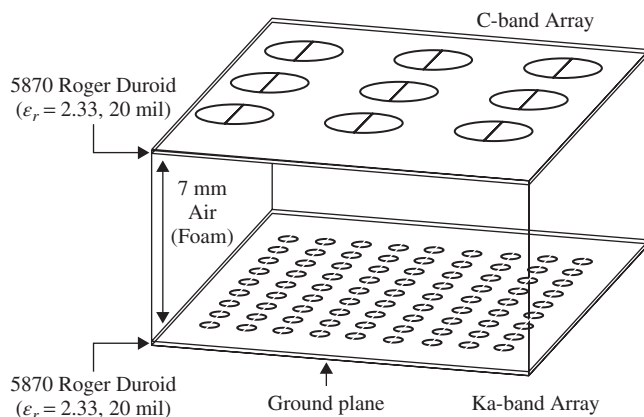
Due to the use of multilayer structure, both designs exhibit a broadband radiation performance. The radiation performance of these two dual-layer CP reflectarrays is summarized in Table 5.4. It is observed that the reflectarray with T-shaped element has a 3-dB AR bandwidth of 28% and 1-dB gain bandwidth of 20% with central frequency of 10 GHz, which is much larger than the conventional single layer reflectarray. Meanwhile, it also exhibits higher radiation efficiency compared to the reflectarray using a rectangular patch as the radiating element.

### 5.5.3 Multi-Band CP Printed Reflectarray

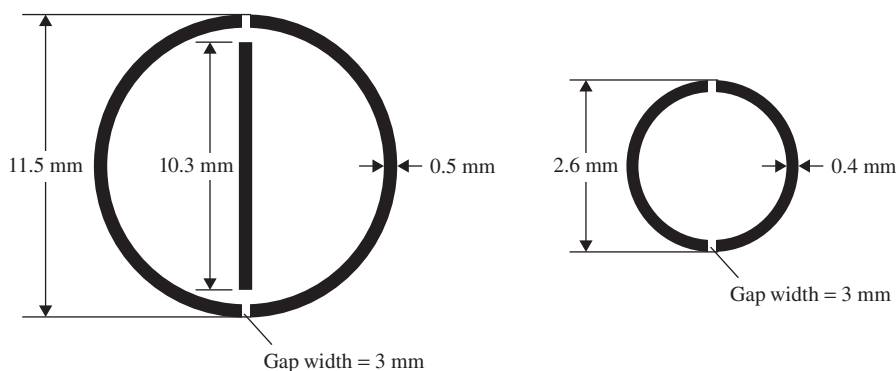
Similar to the design of the multi-band patch antenna, the multi-band CP reflectarray can be realized by employing a multilayer structure. As presented in the last subsection, several single band CP reflectarray elements, like the microstrip ring patch presented in [29], has the potential to be used in a multi-layer structure as it can provide little blockage to the other layers. Based on these studies, a C/Ka dual frequency dual-layer CP reflectarray antenna is proposed in [36]. Figure 5.53 shows the configuration of the CP dual-layer reflectarray using the microstrip rings as the radiating elements and Figure 5.54 presents the geometry of the antenna element for both layers.

It has been demonstrated in [29] that the microstrip ring patch can be used in CP reflectarray design and the phase compensation of the radiating elements can be controlled by rotating the microstrip rings. In this design, two different sizes of the microstrip ring are chosen to operate at the desired frequencies and a dual band CP feed is required to feed the reflectarray. Both layers of the reflectarray are printed on the 0.5-mm thick Roger Duroid 5870 ( $\epsilon_r = 2.33$ ) and are separated by a distance of 7-mm. There are 437 C-band ring elements printed on the top layer, and on the bottom layer there are 9236 Ka-band ring elements. The measurement results show that compared to the single layer design, there is little blockage effect caused by the use of this dual-layer configuration. Figure 5.55 shows the measured radiation patterns of this dual band CP reflectarray at 7.3 and 31.75 GHz. As can be seen from the measurement results, this reflectarray is RHCP with low cross polarization level at both bands. The measured CP gain at 7.3 GHz (C-band) is 28 dBi with a radiation efficiency of 46% while at 31.75 GHz (Ka band), the measured CP gain is 40 dBi with 38% radiation efficiency.

The multi-band CP reflectarray can also be obtained by placing different antenna elements in the same array lattice and thus results a single layer structure. The single layer structure



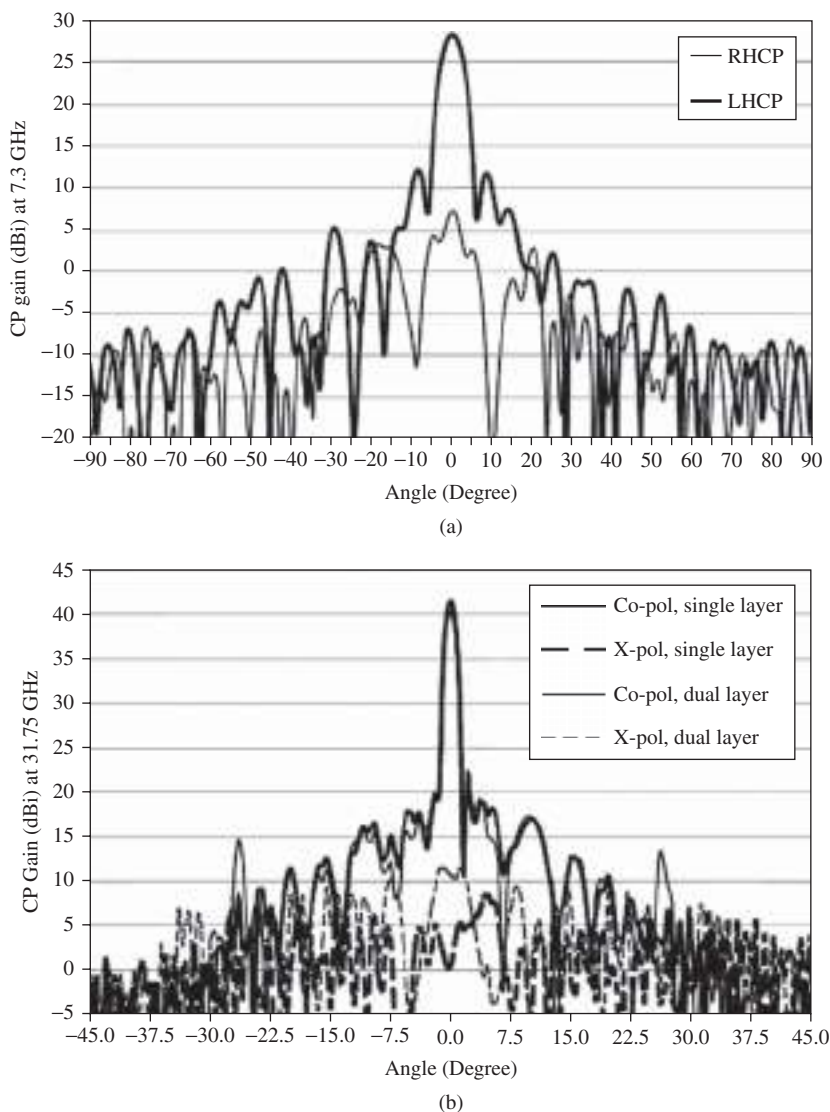
**Figure 5.53** Topology of the CP dual-layer reflectarray [36]. Reproduced with permission of © 2004 IEEE



**Figure 5.54** The configuration of the ring antenna element for the top layer (left) and for the bottom layer (right) [36]. Reproduced with permission of © 2004 IEEE

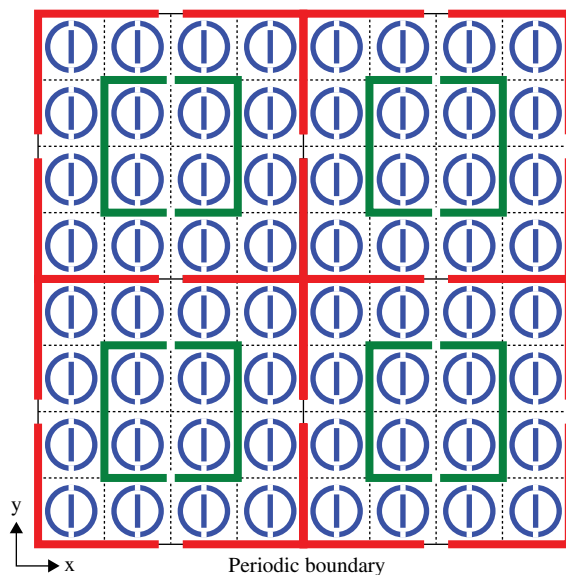
has the advantages of low cost and simply fabrication. To design a single layer multi-band reflectarray with different antenna elements, it is required that each antenna element needs to be chosen in such a way that it has little influence on other elements. One single layer tri-band reflectarray design is presented in [37]. Figure 5.56 shows the top view of this single layer tri-band reflectarray.

As can be seen from Figure 5.56, this tri-band reflectarray consists of three different types of elements: split circular rings, split square loops and cross-dipole elements. All of these three types of elements are printed on the top layer of a grounded RT/Duroid 5870 substrate with a thickness of 62 mm. The split circular ring is responsible for the radiation at the Ka band (32 GHz) while the split square loop is for radiation at the X band (8.4 GHz). The reflection phase of these two types of radiation element can be adjusted by moving the position of slots along the perimeter of the circular ring or the square loop. This *angular rotation technique* has already been presented in previous sections. The cross-dipole is designed to operate at C band (7.1 GHz) and its reflection phase can be adjusted by varying its length.

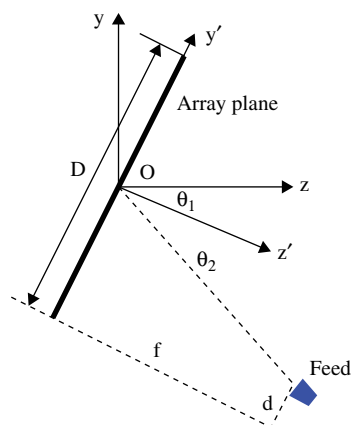


**Figure 5.55** Measured radiation patterns of this dual band CP reflectarray at (a) 7.3 GHz and (b) 31.75 GHz [36]. Reproduced with permission of © 2004 IEEE

Due to the special geometries of these three types of radiation elements, they can be arranged into the same layer without causing any blockage to each other. A CP horn antenna is employed as the feed for the reflectarray and to reduce the feed blockage, an offset feed structure is used. Figure 5.57 shows the configuration of the offset-fed reflectarray design. The antenna elements are aligned on a circular aperture with the diameter  $D = 500$  mm. To achieve the maximum aperture efficiency, the ratio of the focal distance  $f$  (where the feed located) against the diameter of the reflectarray is optimized by calculating both the



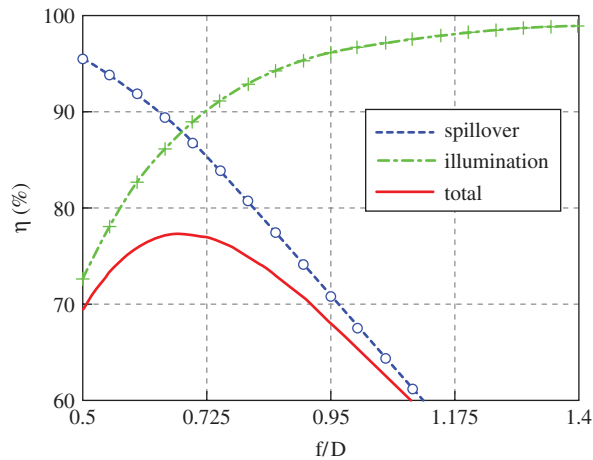
**Figure 5.56** The top view of the single layer tri-band reflectarray [37]. Reproduced with permission of © 2008 URSI



**Figure 5.57** Configuration of the offset design for the CP reflectarray [37]. Reproduced with permission of © 2008 URSI

spillover and illumination efficiency. The calculated overall efficiency against  $f/D$  is presented in Figure 5.58. With the assumption that the feed antenna has a radiation pattern of  $\cos^6(x)$ , the maximum total radiation efficiency can be reached when  $f/D$  equals 0.68.

This CP reflectarray is measured at X band. The measurement results show that it has a gain of 30.4 dBi with 53.7% aperture efficiency. Within the main lobe, the cross polarization is always 20 dB lower than the co-polarization.



**Figure 5.58** Calculated overall efficiency of reflectarray when  $f/D$  varies [37]. Reproduced with permission of © 2008 URSI

## 5.6 Integrated CP Array and Active CP Array

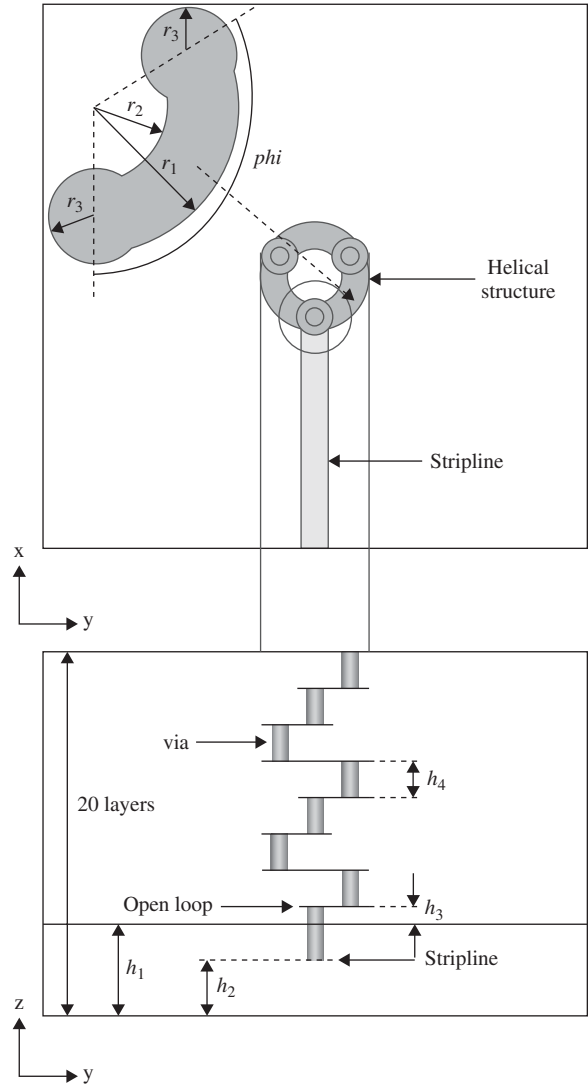
From the aspect of system design, it is important to have an antenna or antenna array as compact as possible and can be easily integrated into the overall system. This requires designing a compact size antenna or antenna array in a way that it is robust and less influenced by the other components. Therefore, the antenna integration becomes an important topic to investigate. In this section, two highly integrated CP antenna arrays for 60 GHz applications will be presented here as design examples.

The Active Integrated Antenna (AIA) consists of active devices and a antenna radiating element. It has the capability of signal processing, such as mixing and amplification [38]. The advantages of an integrated antenna are that there is only one impedance matching network between the active device (e.g. power amplifier) and the antenna instead of having one active device match to a 50- $\Omega$  transmission line, and then matches the transmission line to the input of the antenna. Thus, higher overall radiation efficiency and more compact size can be achieved. This is more advantageous in the millimetre-wave system, where the high loss from the microstrip line is always a concern.

The active integrated antenna is classified by the functions of active devices it integrates with. Therefore, active integrated antennas can be classified into three categories [39]:

- Oscillator type, which integrates active device to generate the RF signal.
- Amplifier type, which integrates amplifier(s) to amplify the RF signal.
- Frequency conversion type, which integrates active device to convert the signal to different frequencies.

In this section, the second type of the active integrated antenna is discussed, which has attracted much research interest in the field of CP antenna array design. For this type of active antenna array, it is important to have a high efficiency power amplifier because even

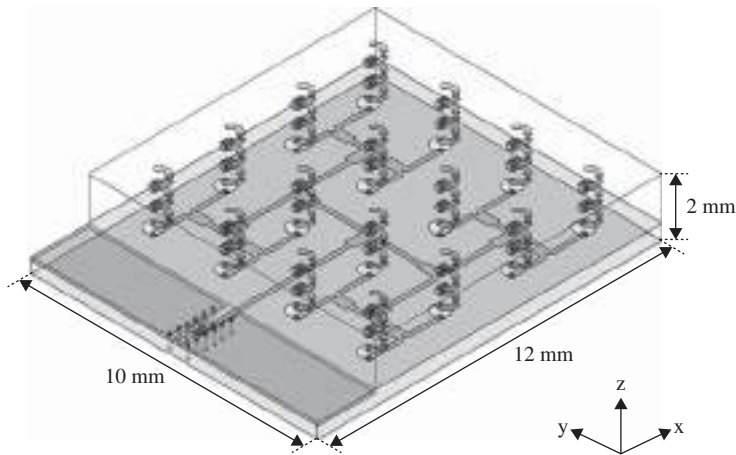


**Figure 5.59** Top and side view of the LTCC CP helical array element [41]. Reproduced with permission of © 2012 IEEE

a few percent of improvement in PAE can be significant if it can be designed without the major degradation in linearity [38,40].

### 5.6.1 Integrated CP Array

Low temperature co-fired ceramic (LTCC) is a key technology for highly integrated antenna design especially for antennas and arrays for millimetre-wave applications. LTCC



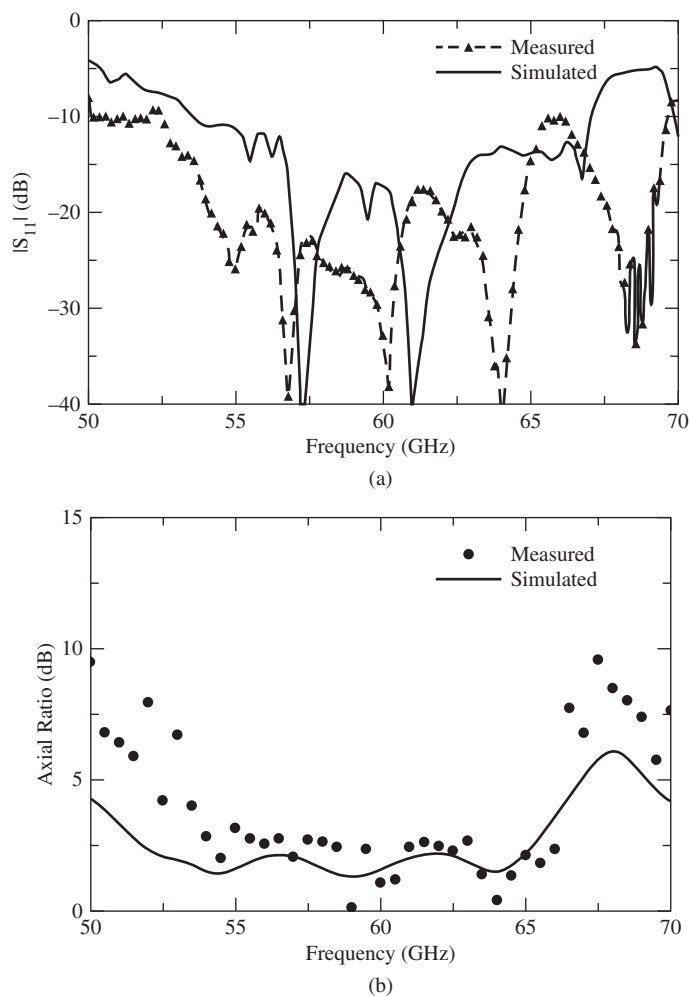
**Figure 5.60** The layout of the  $4 \times 4$  helical antenna array [41]. Reproduced with permission of © 2012 IEEE

technology is usually preferred in the multi-layer fabrication as has advantages of easily introducing blind, buried and through-plated holes anywhere in the substrate compared to conventional PCB multi-layer fabrication. With the help of LTCC technology, a multi-layer structure can be fabricated with good accuracy compared to conventional PCB technology.

Wireless short range communication at 60 GHz has received much attention as it can provide a wideband connection within a short range and support frequency reuse due to the high propagation attenuation of the millimetre wave. Using the LTCC technology to design highly integrated CP antenna arrays for 60 GHz band application has been carried out and reported in [41,42]. One 60 GHz LTCC integrated CP helical antenna array is presented in [41]. This antenna array consists of  $4 \times 4$  elements, each of which is composed of eight  $1/3$ -turn open loops printed in various layers. These layers are connected with via holes to form an axial-mode helical structure to generate travelling wave radiation. Figure 5.59 shows the top and side view of the antenna element for the CP array. The antenna is fabricated using the Ferro A6-M LTCC type system with a fired thickness of 0.1 mm for each layer. The relative dielectric constant of the material is 5.9 and loss tangent is 0.002. The helical antenna is fed by a stripline to suppress the radiation from the feed lines.

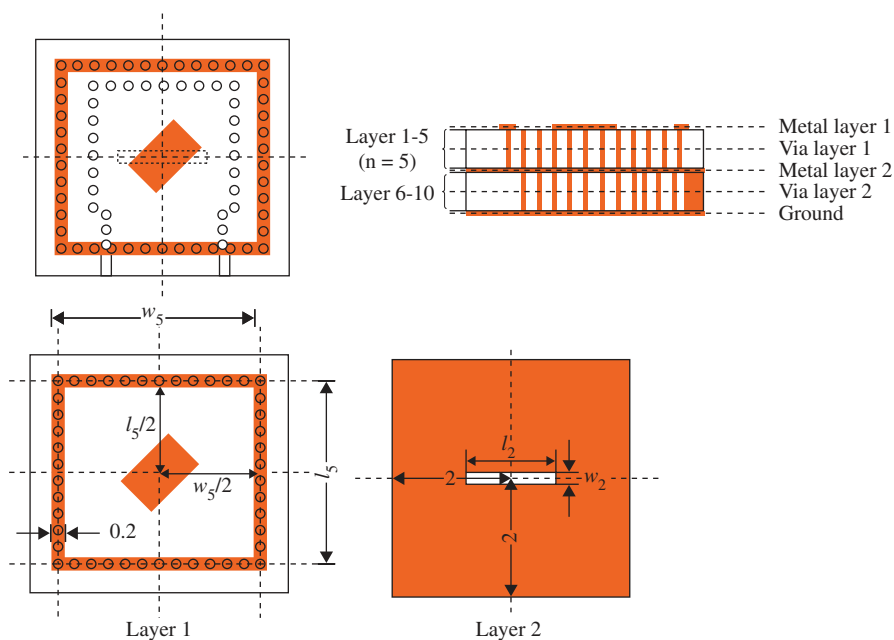
Figure 5.60 shows the overall structure of the  $4 \times 4$  helical antenna array. T-junction feed network with a grounded coplanar waveguide (GCPW) to stripline transition is employed to feed the antenna elements. The distance between each unit cell is 2.5 mm, which is half of the free space wavelength at 60 GHz. The overall size of this CP antenna array is  $12 \times 10 \times 2 \text{ mm}^3$ .

Figure 5.61 presents the measured and simulated return loss as well as the axial ratio of the CP spiral antenna array. The measurement results indicate that this CP antenna array has 10-dB return loss bandwidth from 52.5–65.5 GHz and 3-dB axial ratio bandwidth from 54–66 GHz. It is also found that the agreement between the measured and simulated results is not so good, which is mainly caused by the fabrication accuracy. The average gain within 55 to 64 GHz band is about 14 dBi.

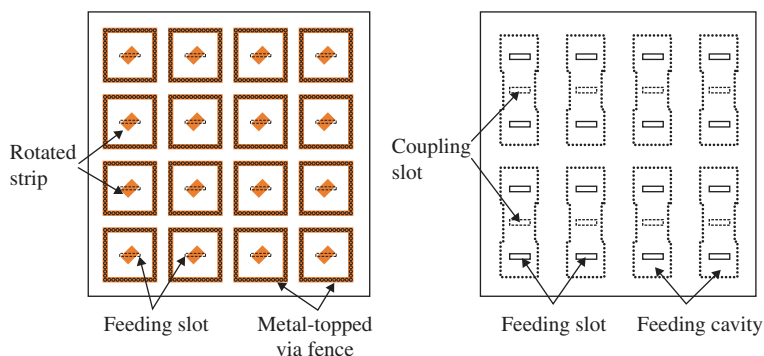


**Figure 5.61** The CP spiral antenna array (a) Simulated and measured  $S_{11}$  and (b) axial ratio [41]. Reproduced with permission of © 2012 IEEE

The transmission line loss and the surface wave loss are two factors that influence the radiation efficiency of the antenna array. To overcome these problems, a CP LTCC antenna array is designed by using the SIW as the antenna feed and introducing a metal-topped via fence to suppress the propagation of the surface wave [42]. SIW has the advantages of low transmission loss compared to the conventional microstrip line at the millimetre waveband. Figure 5.62 shows the structure of this CP antenna array. The antenna element is composed of a 10-layer LTCC substrate of Ferro A6-M and a rotated rectangular patch is used to achieve the CP radiation. A metal-topped via fence is introduced on the same layer of the patch antenna to suppress the surface wave propagation, which can improve the isolation level between adjacent antenna elements. It is shown that the isolation between the two



**Figure 5.62** Structure of the SIW feed CP LTCC antenna array [42]. Reproduced with permission of © 2012 IEEE



**Figure 5.63** Top view of the CP antenna array and the coupling slot for antenna feeding [42]. Reproduced with permission of © 2012 IEEE

adjacent elements can be improved by more than 25 dB with the existence of the metal-topped via fence. The patch antenna is fed by cutting a rectangular slot on the SIW.

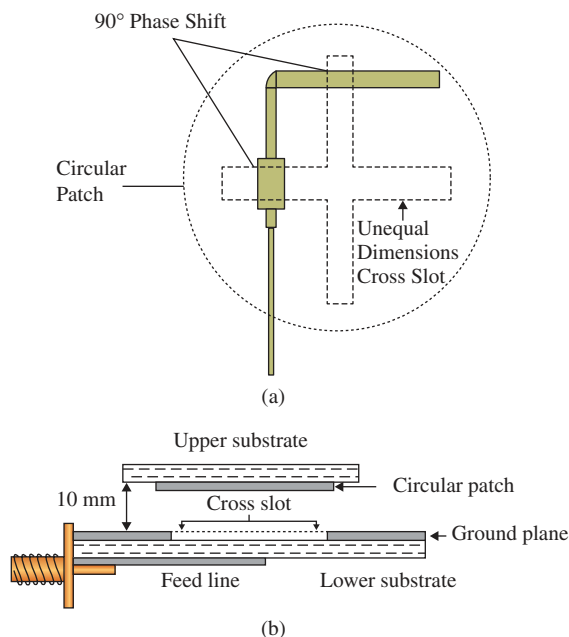
Figure 5.63 shows the top view of the  $4 \times 4$  array and the layout of the coupling slot under the antenna elements. The overall size of the antenna array is  $15.4 \times 15.4$  mm<sup>2</sup>. The distance between each antenna element is 3.8 mm, which is  $0.76\lambda_{60 \text{ GHz}}$ . The measurement results show that this SIW-fed CP antenna array has 6-dB return loss bandwidth from

56.65–65.75 GHz and 3-dB AR bandwidth of more than 7 GHz (60.2–67 GHz). Meanwhile, it exhibits a stable boresight gain better than 12.5 dBi within the AR bandwidth.

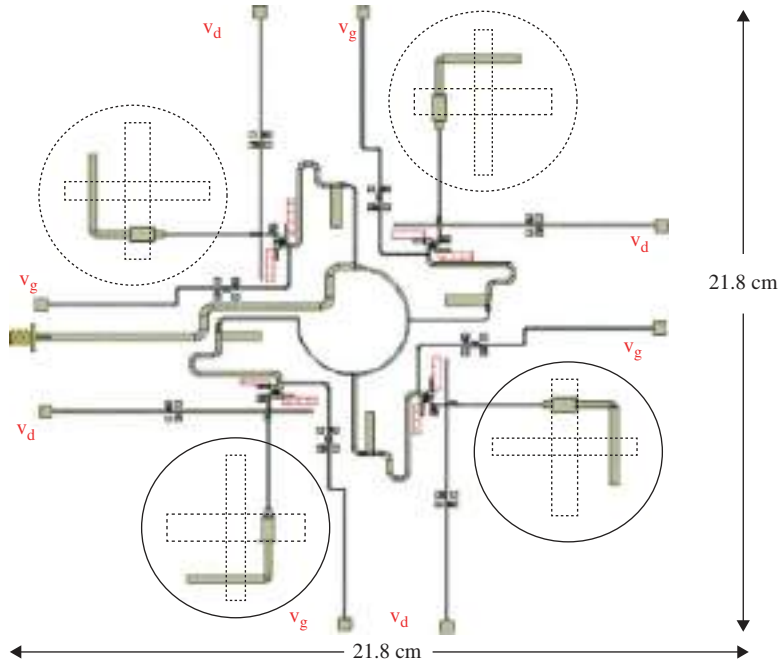
From these two examples, it can be observed that with high antenna integration level the complexity of the overall structure also increases, this brings challenges to the antenna fabrication techniques as there is always needs for accuracy fabrication with moderate cost.

### 5.6.2 CP Active Array

When comes to active CP antenna array design, it is critical to design a proper feed network for the radiating elements because the unequal input power level to the power amplifier can lead to a different compression level. One broadband high-efficiency CP active antenna array is presented in [43]. An aperture coupled circular patch fed by a L-shaped microstrip line is chosen as the antenna element for the CP array. Figure 5.64 shows the top and side view of this broadband CP antenna. This antenna is designed to resonate at 2 GHz. The circular patch is printed on the bottom side of a 0.79-mm thick Roger Duroid 5870 substrate ( $\epsilon_r = 2.2$ ). The cross slot and the L-shaped feed line is printed on another Roger Duroid 5870 substrate of the same thickness. There is a 10 mm air gap between the circular patch and the cross slot. The length of the L-shaped microstrip feed line between two slots is adjusted to be approximately a quarter of the wavelength at the desired operating frequency to introduce the  $90^\circ$  phase difference between two orthogonal modes. The dimensions of the crossed slot need to be optimized to achieve wideband circular polarization while the dimensions of the feed line can be optimized to give broad impedance bandwidth.



**Figure 5.64** Top and side view of the broadband CP antenna fed by an L-shaped microstrip line [43]. Reproduced with permission of © 2006 IEEE



**Figure 5.65** Layout of the CP active array [43]. Reproduced with permission of © 2006 IEEE

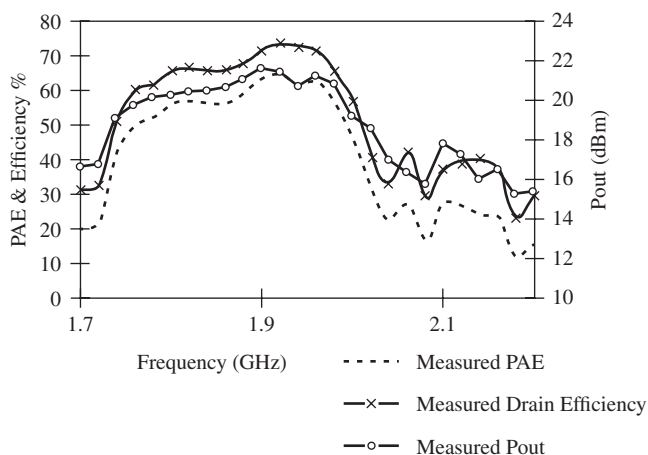
Figure 5.65 shows the layout of the  $2 \times 2$  active CP antenna array using the broadband CP antenna element presented in Figure 5.64. The feed network consists of seven quarter-wave transformers and each of the antenna element has an integrated class-E power amplifier. To ensure the ideal class-E operation, the load network of the amplifier is designed to give an open-circuit condition for all of the higher-order harmonic frequencies. The optimum source and load impedance are obtained by doing simulations in an Agilent ADS.

Figure 5.66 shows the measured PAE, drain efficiency and output of the power amplifier. The PAE of this AIA amplifier module is determined by using the standard definition of the PAE (equation 5.18) and the Friis transmission equation (equation 5.19). In these two equations,  $P_{out}$  and  $P_{in}$  represent the output and input power of the CP antenna array,  $P_{DC}$  is the power of the input DC signal,  $P_{re}$  is the received power,  $G_{sd}$  is the gain of the standard horn used during the measurement,  $G_A$  is the predetermined passive antenna gain and  $L_{cable}$  is the power loss in the cable.

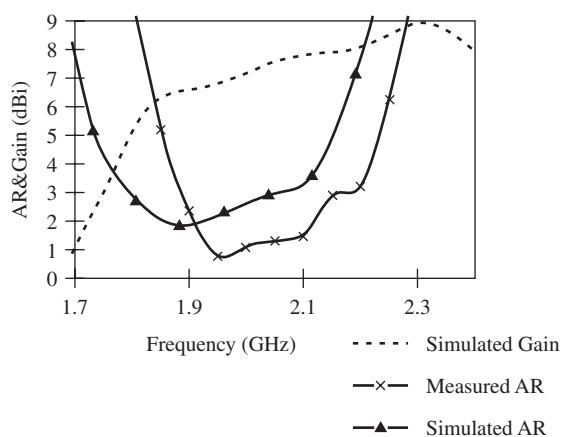
$$\eta = \frac{P_{out} - P_{in}}{P_{DC}} \times 100\% \quad (5.18)$$

$$P_{out} = \frac{P_{re}}{G_A G_{sd}} \times \left( \frac{4\pi r}{\lambda} \right)^2 \times L_{cable} \quad (5.19)$$

It can be seen that the active CP antenna array has a peak drain efficiency of 71.35% and a peak PAE of 64.8%. The PAE is over 50% within the 22.6% bandwidth from 1.72–2.16 GHz while the drain efficiency is over 60% from 1.82–2.14 GHz. The peak output power of the class-E PA array is 27.5 dBm at the central frequency (2 GHz).



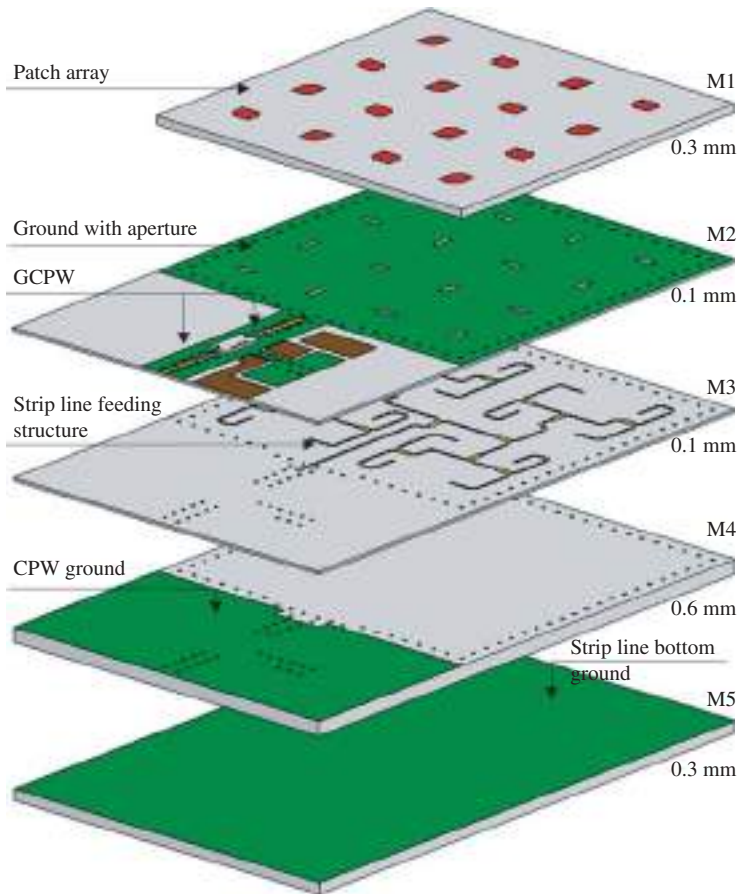
**Figure 5.66** Measured power-added efficiency (PAE), drain efficiency and output of the power amplifier [43]. Reproduced with permission of © 2006 IEEE



**Figure 5.67** Measured and simulated axial ratio of the active CP array [43]. Reproduced with permission of © 2006 IEEE

Figure 5.67 presents the measured and simulated axial ratio of the active CP array at the broadside. It has a 3-dB axial ratio bandwidth of 27% from 1.73–2.26 GHz and the simulation result shows that it has a peak gain of 9 dBi at 2.3 GHz.

To realize the highly integrated CP active antenna array, the LTCC technique and multilayer configuration can be employed, similar to the designs presented in last section. An integrated active CP array and LNA in LTCC for 60 GHz is presented in [44]. In this work, a  $4 \times 4$  CP patch antenna array, which is corner truncated and fed by sequential rotation techniques, and a 21 dB low noise amplifier (LNA) are packaged into the same substrate. Figure 5.68 shows the 3D exploded view of this multilayer CP antenna array. The active CP array antenna is designed in a Ferro-A6M LTCC package with a relative dielectric constant of  $\epsilon_r = 5.9$  and a loss tangent of  $\tan \delta = 0.002$ . It consists of five co-fired laminated ceramic

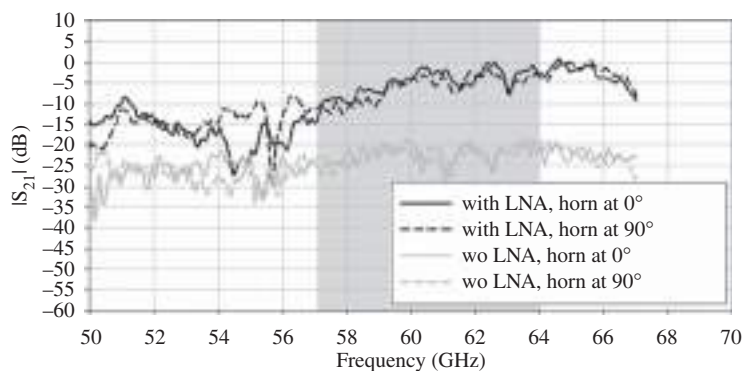


**Figure 5.68** 3D exploded view of the highly integrated CP antenna array [44]. Reproduced with permission of © 2011 IEEE

layers, where the radiating element is printed on the first layer, the LNA lies in the second layer, the feed lines and grounds are located in the other layers.

The final fabricated prototype has a compact size of only  $13 \times 20 \times 1.4 \text{ mm}^3$ . The simulation and measurement results reported in [44] demonstrates that with such high level of system integration, the antenna array exhibits good CP performances with an estimated overall peak gain of at least 35 dBi. Moreover, compare to the case that there is no LNA integrated, approximately 19 dB gain improvement has been observed in the 60.7–65 GHz band while the actual gain of the amplifier is around 21 dB. These measurement results are presented in Figure 5.69. The final estimated peak gain of this active antenna is at least 35 dBi.

By comparing these two types of CP active integrated antenna arrays presented in Figure 5.64 and Figure 5.68, it can be concluded that integrating the active components on the same layer of the antenna elements or feed network can result in a planar structure, which is easy to fabricate with low cost. However, the overall size of the antenna array can be relatively large. Employing the LTCC technique, multi-layer CP active integrated



**Figure 5.69** Measured transmission coefficient of the CP antenna array with and without the LNA [44]. Reproduced with permission of © 2011 IEEE

antenna arrays can be realized and different components can be printed in separate layers. This requires high fabrication accuracy, thus the fabrication cost is increased.

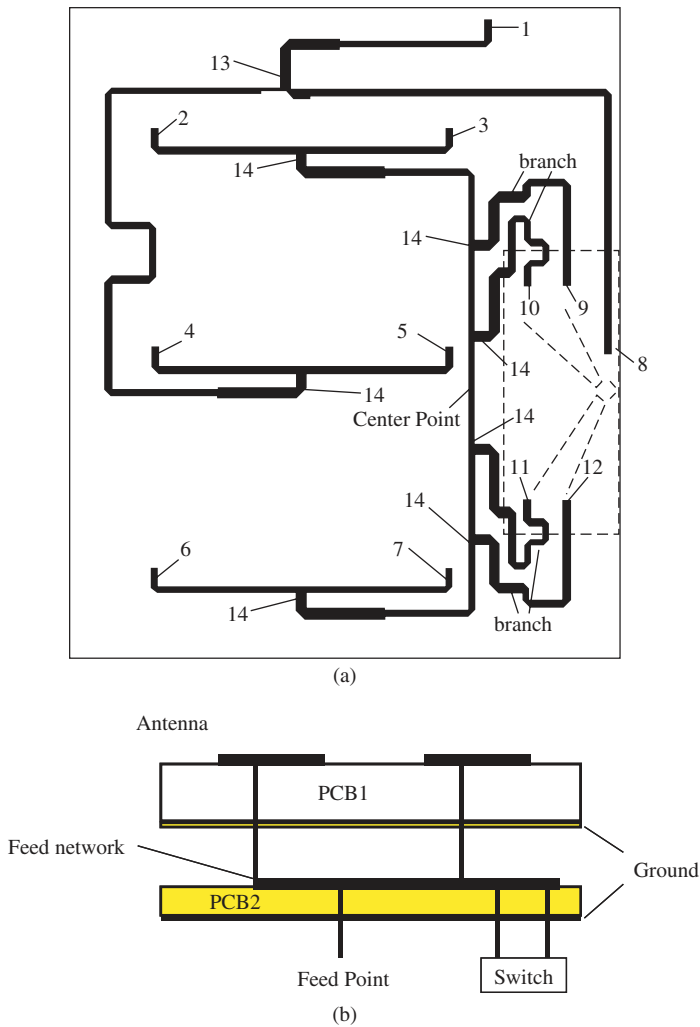
## 5.7 CP Array with Reconfigurable Beams

### 5.7.1 Beam-Switching CP Array using a Reconfigurable Feed Network

It is known that the total radiation field of the antenna array is determined by the vector addition of the far-fields radiated from the radiating elements. As the result, the radiation pattern of the antenna array can be controlled by adjusting the excitation amplitude and phase of the individual elements. The design of the CP antenna array has already been presented in the previous sections. In order to give the CP antenna array the ability to do the beam-switching or beam-steering, it is critical to design a suitable feed network that is able to control the phase as well as the amplitude of the excitation signal for each antenna element. One method that can be employed to design a beam-switching CP antenna array is to use a reconfigurable switch feed network, such as the one presented in [45]. The advantages of this technique are low cost and easy fabrication. Figure 5.70 shows the layout of the switch feed network proposed by [45] and the side view of the overall antenna array structure.

To switch the main beam to different angles, a single-pole multi-throw (SPMT) microwave P-I-N switch is used, which can switch between different feed lines with varied length and can result a CP array with different excitation phase. The function of the ports labeled in the Figure 5.70 is explained in Table 5.5.

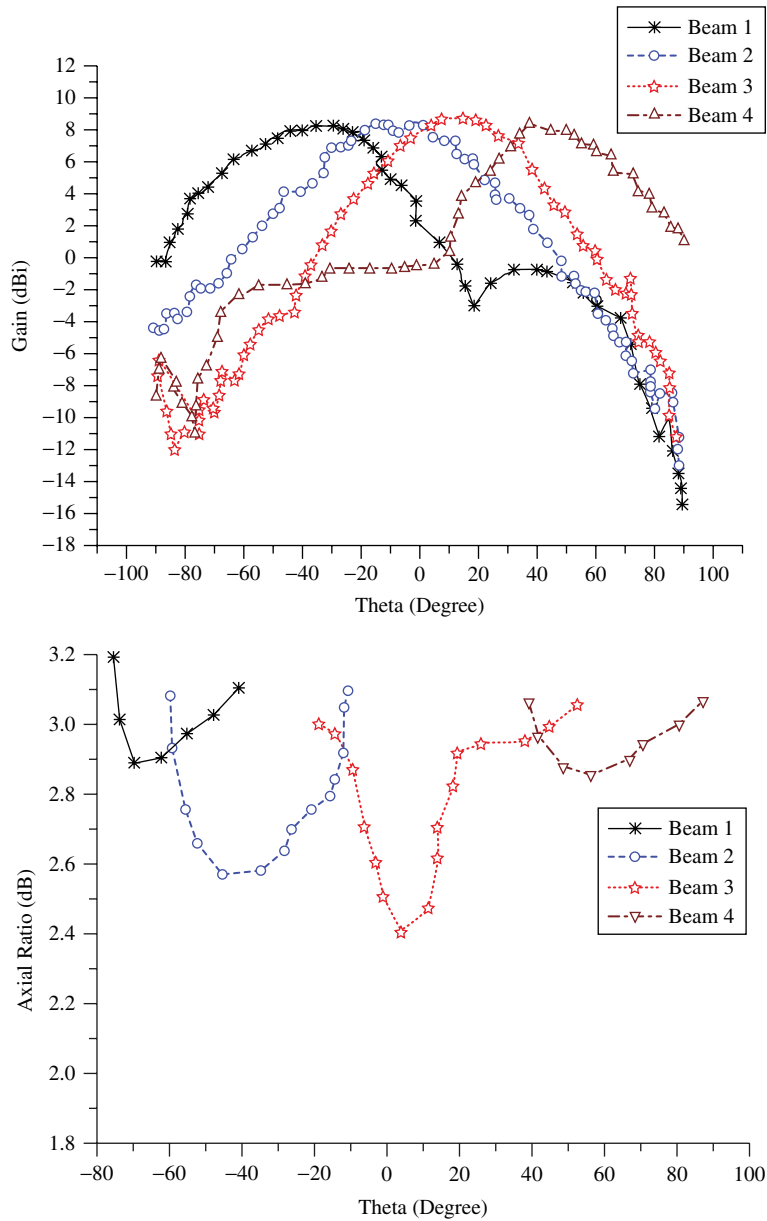
The antenna element used in this design is a simple square patche with a feeding probe on the diagonal line of the patch. By connecting the Port 8 to a different input port of the feed line (Port 9 to Port 12), several sets of phase distribution for the radiating element can be created. For instance, by connecting Port 8 to Port 10, the excitation phase of the antenna array is  $(0^\circ, 0^\circ, 140^\circ, 140^\circ, 275^\circ, 275^\circ)$ . The antenna array is designed to operate at 1.8 GHz and is printed on a 3.8-mm thick substrate with a relative dielectric constant of 4.316. The feed network is printed on a 0.787-mm thick PCB board with relative dielectric constant of 2.55. Figure 5.71 shows the measured radiation patterns and AR bandwidth of this array



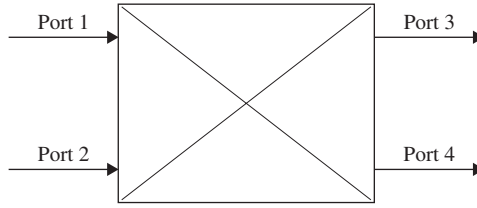
**Figure 5.70** The switch feed network for the switch-beam antenna array [45]. Reproduced with permission of © 2011 IEEE

**Table 5.5** Meaning of the ports shown in Figure 5.70

|                   | Function                                          |
|-------------------|---------------------------------------------------|
| Port 1            | Input port                                        |
| Port 2 to Port 7  | Output port for antenna feeding                   |
| Port 8            | Input port connected to the SPMT                  |
| Port 9 to Port 12 | Input port for feeding lines with different phase |



**Figure 5.71** Measured radiation patterns and AR bandwidth of the antenna array with different sets of excitation phase [45]. Reproduced with permission of © 2011 IEEE



**Figure 5.72** A generic version of a two-beam Butler Matrix

with different sets of excitation phase. It is found that this antenna array can have its main beam switched up to  $40^\circ$  with good CP radiation performance.

### 5.7.2 Beam-Switching CP Array using Butler Matrix Network

Another approach can be employed to design beam-switching CP antenna array is to use the Butler Matrix network. The Butler Matrix is a type of beam-forming network and has been extensively used in the field of beam switching antenna array design. It has N input ports and depends on which input port is excited, the beam of the antenna can be steered to a specific direction in one plane. Figure 5.72 shows a generic version of a two-beam Butler Matrix.

The scattering matrix of the two-beam Butler Matrix is [46]:

$$B = \left( \frac{1}{\sqrt{2}} \right) \times \begin{bmatrix} 0 & 0 & 1 & -j \\ 0 & 0 & -j & 1 \\ 1 & -j & 0 & 0 \\ -j & 1 & 0 & 0 \end{bmatrix} \quad (5.20)$$

Let  $I_i$  and  $Output_i$  respectively represent the input and output of this four-port system, as the one shown in Figure 5.72, it can be derived that:

$$\begin{bmatrix} Output_1 \\ Output_2 \\ Output_3 \\ Output_4 \end{bmatrix} = B \times \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \end{bmatrix} \quad (5.21)$$

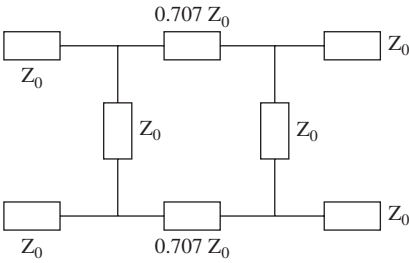
Solve this equation, we can get:

$$Output_1 = 0 \quad (5.22)$$

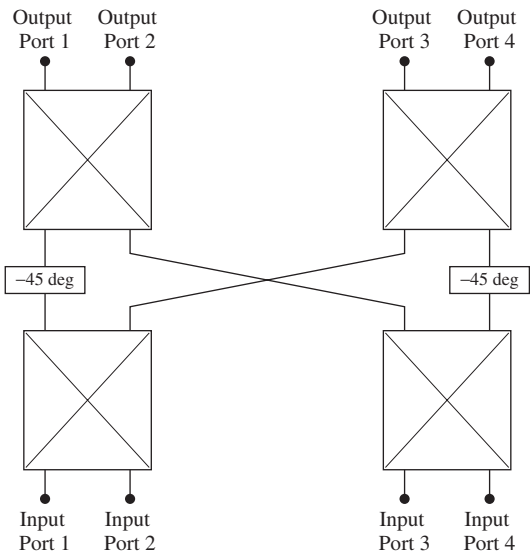
$$Output_2 = 0 \quad (5.23)$$

$$Output_3 = \left( \frac{I_1}{\sqrt{2}} \right) \times \angle 0^\circ \quad (5.24)$$

$$Output_3 = \left( \frac{I_1}{\sqrt{2}} \right) \times \exp \left( -\frac{\pi}{2} \right) \quad (5.25)$$



**Figure 5.73** Circuit model of the branchline coupler



**Figure 5.74** A generic version of a four-beam Butler Matrix

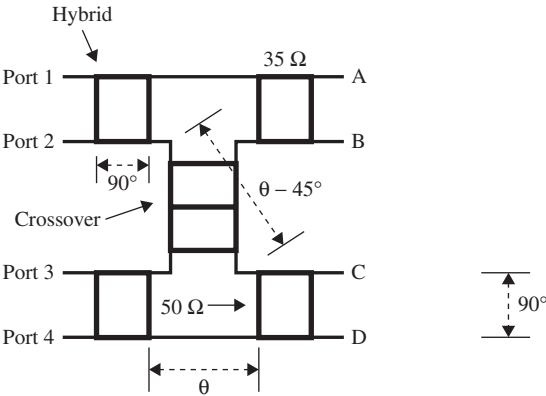
The Butler Matrix can be realized by a branchline quadrature coupler. The equivalent circuit model of the branchline coupler is depicted in Figure 5.73. In this model, each section of the transmission line has a length of quarter wavelength at the desired frequency.

To produce more beams, several two-beam Butler Matrixes can be connected by introducing the crossover structure. Figure 5.74 shows a generic version of a four-beam Butler Matrix. Table 5.6 summarizes the phase of output signals when different ports are excited.

Several beam-switching CP antenna arrays using the Butler Matrix as the feed network have been reported in the literature [47,48,49]. In [47], a planar microstrip antenna array with a Butler Matrix is designed to operate at 2.4 GHz for indoor wireless dynamic environments. The microstrip antenna array consists of four sequentially rotated inset-fed rectangular patch antennas to achieve CP radiation. Figure 5.75 shows the layout of the feed network used in this design. The  $4 \times 4$  Butler Matrix consists of four branchline couplers and a crossover to isolate the cross-lines. By selecting a different input port, different sets of the output phase can be produced and either RHCP or LHCP waves can be excited.

**Table 5.6** The phase of output signal when different ports are excited

|              | Output Port 1 | Output Port 2 | Output Port 3 | Output Port 4 |
|--------------|---------------|---------------|---------------|---------------|
| Input Port 1 | −45           | −90           | −135          | 180           |
| Input Port 2 | −135          | 0             | 135           | −90           |
| Input Port 3 | −90           | 135           | 0             | −135          |
| Input Port 4 | 180           | −135          | −90           | −45           |



**Figure 5.75** Layout of the feed network [47]. Reproduced with permission of © 2009 IEEE

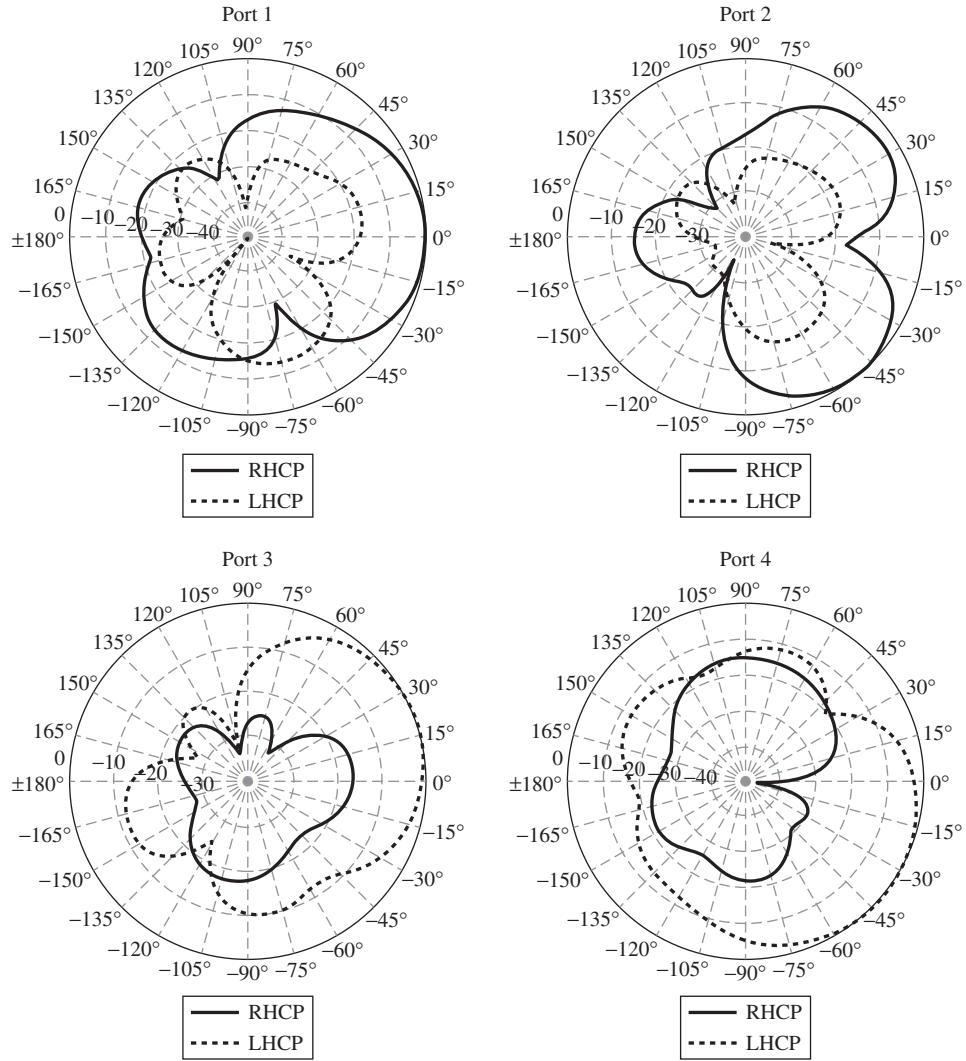
Figure 5.76 shows the measured normalized radiation patterns of this microstrip antenna array at 2.4 GHz when different input port is excited. As can be seen from the measured results, RHCP can be obtained when port 1 or port 2 is selected while LHCP can be generated when port 3 or port 4 is excited. Moreover, beam-switching can also be reached when port 1/port 3 or port 2/port 4 is selected.

Similar design approaches are demonstrated in [48,49], except that these CP arrays are designed to operate at different frequencies and different antenna elements are used. All of these designs have planar structures, which is simple to fabricate with low cost.

### 5.7.3 Beam-Switching Reflectarray using MEMS

Recently, a new concept of the beam-switching antenna array design has been proposed in [50]. In this work, a CP reflectarray antenna that is capable of switching the beam in both K and Ka bands is presented. The dual band operation of the reflectarray is realized by employing two different sizes of microstrip rings as the radiating elements and arranging them in the same lattice, such as the one depicted in Figure 5.77, where the microstrip rings of larger sizes are responsible for the operation at K band; the smaller rings are for the radiation at the Ka band.

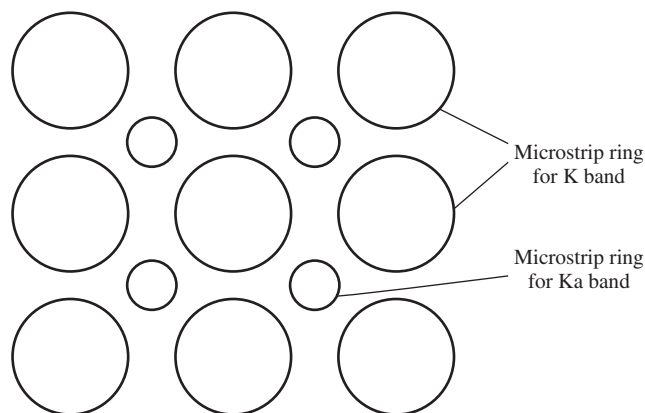
It is known that the beam of a reflectarray can be steered if the phase delay provided by the antenna element can be controlled. As a result, in [50] RF MEMS switches are introduced to



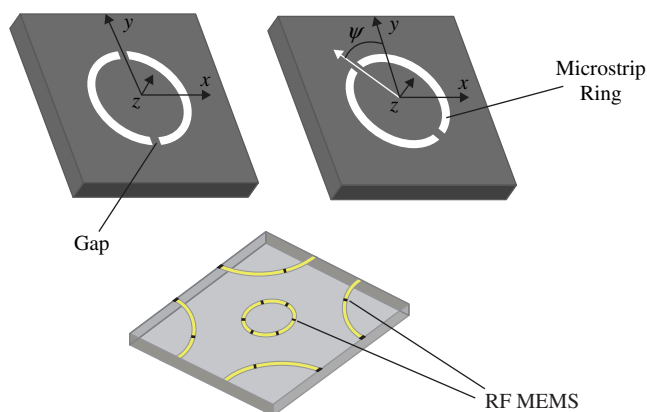
**Figure 5.76** Measured normalized radiation patterns of the microstrip antenna array with different excitation port [47]. Reproduced with permission of © 2009 IEEE

the microstrip ring to activate different slots on the ring, which is equivalent to controlling the gap rotation angle on the antenna element and thus modifying its corresponding phase response. Figure 5.78 shows the concept of the reconfigurable antenna unit cell.

The passive models of the reflectarray are fabricated to verify this design concept. The antenna elements are printed on a 0.5-mm thick substrate with a relative dielectric constant of 3.78. The outer and inner radius of the large ring are 2.4 and 2.2 mm, respectively. For the smaller size ring, the outer and inner radius are 1 and 0.8 mm, respectively. The final array consists of 109 large and 124 small split-rings. Figure 5.79 shows the measured and



**Figure 5.77** Layout of the dual band reflectarray [50]. Reproduced with permission of © 2012 IEEE



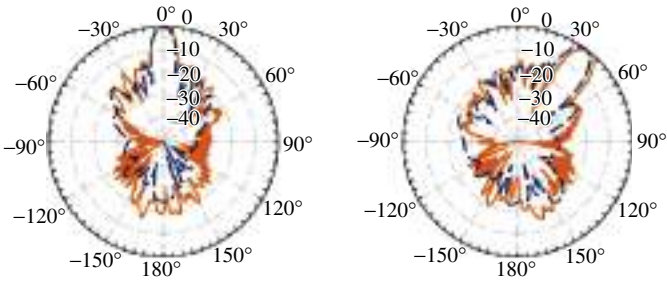
**Figure 5.78** Illustration of the reflective element with RF MEMS [50]. Reproduced with permission of © 2012 IEEE

simulated radiation pattern of the reflectarray at 24.4 GHz with different configuration of the slot position, which is determined by the status of the MEMS in the active model. It is observed that the beam of the reflectarray can be switched from broadside to  $35^\circ$ .

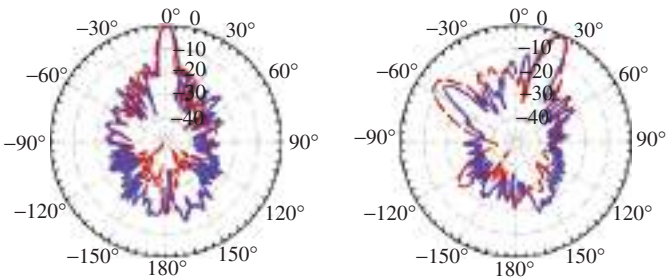
Figure 5.80 shows the measured and simulated radiation pattern of the reflectarray at 35.5 GHz. Similar to the results measured at the K band, it is found that the beam can be switched from broadside to  $24^\circ$ .

#### 5.7.4 Integrated Beam-Steering Array in LTCC

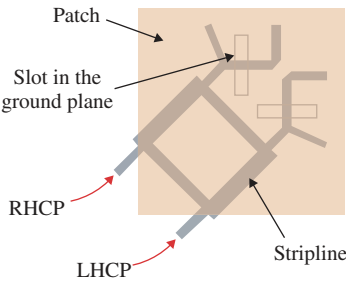
When the electronically steerable antennas are designed to operate at higher frequencies such as the Ka band, the system integration level and the packaging density is very high. In this case, there is a need to resort to the LTCC technology to fabricate the multilayer



**Figure 5.79** Measured (solid line) and simulated (dashed line) of the radiation patterns of the reflectarray when operating at the K band when the MEMS is switched ON (left) and OFF (right) [50]. Reproduced with permission of © 2012 IEEE

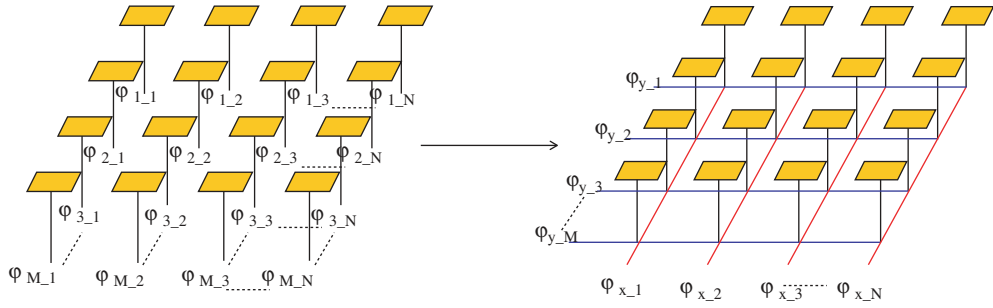


**Figure 5.80** Measured (solid line) and simulated (dashed line) of the radiation patterns of the reflectarray when operating at Ka band when the MEMS is switched ON (left) and OFF (right) [50]. Reproduced with permission of © 2012 IEEE

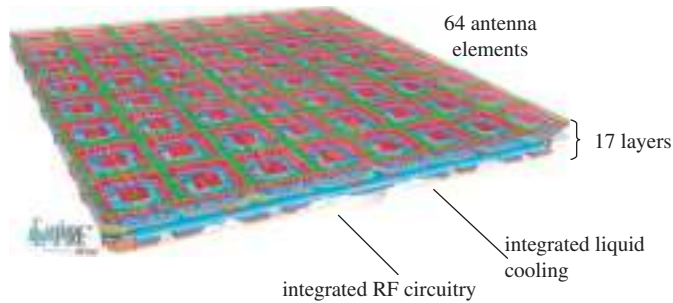


**Figure 5.81** The top view of the dual-CP antenna for 20 GHz operation [51]. Reproduced with permission of © 2010 EurAAP

structure. One highly integrated beam-steering active antenna array for Ka band application is reported in [51,52]. Two separate modules are required for this project, one is for the transmitter at 30 GHz and the other is for the receiver at 20 GHz. A dual-CP antenna element is used as the radiating element and Figure 5.81 shows the top view of the antenna. This antenna is composed of one square patch, one ground plane with two slots and fed by striplines. The stripline with a Butler Matrix network is used to electromagnetically couple the energy to



**Figure 5.82** Principle of beam steering of an  $N \times M$  array [52]. Reproduced with permission of © 2009 EurAAP



**Figure 5.83** 3D view of the LTCC Tx fronted module [51]. Reproduced with permission of © 2010 EurAAP

the square patch through the rectangular slots embedded on the ground plane. The RHCP and LHCP radiation of the antenna can be adjusted by using a different feeding port.

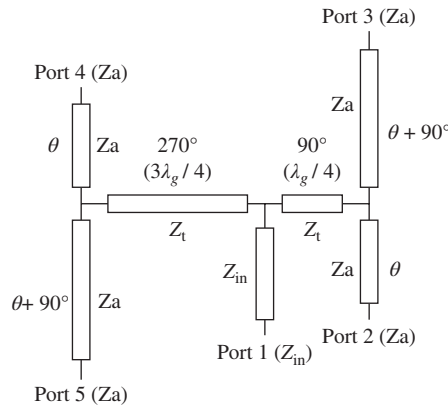
The basic principle of this system architecture is depicted in Figure 5.82. Instead of using a independent phase shifter for each antenna element, a dedicated phase shift in the x- direction as well as a dedicated phase shift in the y-direction is employed. In this way, the total amount of the control lines can be reduced from  $M \times N$  to  $M + N$ , where  $N$  is the number of element in X direction and  $M$  is the number of element in Y direction. This is especially useful when a large number of antenna elements is employed in an active antenna array.

Figure 5.83 shows the 3D view of the LTCC Tx fronted module. It has  $8 \times 8$  elements and consists of 17 LTCC layers with 18 metallization layers. This module includes antenna elements, a calibration network, active RF circuits, LO distribution networks, power and a DC supply and liquid cooling system, which is a highly integrated module.

## 5.8 Other CP Arrays

### 5.8.1 Compact Sequentially Rotated CP Arrays

The sequential-phase feed in conjunction with sequential rotated elements can improve the radiation performance of a CP array. However, when they are placed on the same layer



**Figure 5.84** Schematic diagram of the compact sequential-phase feed [53]. Reproduced with permission of © 2011 IEEE

of the array, the size of the feed network can increase the antenna element spacing. The increased element spacing could raise the side lobe level of the array. A compact sequential-phase feed using uniform transmission lines for CP sequential-rotation arrays is proposed in [53]. In this design, instead of using multiple segments of quarter wavelength transformers for impedance matching, the sequential-phase feed is realized by only using a single-stage transition and keeping the transmission line width uniform. Figure 5.84 shows the design schematic diagram of the equivalent transmission line model for this compact sequential-phase feed.

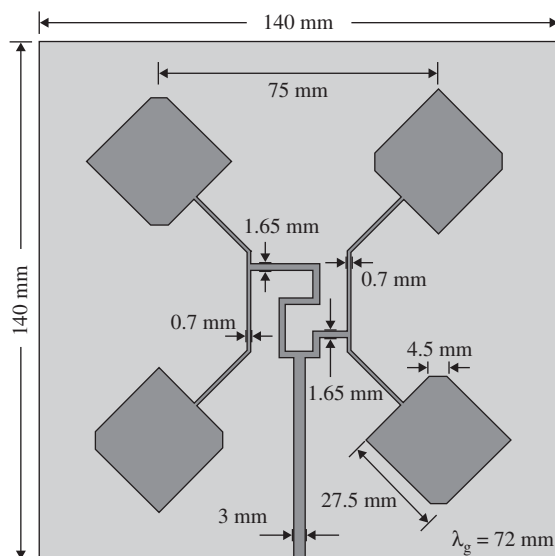
The relation between  $Z_{in}$ ,  $Z_a$  and  $Z_t$  satisfies:

$$Z_t^2 = Z_{in} \cdot Z_a \quad (5.26)$$

If the antenna impedance is designed to match the input port impedance, then:

$$Z_t = Z_{in} = Z_a \quad (5.27)$$

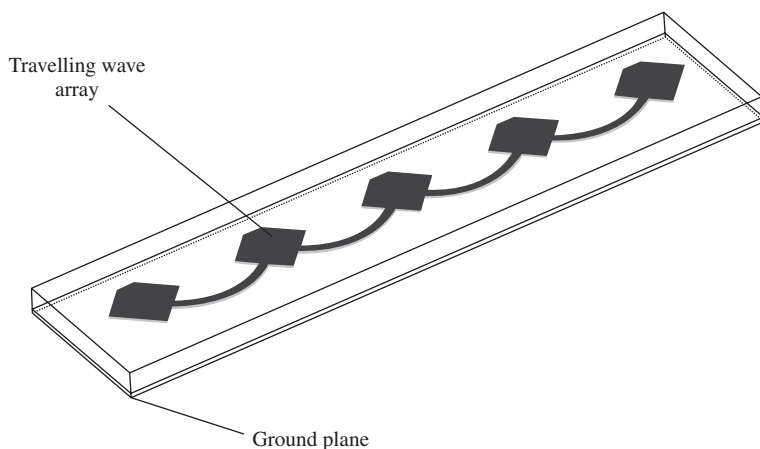
Figure 5.85 shows the layout of the  $2 \times 2$  sequential-rotated patch antenna array with the compact sequential-phase feed network. The patch used in this design is a square patch with two corners truncated for CP radiation. The array is designed on a 0.6-mm thick FR4 substrate and the central frequency of the array is 2.5 GHz. Meander lines are employed in the feed network to provide the desired phase delay and the size of the feed area is about  $0.25\lambda_{2.5 \text{ GHz}} \times 0.25\lambda_{2.5 \text{ GHz}}$ . Due to the coupling of the meander line segments, the dimensions of the feed is optimized in the simulation tool to get the optimum performance. The measurement results show that this compact sequential-phase feed has a bandwidth for the 1.5 dB balance level range from 2.1–2.9 GHz and  $1^\circ$  phase balance from 2.22 GHz to 2.76 GHz. The simulation results show that the  $2 \times 2$  array has 3-dB AR bandwidth from 2.45–2.6 GHz with a peak gain of 7 dBi at 2.52 GHz. Because of the small size of the feed network, this technique can be extended to larger scale  $2^N \times 2^N$  array without increasing the spacing between antenna elements.



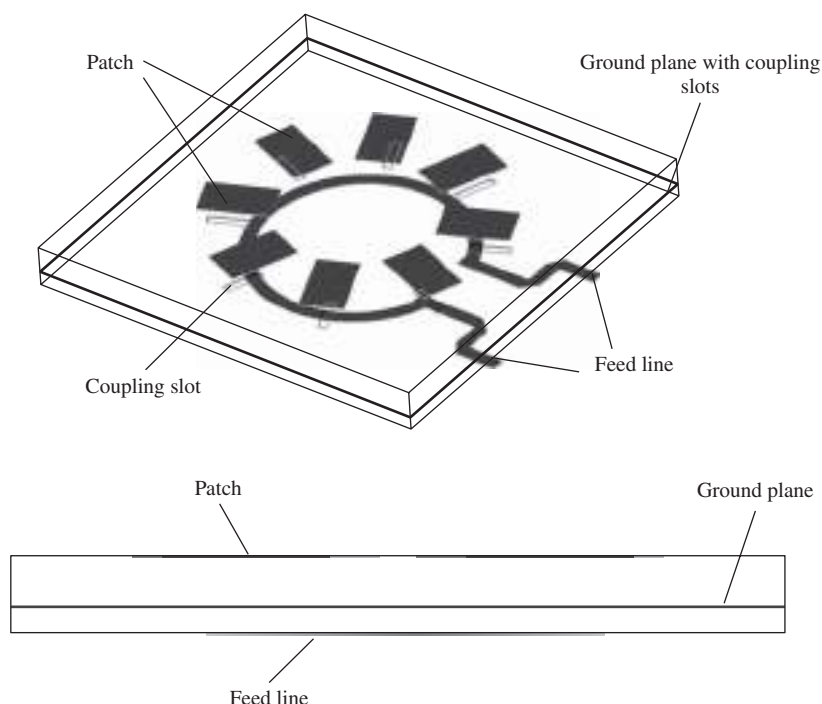
**Figure 5.85** Layout of a  $2 \times 2$  sequential-rotated patch antenna array with the compact sequential-phase feed [53]. Reproduced with permission of © 2011 IEEE

### 5.8.2 Travelling-Wave CP Arrays

The CP patch antenna array can be designed by using a series feed, which is simpler than using a corporate feed network. In the field of series feed antenna array design, the travelling-wave series feed array is preferred due to its advantage of wider bandwidth. One CP travelling-wave array antenna consisting of microstrip square patches is presented in [54]. Figure 5.86 shows the structure of this travelling-wave array. Each of the square patches has one truncated corner and is designed for operation at 5.6 GHz.



**Figure 5.86** Structure of the travelling-wave CP array [54]



**Figure 5.87** Configuration of the travelling-wave-fed circularly polarized concentric ring array [55]

This antenna array is printed on a Taconic RF-30-0300-CL1/CL1 substrate ( $\epsilon_r = 2.92$ ) with 0.76 mm thickness. The width of the feeding line is 2 mm to give  $50\ \Omega$  microstrip line impedance. The resonant frequency of the antenna can be controlled by adjusting the size of the square patch while the CP radiation of the antenna can be optimized by searching the optimum degree of truncating on the square patch. Meanwhile, feeding the array from the left side generates LHCP radiation and RHCP radiation can be obtained by feeding the array from the right side. The distance between each antenna element is 49 mm, which is about  $0.91\ \lambda_{5.6\text{GHz}}$ . The measurement results show that this antenna array has a 10-dB return loss of 12% and good CP radiation at central frequency (5.6 GHz). The gain of the antenna array is 12 dBi with cross-polarization 18 dB lower than the co-polarization at beam peak.

To avoid the radiation from the microstrip feedline, it is possible to design the travelling-wave-fed patch array using the aperture-coupled technique. Since this configuration means that the microstrip feedline lies below the ground plane, the unwanted radiation from the feed network can be suppressed. Another advantage of using aperture-coupled feeding is that it can improve the operation bandwidth of the antenna. An X-band travelling-wave-fed CP concentric ring array using linearly polarized patch antennas as radiating elements is presented in [55]. Figure 5.87 shows the configuration of this proposed antenna array. The radiating elements, square patches, are printed on the top side of the 1.6-mm thick RT/Duriod 5880 ( $\epsilon_r = 2.2$ ) substrate and fed by microstrip lines on another 0.5 mm RT/Duriod 5880. The apertures are cut in the shared ground plane of these two layers. The patch antenna elements are positioned in a sequential rotated way along the concentric slot ring to produce

**Table 5.7** Measurement results of the antenna array proposed in [55]

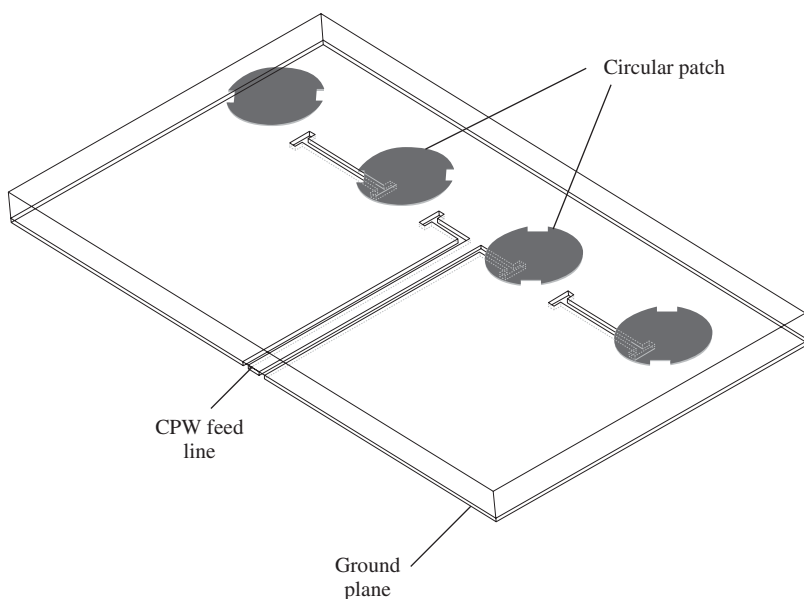
|                           | Four-element array        | Eight-element array      |
|---------------------------|---------------------------|--------------------------|
| 3-dB AR bandwidth         | 7.2% ( $f_0 = 10.05$ GHz) | 7.2% ( $f_0 = 10.1$ GHz) |
| AR at central frequency   | 0.6 dB                    | 0.8 dB                   |
| Gain at central frequency | $\sim 9$ dBi              | $\sim 12.5$ dBi          |
| 10-dB impedance bandwidth | $>20\%$                   | $>20\%$                  |

the desired CP waves. To ensure that each radiation element receives the same coupled power, progressive offset distance of the patches from the slotline is used and its value is optimized by performing numerical simulations.

One four-element and one eight-element array are designed and fabricated by [55] and the measurement results are summarized in Table 5.7. The measurement results show that both of the antenna arrays exhibit good AR bandwidth and CP radiation at the desired frequency band. The measurement results also indicate that both arrays have a very symmetric radiation pattern in the broadside direction, which implies that each radiation element contributes equal radiation power. However, it is important to point out that to reach the desired radiation performance, the alignment accuracy needs to be high as the misalignment of more than 5% can significantly influence the antenna performance [55].

### 5.8.3 CPW-Fed CP Arrays

CPW technique can also be employed in the CP patch antenna array design. The CPW technique has attracted much research interest in recent years due to its advantage of low

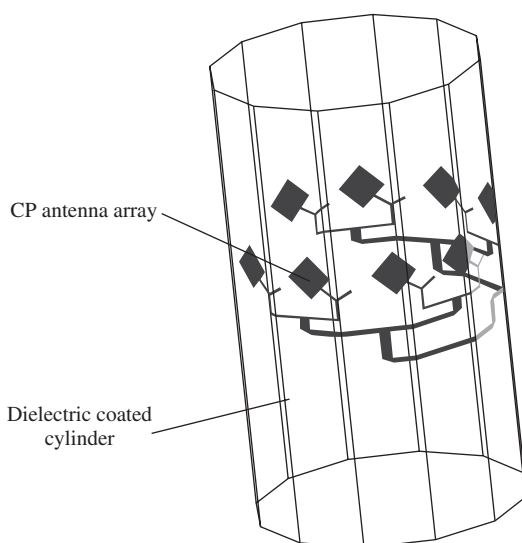
**Figure 5.88** Structure of the CPW-fed CP patch antenna array [56]

radiation losses, less dispersion and easy integration to the system. The CPW-fed CP patch antenna array has also been studied, as the one presented in [56]. The structure of this antenna array is given in Figure 5.88.

This  $1 \times 4$  array is designed on a 1.6-mm thick FR4 substrate for operation at 4 GHz. The circular patches and the feeding CPW are placed on different sides of the substrate and all the patches are aligned in the transverse direction of the CPW. Two perturbed segments,  $45^\circ$  inclined from the main axis, are cut on each of the circular patch to realize CP radiation. The antenna elements are fed by slot lines of  $100 \Omega$  and a T-shaped CPW-to-slotline feeding structure is used. The measurement results show that this antenna array has a gain of 9.57 dBi with 3-dB AR bandwidth of 0.9%. Within the AR bandwidth, the return loss is always better than 15 dB.

#### 5.8.4 Omni-directional CP Antenna Arrays

For certain applications, such as space vehicles, it is required to have an omni-directional CP antenna array to maintain a good communication link with a central receiving units and avoid selective reception. One method of obtaining an omni-directional radiation pattern is to design an conformal CP antenna array[57]. Figure 5.89 shows the generic configuration of the conformal CP patch array wrapped as a cylindrical dielectric cylinder. The patch antenna array employed in this design is an eight-element CP array printed on a 1.58-mm thick substrate with relative permittivity of 2.2. The two corners at the diagonal position of the patch are truncated in order to generate two orthogonal modes and thus obtain CP radiation. This conformal CP antenna array is designed to resonate at 2.25 GHz and it is wrapped into a cylinder with diameter of 20 cm. It is found that effect of the curvature is negligible for mounting on a cylinder whose radius is greater than one guided wavelength.



**Figure 5.89** Conformal CP patch array on a dielectric cylinder [57]

**Table 5.8** A summary of the CP array antenna in literature

| Ref. | Type                 | $f_0$ (GHz) | element number | Max. Gain (dBi) | 3-dB AR bandwidth (%) | Fabrication Complexity |
|------|----------------------|-------------|----------------|-----------------|-----------------------|------------------------|
| [5]  | Stacked-Patch        | 10.2        | $2 \times 2$   | 12.5            | 23.5                  | medium                 |
| [7]  | Dual-feed Patch      | 3.55        | $2 \times 2$   | 12.8            | 30                    | medium                 |
| [11] | Single-feed Patch    | 29          | $8 \times 8$   | ~25             | 13.8                  | low                    |
| [15] | DRA                  | 4.4         | $2 \times 2$   | 12              | 16                    | medium                 |
| [19] | Slot Array           | 2.4         | $2 \times 2$   | 9               | 15                    | low                    |
| [20] | Slot Array           | 60          | $2 \times 4$   | 15.6            | ~25                   | medium                 |
| [23] | Waveguide Slot Array | 7.55        | 15             | 18              | >1.3                  | medium                 |
| [26] | RLSA                 | 7.5         | —              | 25              | >7                    | medium                 |
| [29] | Reflectarray         | 7.1         | >100           | 27.6            | 18                    | medium                 |
| [34] | Reflectarray         | 10          | 81             | 19.4            | 28                    | medium                 |
| [44] | Integrated CP array  | 60          | $4 \times 4$   | 35              | 15                    | high                   |

Instead of using a patch antenna array, cross-shaped slot antenna array can also be used [58], which exhibits similar radiation performance to the one presented in [57].

## 5.9 Summary

This chapter presents a review of various CP arrays, including CP patch arrays, CP DRA arrays, CP printed slot arrays, CP waveguide slot arrays, CP reflectarrays, CP active arrays, millimetre-wave CP arrays in LTCC and CP arrays with beam reconfigurability. Table 5.8 summarizes the performance of some typical CP arrays available in the literature. It is found that among antennas listed in this table, the reflectarray presented in [34] and the CP patch array reported in [5] have the most wide 3-dB AR bandwidths. The broadband performances of both arrays are achieved by employing a multilayer configuration with two layers of radiating elements. The dual-feed patch array proposed in [7] can reach a broad 3-dB AR bandwidth of 30% by employing a thick (11 mm air layer) low permittivity substrate and patches fed by slot coupling technique. The integrated CP array [44] is small but it has high fabrication complexity and LTCC technology is required to realize such multilayered structure.

As there are a huge number of publications on CP arrays, the chapter can only select a limited amount of results from literature. Table 5.8 can only provide readers a general idea due to limited examples included in the table. A good reference on CP antenna arrays is [59].

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# 6

## Case Studies

### 6.1 Introduction

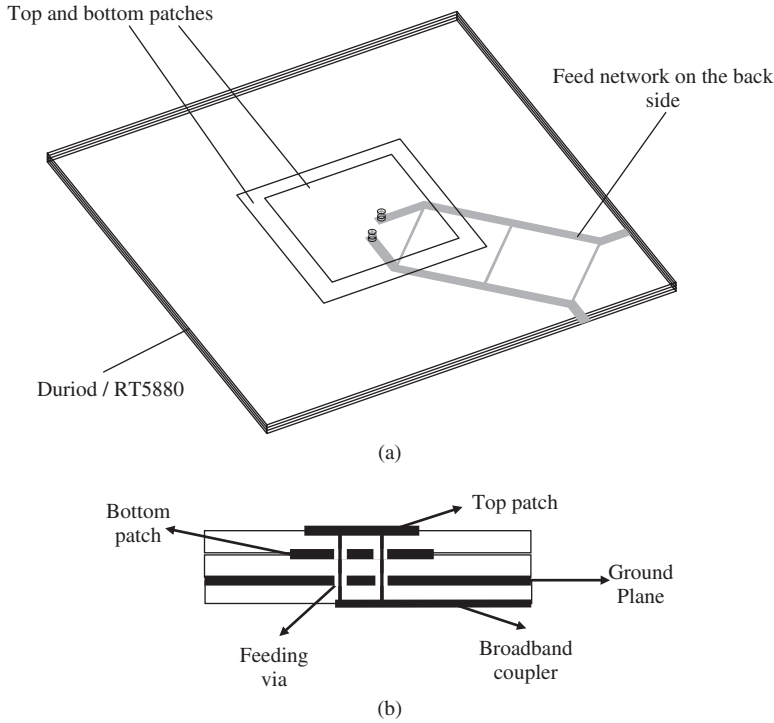
In previous chapters, different techniques for designing CP antenna elements and arrays have been presented. This chapter aims to provide step-by-step guidance to readers who would like to get started on CP antenna and array design. Design procedures of some different types of CP elements and arrays are described in detail. It is expected that readers will be able to follow the design steps as given in the case studies.

Five case studies; one dual band CP array for GPS application (1.575 GHz and 1.227 GHz), one small PQHA for SATCOM (2 GHz), one broadband CP antenna for RFID reader (2.4 GHz), one CP reflectarray for Ka band (30 GHz) application and one wideband logarithmic spiral antenna that can operate from 3.4–7.7 GHz, will be presented. The design guidelines for these CP antennas will be given and detailed results are presented.

### 6.2 Dual-Band CP Patch Array for GNSS Reflectometry Receiver on Board Small Satellites

The Global Navigation Satellite System (GNSS) is a system of satellites that provides autonomous geo-spatial positioning with global coverage. Receivers can determine their location precisely by calculating the time signals received from GNSS satellites. Although the major market for GNSS remains with guidance and navigation users at the moment, GNSS remote sensing has become increasingly important for the scientific and research community due to the availability of GNSS signals for studying climate changes, ocean roughness and salinity, soil moisture, ice, wind speed, disaster monitoring, atmosphere and ionosphere measurement. GNSS reflectometry is a passive remote-sensing instrument which uses a receiver to collect GNSS signals reflected from earth surfaces. GNSS reflectometry receivers are low cost and can be put on board small satellites in low earth orbits.

To design one antenna array that can be mounted on a small remote sensing satellite, besides being light and compact, it is required that the antenna array can operate at both GPS L1 (1.575 GHz) and L2 (1.227 GHz) bands with 15 and 4 MHz bandwidth, respectively.



**Figure 6.1** Top and side view of the dual band CP patch antenna

Moreover, the antenna array should radiate LHCP waves and have a gain greater than 11 dBi with a shaped beam pattern ( $20^\circ$  along array axis and  $80^\circ$  across array axis).

The patch antenna is chosen for this application due to its advantages of low cost, easy fabrication and light weight. This work starts with designing one dual band CP patch antenna operating GPS L1 and L2 bands. The antenna element has a stacked structure. On the top, there is one dual-fed square patch for higher band operation. Below that patch, there is a second patch for lower band operation which is fed by electromagnetic coupling. Figure 6.1 shows the top and side views of the proposed multi-layer dual band CP patch antenna [1].

The substrates used in this work for both antenna and feed network is Duriod/RT 5880 ( $\epsilon_r = 2.2$ ) with a thickness of 1.6 mm. The initial dimensions of patch antenna can be calculated by using the formulas given in [2]:

$$W = \frac{c}{2f_0} \sqrt{\frac{2}{\epsilon_r + 1}} \quad (6.1)$$

$$L = L_{eff} - 2\Delta L \quad (6.2)$$

where  $W$  is the width of the patch,  $L$  is the length of the patch,  $c$  is the speed of light,  $f_0$  is the resonant frequency of the patch and  $\epsilon_r$  is the relative dielectric constant of the substrate.

The  $L_{eff}$  and  $\Delta L$  can be obtained by:

$$L_{eff} = \frac{c}{2f_0 \sqrt{\epsilon_{eff}}} \quad (6.3)$$

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[ 1 + 12 \frac{h}{W} \right]^{-1/2} \quad (6.4)$$

$$\Delta L = 0.412h \frac{(\epsilon_{eff} + 0.3) \left[ \frac{W}{h} + 0.264 \right]}{(\epsilon_{eff} + 0.258) \left[ \frac{W}{h} + 0.8 \right]} \quad (6.5)$$

where  $h$  is the height of the substrate. In this study, it is decided to use the square patch because we plan to design a CP antenna with dual-feed; thus it is important to keep the patch with a symmetry structure. After calculating the initial dimensions of the patch, this patch antenna is simulated in CST microwave studio [3] to be optimized.

The top square patch, which has a side length of 62 mm, is designed to operate at the GPS L1 band at 1.575 GHz. To achieve the CP radiation, this patch is fed by two probes with  $90^\circ$  phase difference as depicted in Figure 6.1. The bottom patch, which is for GPS L2 band operation at 1.227 GHz, is printed on another substrate layer and has dimensions of  $79.6 \times 79.6 \text{ mm}^2$ . This patch is electromagnetically coupled to the top patch. It is found that good impedance matching can be reached for the higher band patch when the feeding position is close to the centre of the patch. However, for the lower band patch, good impedance matching can be obtained when the feeding position is close to the edge of the patch. Therefore, it is important to find an optimal position to locate the feeds to achieve promising impedance matching for both bands. After doing some numerical studies in the CST microwave studio, the feeding position is selected to be 9.5 mm away from the centre of the patch.

A broadband three-branch coupler is used as the feed network [4]. Figure 6.2 shows the layout of the three-branch coupler. Compared to the branchline coupler presented in Figure 5.73 in Chapter 5, this three-branch coupler can provide wider bandwidth. This coupler is designed to provide the required 3-dB power split and phase quadrature over the required frequency band.

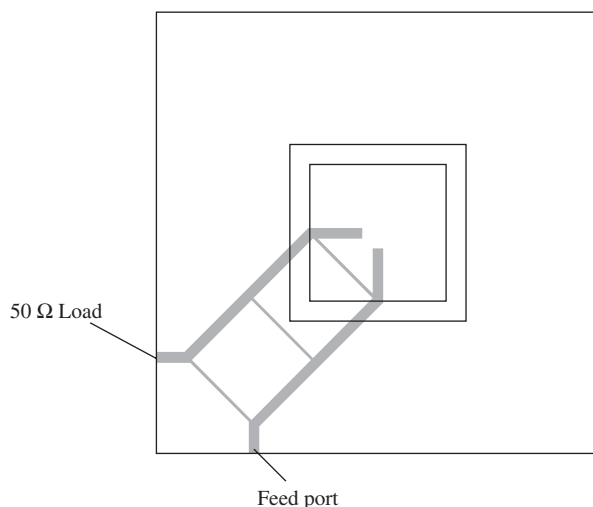
As shown in Figure 6.2, the three-branch coupler can be divided into seven sections, each of them is a microstrip line with a length of quarter wavelength. The equivalent circuit model of this three-branch coupler is given in Figure 6.3. The characteristic impedance of the microstrip lines are presented by  $Z_0$ ,  $Z_1$  and  $Z_2$ .

The characteristic impedance of microstrip lines can be calculated by using the formulas:

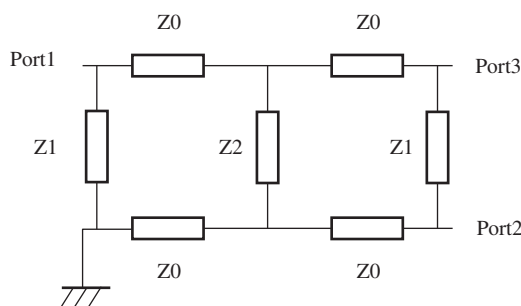
$$Z_1 = (1 + \sqrt{2}) \times Z_0 \quad (6.6)$$

$$Z_2 = \sqrt{2} Z_0 \quad (6.7)$$

This three-branch coupler is simulated in Agilent ADS[5] and Figure 6.4 shows the simulated transmission coefficient and phase of the output signal at Ports 2 and 3 as labelled in Figure 6.2. As can be seen from the simulation results, at the desired frequency (1.227 GHz), the power of the input signal from Port 1 is split equally with a  $90^\circ$  phase difference.



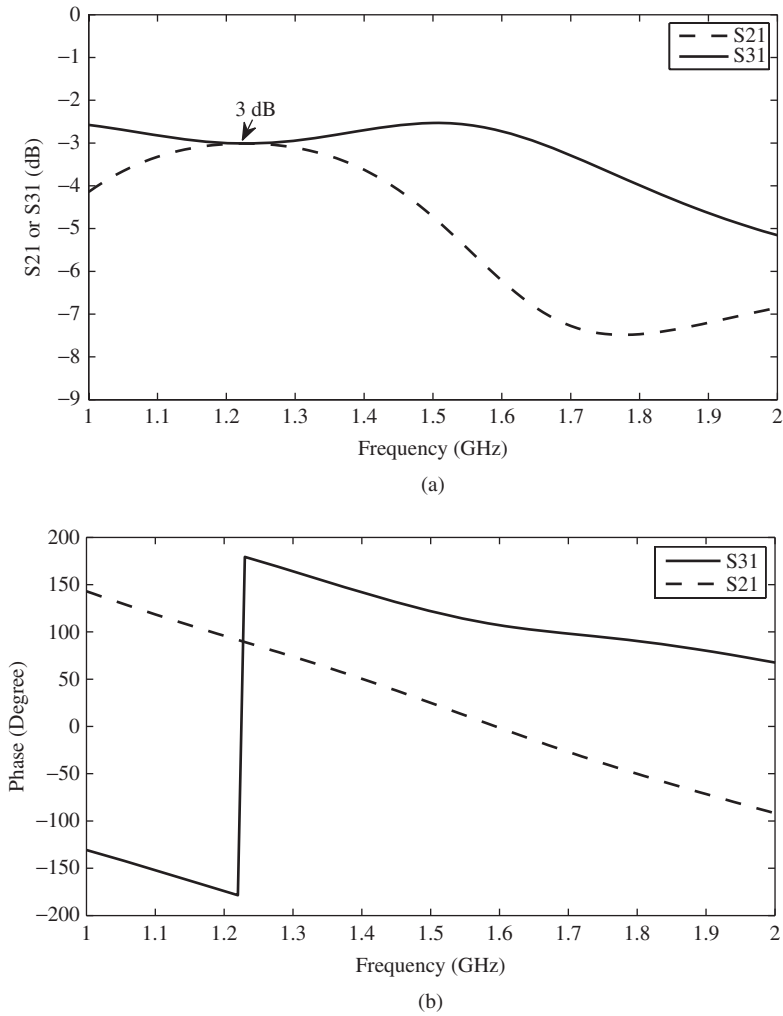
**Figure 6.2** The layout of the broadband three-branch coupler [1]



**Figure 6.3** Equivalent circuit model of the three-branch coupler

After the theoretical calculation, the dimensions of the branch coupler are further optimized in the simulation tool. This antenna is fabricated and measured at Surrey Space Centre, University of Surrey. Figure 6.5 presents the measured and simulated  $S_{11}$  of the antenna prototype. It is seen that the antenna can achieve good impedance matching at both GPS L1 and L2 bands. There is a good agreement between the simulated and measured results except some frequency shifts, which is caused by the fabrication accuracy and numerical errors. The resonant frequencies can be easily tuned by re-sizing the side length of the top and bottom square patches. There is a big dip at around 1.4 GHz, which is the resonant frequency of the branch line coupler.

Figure 6.6 presents the measured radiation patterns of the proposed antenna element at the central frequencies of the GPS L1 and L2 band. It can be observed that this antenna exhibits a hemi-spherical radiation pattern with a 3-dB beam-width of nearly  $80^\circ$ , which fulfills the design objectives.

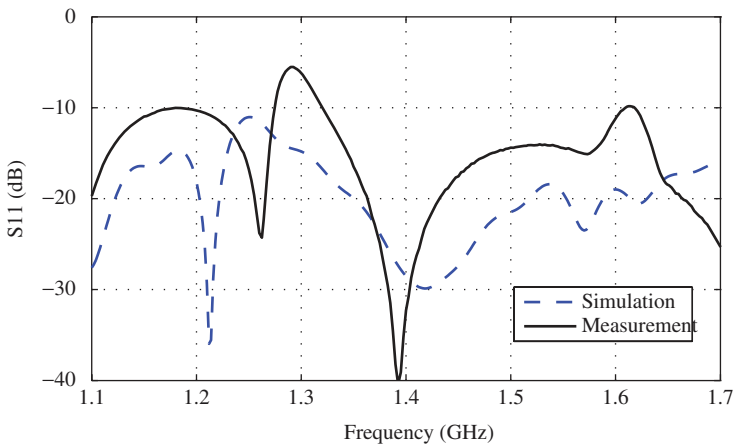


**Figure 6.4** Simulated transmission coefficient and phase of the output signal of the three-branch line coupler

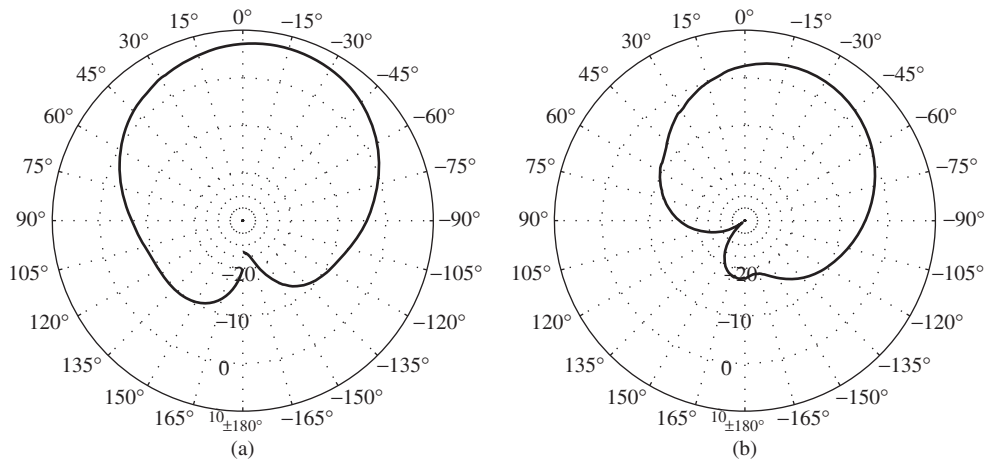
A summary of the measured gain and simulated axial ratio of this antenna element is shown in Table 6.1. It can be seen that within the frequency band interested, the gain is always better than 3 dBi.

To achieve higher gain, it is decided to use an array antenna and the antenna designed here will be one element in the array. Four patch elements are used and the distance between each element is about  $0.6\lambda_{1.2 \text{ GHz}}$ , where  $\lambda_{1.2 \text{ GHz}}$  represents the free space wavelength at 1.2 GHz. This distance is determined by allowing sufficient space to allocate the feed network. The configuration of this antenna array is presented in Figure 6.7.

The broadband three-branch coupler used for the single antenna element is employed in the array design and one power divider, which consists of several quarter wave transformers for matching  $100\text{--}50\Omega$  is introduced, as demonstrated in Figure 6.8.



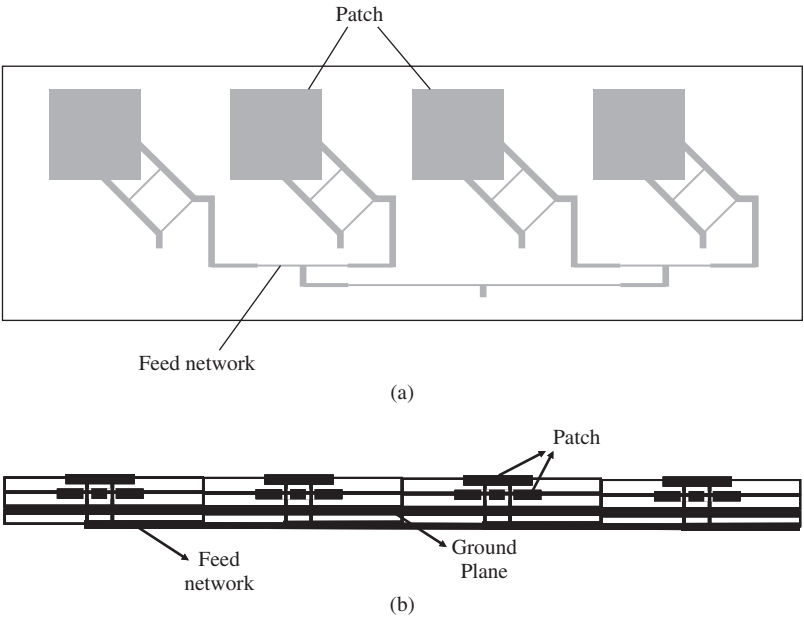
**Figure 6.5** Comparison of the measured and simulated S11 of the dual band GPS antenna



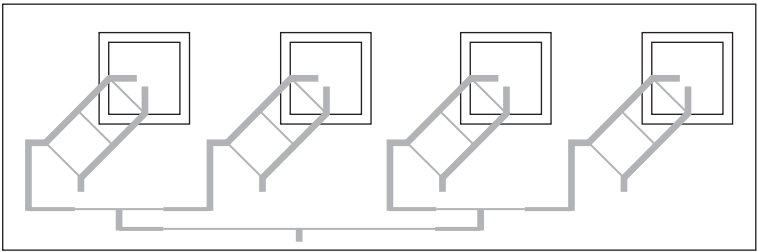
**Figure 6.6** Measured radiation patterns of the proposed antenna at GPS L1 1.575 GHz (left) and L2 band 1.227 GHz (right)

**Table 6.1** Measured gain and simulated AR of the dual band GPS patch antenna

| Frequency (GHz) | Measured Gain (dBi) | Simulated AR (dB) at $\theta = 0^\circ, \varphi = 90^\circ$ |
|-----------------|---------------------|-------------------------------------------------------------|
| 1.225           | 4.1                 | 2.66                                                        |
| 1.227           | 3.9                 | 2.71                                                        |
| 1.229           | 3.6                 | 2.77                                                        |
| 1.5675          | 5.5                 | 1.42                                                        |
| 1.575           | 6.3                 | 1.10                                                        |
| 1.5825          | 6.8                 | 0.86                                                        |



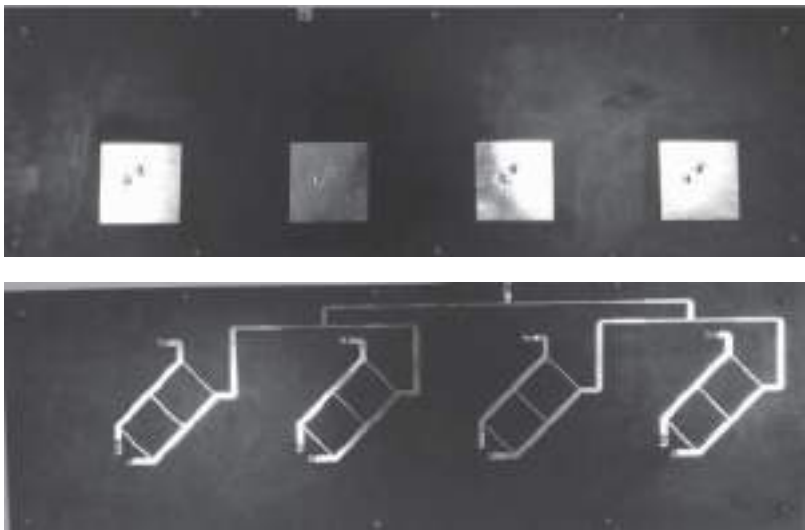
**Figure 6.7** Top and side view of the four element dual band antenna array



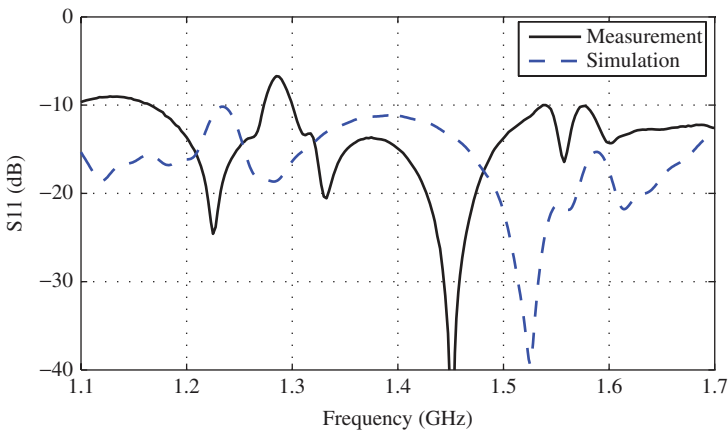
**Figure 6.8** The layout of the feed network for the antenna array [1]

Figure 6.9 shows the photos of fabricated array antenna. Both the patch array (on the top) and the feed network (at the back) are shown. Figure 6.10 presents the measured and simulated return loss of this antenna array prototype. The measurement results show that this antenna array can operate properly in the required GPS L1 and L2 band. The simulation result matches well with the measurement result except for some frequency shifts, which is caused by the fabrication accuracy.

The simulated radiation performances of this antenna array at GPS L1 and L2 are summarized in Tables 6.2 and 6.3. It can be seen that at both bands, the antenna array exhibits peak gain better than 11 dBi, which meets the requirements for the GNSS remote sensing for small satellite application. Meanwhile, the axial ratio at broadside is always smaller than 1.8 dB within the GPS L1 and L2 bands.



**Figure 6.9** Photo of the fabricated antenna array



**Figure 6.10** Measured and simulated S11 of the fabricated dual band GPS antenna array

**Table 6.2** Antenna performance across GPS L1 band

|                                                    | 1.5675 GHz | 1.575 GHz | 1.5825 GHz |
|----------------------------------------------------|------------|-----------|------------|
| Peak Gain at $\varphi = 0^0$ (dBi)                 | 11.8       | 13.5      | 13.8       |
| Peak Gain at $\varphi = 90^0$ (dBi)                | 11.9       | 13.5      | 13.9       |
| Axial Ratio at $\theta = 0^0, \varphi = 90^0$ (dB) | 1.71       | 1.35      | 1.03       |

**Table 6.3** Antenna performance across GPS L2 band

|                                                            | 1.225 GHz | 1.227 GHz | 1.229 GHz |
|------------------------------------------------------------|-----------|-----------|-----------|
| Peak Gain at $\varphi = 0^\circ$ (dBi)                     | 11.6      | 11.9      | 12.1      |
| Peak Gain at $\varphi = 90^\circ$ (dBi)                    | 11.6      | 11.9      | 12.1      |
| Axial Ratio at $\theta = 0^\circ, \varphi = 90^\circ$ (dB) | 0.78      | 0.94      | 1.10      |

### 6.3 Small Printed Quadrifilar Helix Antenna for Mobile Terminals in Satellite communications

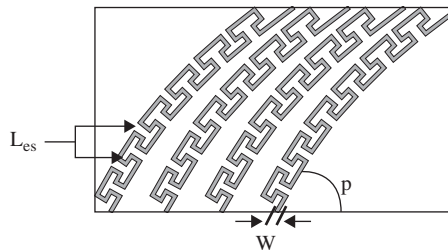
PQHA is widely used in handheld devices for satellite communication systems as its advantages of good circular polarization, hemi-spherical radiation pattern and easy integration with mobile terminals. There is an increasing need to have one small antenna equipped with portable devices because it is always desirable to have a portable device as compact as possible. In this case study, one technique that can be used to effectively reduce the size of the PQHA will be demonstrated.

One novel kind of PQHA called the Meandered Variable Pitch Angle Printed Quadrifilar Helix Antenna (MVPQHA) is proposed in one of our recent works [6, 7]. Using the meander line to design compact PQHA has been reported in several studies and some of them have been presented in Chapter 2; however, in our study it is found that better bandwidth can be achieved when the PQHA is designed in such a way that the linear segments exhibit variable pitch angles, as depicted in Figure 6.11, where  $L_{es}$  represents the length of each meandered element,  $w$  is the width of the microstrip line and  $p$  is the initial pitch angle. The individual meandered element is shown in Figure 6.12. The reason why this geometry is chosen is that it can provide less mutual coupling between adjacent elements compared to the conventional one that is shown in Figure 2.26 of Chapter 2.

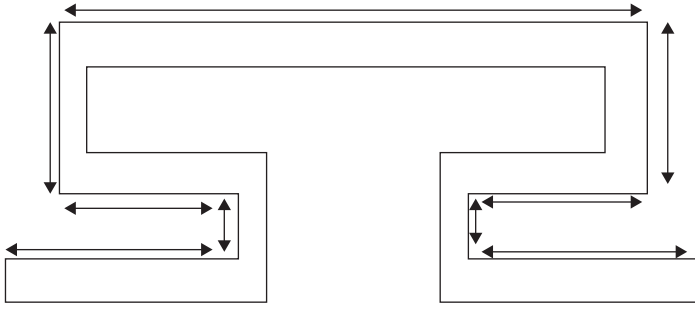
The axial length of the MVPQHA is calculated by dividing it into elements and making the sum of its corresponding axial length, as demonstrated in Figure 6.13.

Then, the axial length can be calculated by using the following formula:

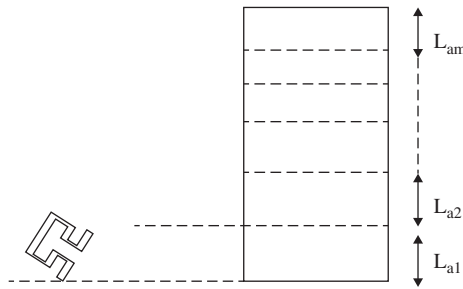
$$L_{Axial}(MVPQHA) = \sum_{i=1}^m L_{ai} \quad (6.8)$$



**Figure 6.11** The meandered line with variable pitch angle for PQHA design [6] Reproduced with permission of © 2009 IEEE



**Figure 6.12** The individual meandered element for the MVPQHA



**Figure 6.13** Calculation of the axial length of the MVPQHA

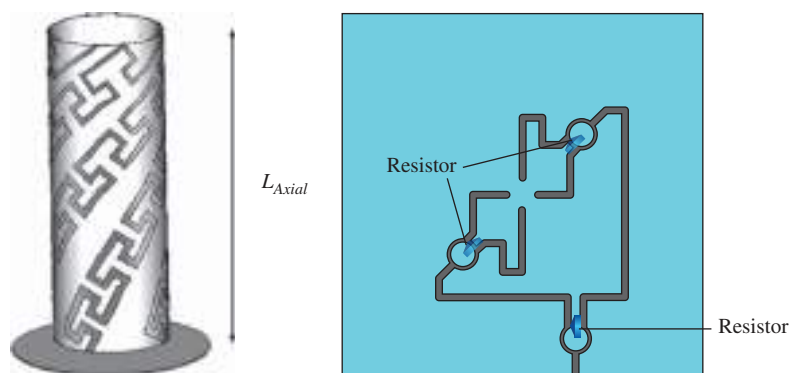
where  $m$  is the number of the elements and  $L_{ai}$  corresponds to the axial length of each individual element. The pitch angle is represented by  $p$  and the step size for varying pitch angle is  $dp$ . This means that the pitch angle of each square element differs from the other by  $dp$  degrees. If the pitch angle varies equally around the central angle, the axial length of each element can be calculated by using the following formula:

$$L_{ai} = \frac{1}{m} (\tan(p_i) \times N \times 2\pi r) \quad (6.9)$$

where  $r$  is the radius of the antenna and  $N$  is the number of turns. Therefore, the total axial length can be derived as:

$$L_{Axial}(MVPQHA) = \frac{1}{m} \sum_{i=1}^m \tan(p + dp \times i) \times N \times 2\pi r \quad (6.10)$$

The 3D model of the proposed MVPQHA is presented in Figure 6.14. As the four helices of the PQHA need to be fed with equal amplitudes with phases of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$ , a feed network that consists of three Wilkinson Power dividers has been designed and is also shown in Figure 6.14. A  $50 \Omega$  feed is given to the input of the first Wilkinson power dividers which divides the power equally into two arms. The equal division of the power can be achieved by having two arms of length equal to a quarter of a guided wavelength and with an impedance of  $70.7 \Omega$ . The end of the two arms is connected to the microstrip lines with impedance of  $100 \Omega$ . The two arms from the Wilkinson power divider in turn act



**Figure 6.14** The 3D model of the MVPQHA and its feeding network

as the feed for another two Wilkinson power dividers. A difference in length between two feed lines is maintained in order to have a phase difference of  $180^\circ$ . The output of the two Wilkinson power dividers has the electrical length difference of  $90^\circ$  each. In this way phase differences of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$  and  $270^\circ$  are created with equal amplitude.

A photo of the fabricated MVPQHA with feed network is given in Figure 6.15. The feed circuit is fabricated on an FR4 substrate ( $\epsilon_r = 4.4$ ) of thickness 1.6 mm. The MVPQHA is printed on a thin film with a relative permittivity of 2.2. The width ( $w$ ) of the lowest meandered element of helix is 1 mm and it is reduced in successive meandered elements by 0.1 mm. The height of the antenna ( $L_{Axial}$ ) is 38.4 mm and the radius of the cylinder is 7 mm. Each helix is divided into a number of meandered elements with decreasing width.

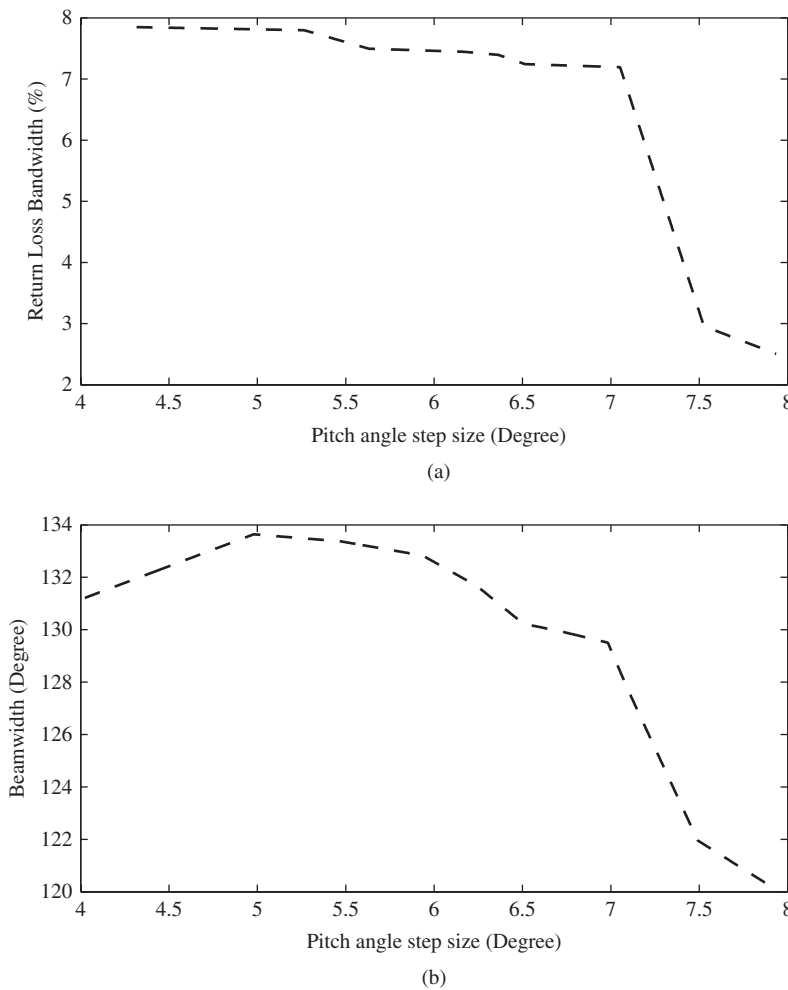


**Figure 6.15** Photo of the fabricated MVPQHA with feed network [6] Reproduced with permission of © 2009 IEEE

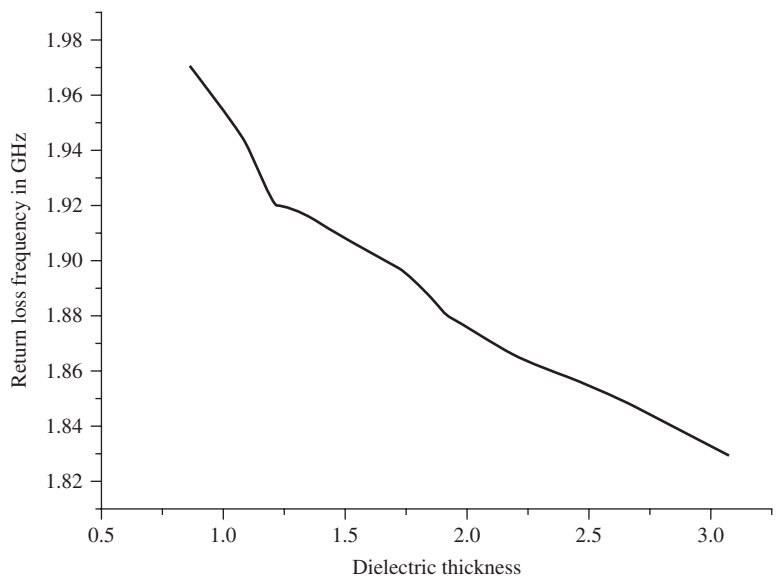
The length of each meandered element ( $L_{es}$ ) is 25 mm. The pitch angle varies from  $62.5^\circ$  to  $37.5^\circ$  with a step size decrease of  $5^\circ$ .

The optimum step size of the pitch angle is chosen by doing numerical simulations in the CST microwave studio [3]. The effect of pitch angle on the performance of MVPQHA is studied by varying the pitch angle from  $4^\circ$  to  $8^\circ$  while keeping the mid-point pitch angle about  $50^\circ$  and then studying the antenna return loss as well as the half power beam width. It is found that the maximum return loss and half power beam width coverage are obtained when the pitch angle step size is  $5^\circ$ , as shown in Figure 6.16.

The effect of loading the dielectric is also studied and the simulation result is presented in Figure 6.17. Increasing the thickness of the dielectric means decreasing the resonant frequency. The antenna resonates at about 2 GHz when printed on a dielectric of  $50\ \mu\text{m}$  thick.



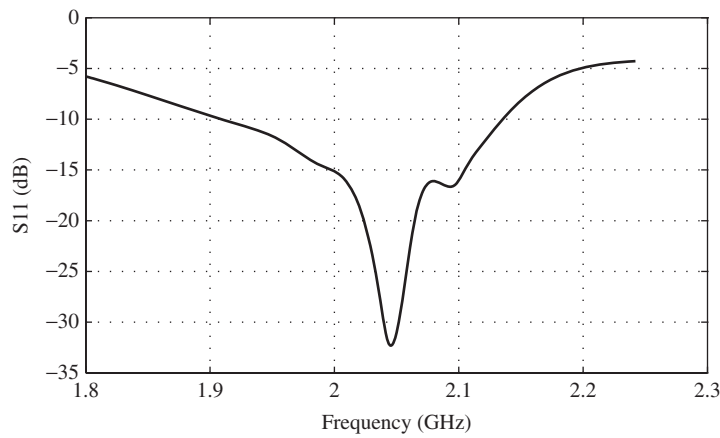
**Figure 6.16** The influence of the pitch angle step size to antenna bandwidth and half power beamwidth



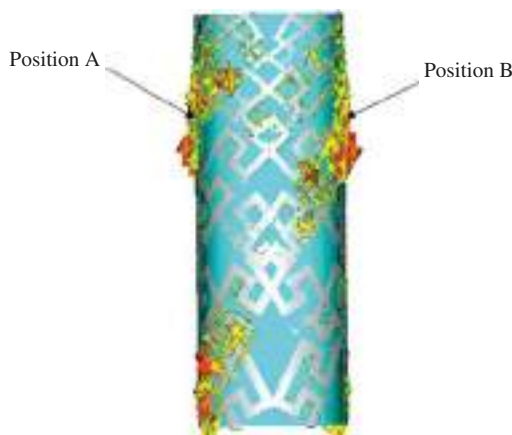
**Figure 6.17** The effect of the dielectric thickness to the resonant frequency of the MVPQHA

When the dielectric substrate is made more solid with the thickness of 3 mm, the resonant frequency of the antenna shifts to 1.82 GHz. It is also observed that with the increasing thickness of the dielectric, the impedance matching becomes worse. Therefore, in this design the thickness of the substrate is chosen to be 2 mm, in order to have a small antenna while having a good impedance bandwidth.

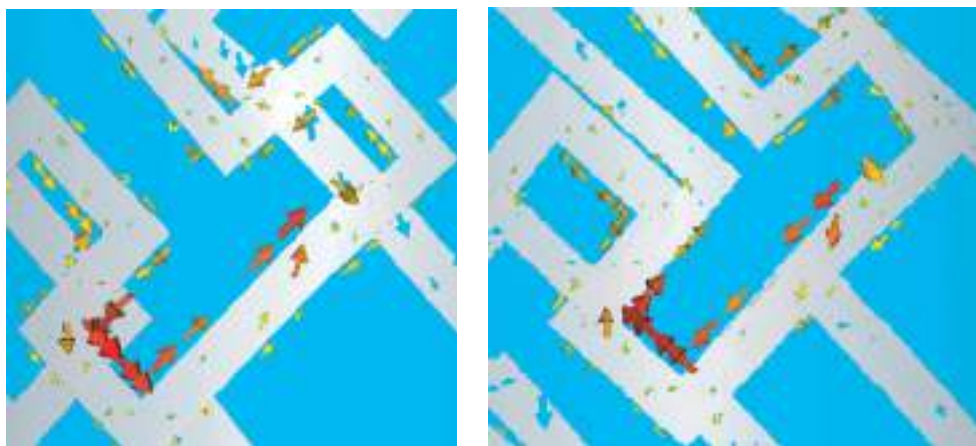
Figure 6.18 shows the measured return loss of the fabricated antenna prototype. The measurement results show that this antenna has 10-dB return loss bandwidth from 1.9–2.1 GHz



**Figure 6.18** Measured S11 of the proposed MVPQHA



**Figure 6.19** Surface current of the MVPQHA at 2 GHz

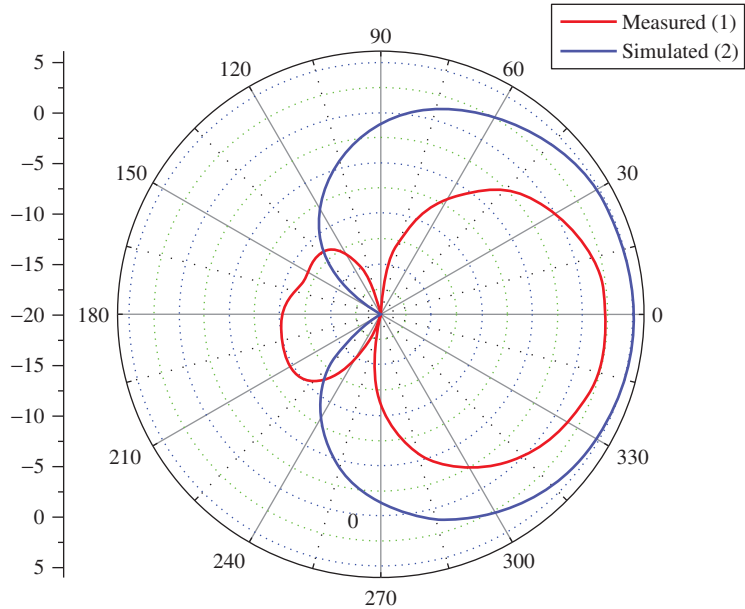


**Figure 6.20** Current flow in position A and position B

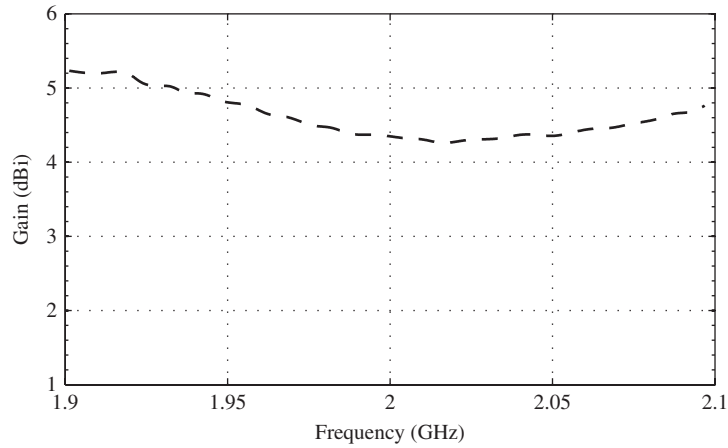
with a central frequency of 2.05 GHz, which covers the required frequency band for earth to space communications (1.98–2.01 GHz).

Figure 6.19 shows the simulated surface current behavior of the MVPQHA antenna. The surface current is the combination of tangential and normal current. The two maximum current positions have been named A and B. At these two points, the direction of current is  $180^\circ$  out of phase at any given instance of time. Figure 6.20 provides a detailed view of the current flow in position A and position B.

The measured far-field radiation pattern of the MVPQHA at 2 GHz is shown in Figure 6.21 and the simulated gain is given in Figure 6.22. The MVPQHA is found to have small front-to-back ratio. The simulation is made without taking into consideration of the feed circuit. The measured results show that the gain of this antenna is about 2 dB lower than



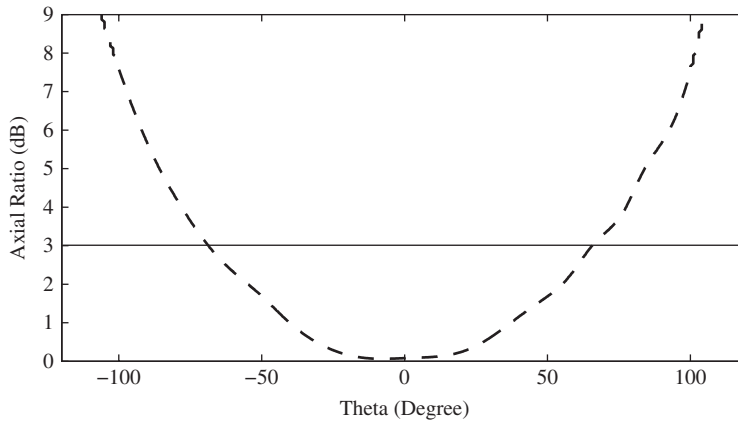
**Figure 6.21** Measured and simulated radiation pattern of the MVPQHA at 2 GHz [6] Reproduced with permission of © 2009 IEEE



**Figure 6.22** Simulated gain of the MVPQHA without the feed network

the simulated one, which is mainly due to the fact that the feed circuit is printed on a lossy FR4 dielectric.

Figure 6.23 shows the simulated axial ratio performance of the MVPQHA at 2 GHz. The simulation result shows that the antenna has a good circular polarization at 2 GHz over the vertical theta ( $\theta$ ) angle of about  $140^\circ$ .



**Figure 6.23** Simulated axial ratio of the MVPQHA at 2 GHz for different theta angles

This case study presents a small PQHA with variable pitch angle and its feed network. The design techniques shown in this case study can also be applied to the design of other types of small PQHAs such as the ones given in Section 2.4, Chapter 2.

## 6.4 Printed Broadband CP Rectangular Bi-Loop Antenna for RFID Readers

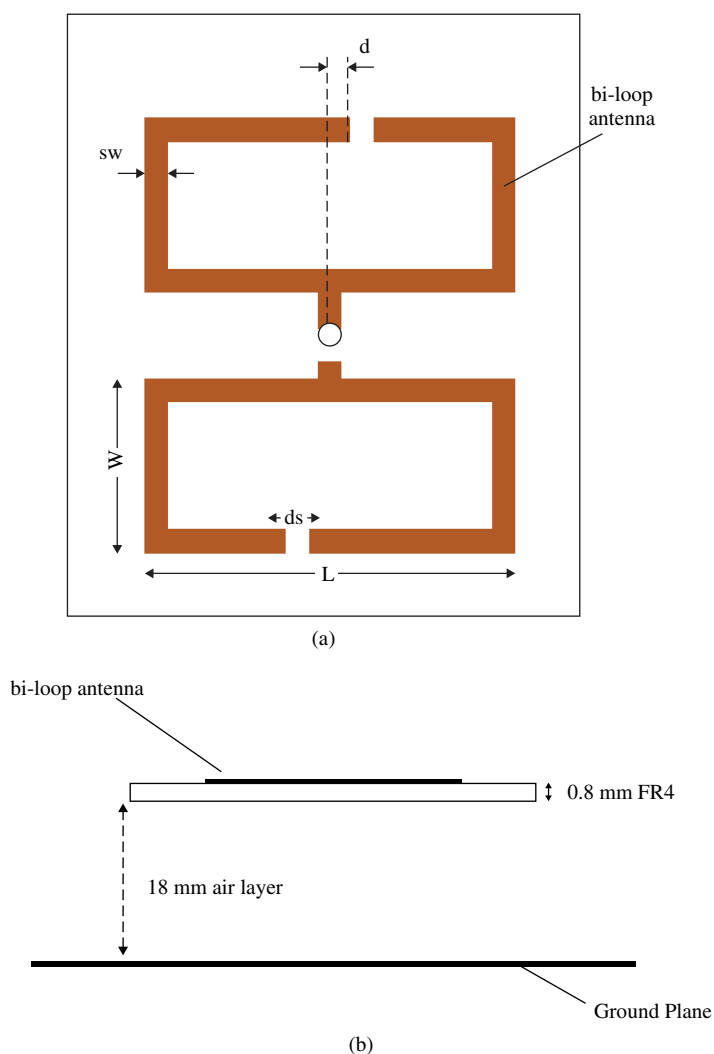
RFID is used for the identification of different objects and has gained wide application in recent years. For the design of the antenna for RFID readers, it is always desirable to have one antenna with broad bandwidth, moderate gain and more importantly, circular polarization. Since most of the antennas for RFID tags are linearly polarized, having a CP antenna equipped with the RFID reader can enable the system to detect the objects regardless of the orientation of the RFID tags.

In this case study, one printed broadband CP rectangular bi-loop antenna for a RFID reader with a resonant frequency at 2.4 GHz is demonstrated. This type of antenna has been investigated by [8, 9] but here in this study, instead of using a cylindrical wire, a printed rectangular loop is employed to simplify the fabrication process and improve the fabrication accuracy as well. Figure 6.24 shows the top and side view of the printed CP broadband rectangular bi-loop antenna.

The parameters of the bi-loop antenna are indicated in Figure 6.24, where  $L$  and  $W$  represent the length and width of the loop, respectively. Here,  $s_w$  represents the width of the microstrip line,  $ds$  is the length of the slot cut on the loop and  $d$  indicates the location of the slot. The bi-loop antenna can be regarded as two loop antennas fed by a short dipole and the circumference of each loop needs to be approximately one wavelength long at the resonant frequency:

$$2 \times (L + W) = \lambda_0 \quad (6.11)$$

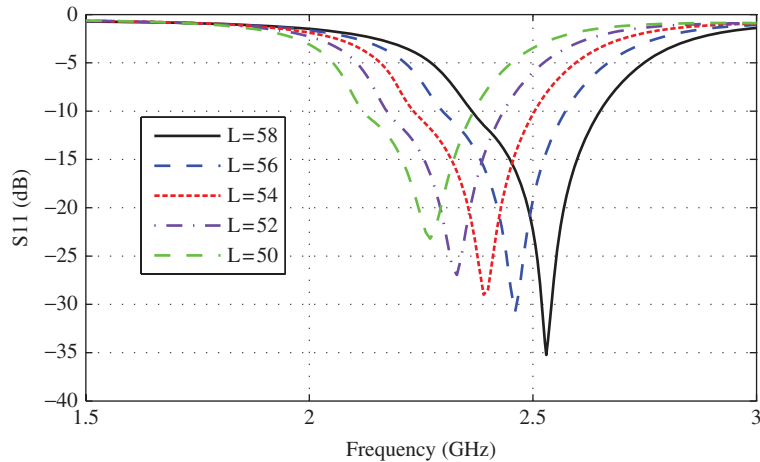
To obtain the CP radiation, the loop antenna needs to have a travelling wave current distribution with constant amplitude and linear phase changing [8]. One approach to reach this



**Figure 6.24** Top and side view of the printed broadband circular polarization rectangular bi-loop antenna

condition is to load a reactance on the loop antenna [10]. The reactance can be introduced by cutting a small gap on the loop antenna, as depicted in Figure 6.24. As can be seen from Figure 6.24, there is a ground plane below the antenna element. The ground plane works as the reflector for the antenna to increase its directivity and reduce the backside radiation.

Parametrical studies have been conducted to demonstrate how the radiation performance of the antenna can be optimized by varying the relevant parameters. Figure 6.25 presents the simulated  $S_{11}$  of the printed bi-loop with different loop length ( $L$ ). As can be observed from these simulation results, the resonant frequency of the antenna can be tuned by changing the circumference of the loop.



**Figure 6.25** The simulated S11 of the bi-loop antenna when varying the value of  $L$ . The rest of parameters for this study are:  $W = 25$ ,  $sw = 3.5$ ,  $d = 3$ ,  $ds = 3.5$  (units are all millimetres)

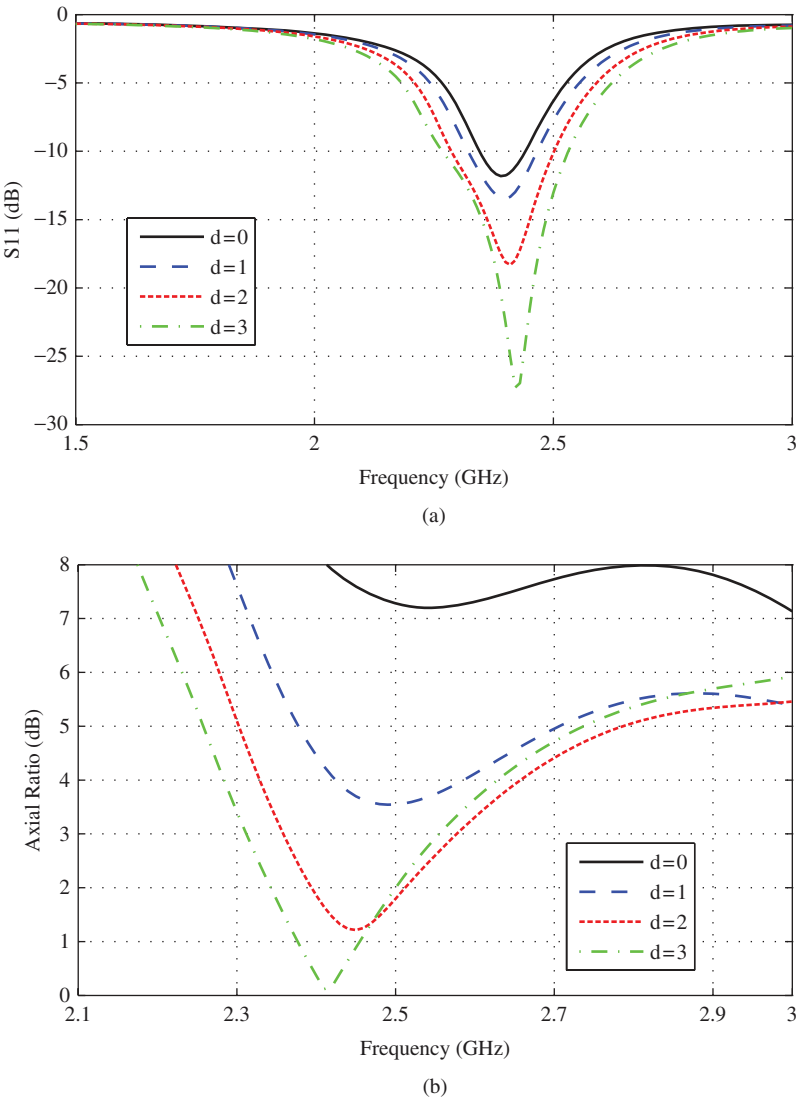
The effects of the gap size and its position on the loop antenna to the AR and impedance bandwidth of the bi-loop antenna have also been studied. Figure 6.26 and Figure 6.27 respectively show the simulated AR and impedance bandwidth of the antenna with the gap in different position and with varied gap size. These parametrical studies show that the size and position of the two gaps, which are rotatively symmetry, can influence both the AR and impedance bandwidth of the antenna; as the result, values of these parameters need to be optimized in order to reach the best overall radiation performance. In this study, the dimensions of the bi-loop antenna are optimized by doing numerical simulations in Ansoft HFSS V14 [11] and the values of the final parameters are given in Table 6.4.

Figure 6.28 shows the current distribution of the bi-loop antenna at 2.4 GHz. The size and position of the two gaps are optimized to let the bi-loops have a travelling wave current distribution with constant amplitude and linear phase changing.

Figure 6.29 shows the top view of the fabricated antenna prototype. This Bi-loop is printed on a 0.8-mm thick FR4 ( $\epsilon_r = 4.4$ ) and is mounted 18-mm above the ground plane. The hole drilled in the middle of the PCB is reserved for adding a semi-rigid cable to feed the antenna.

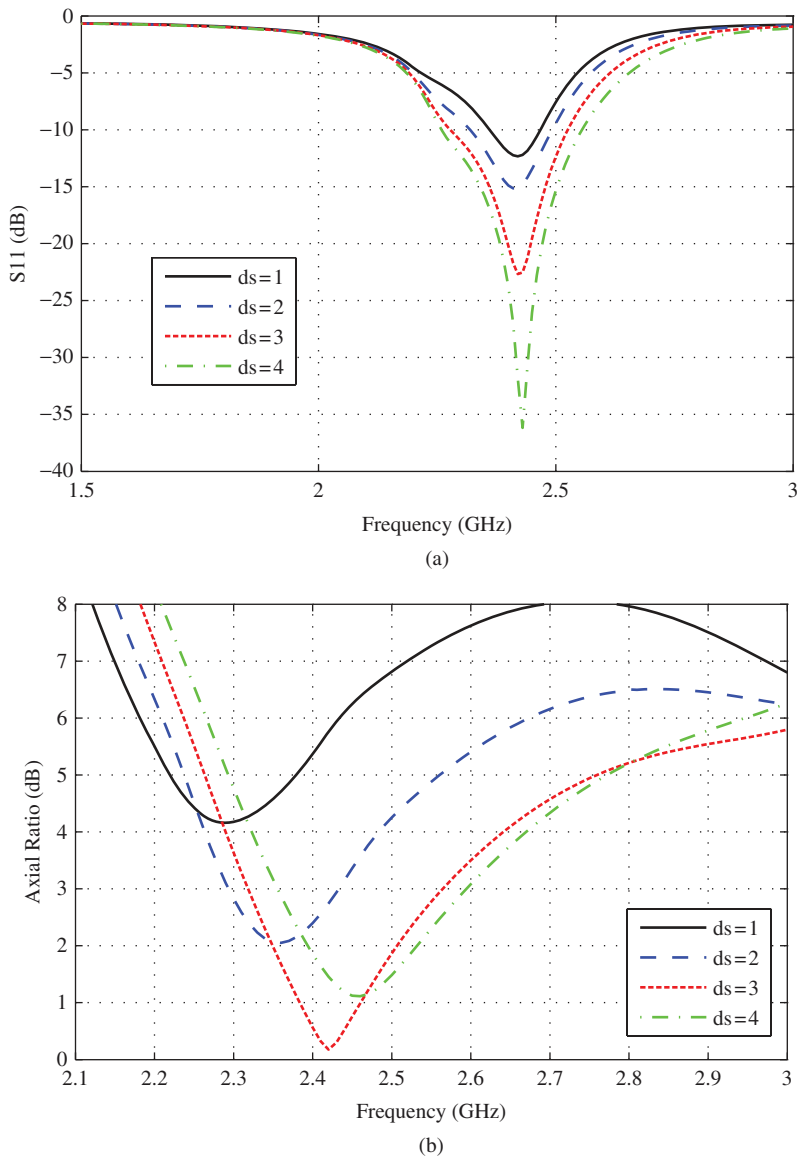
**Table 6.4** Parameters of the planar bi-loop antenna

| Name | Value (mm) |
|------|------------|
| L    | 53         |
| W    | 25         |
| sw   | 3.5        |
| d    | 2.8        |
| ds   | 3.5        |

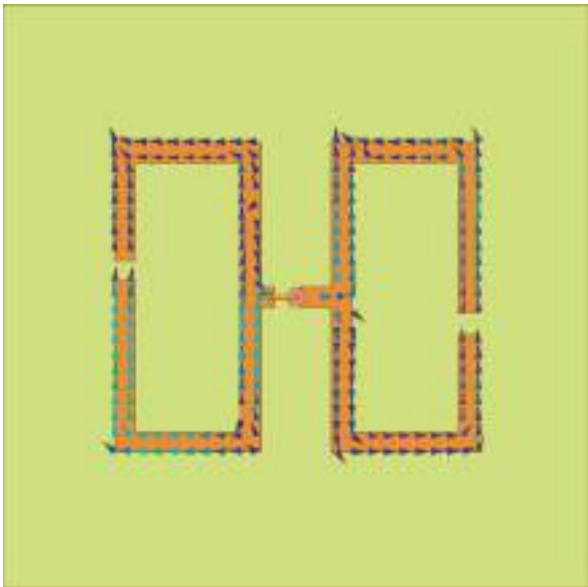


**Figure 6.26** The simulated  $S_{11}$  and AR bandwidth of the bi-loop antenna when varying the position of the gap ( $d$ ). The rest of parameters for this study are:  $L = 53$ ,  $W = 25$ ,  $sw = 3.5$ ,  $ds = 3.5$  (units are all millimetres)

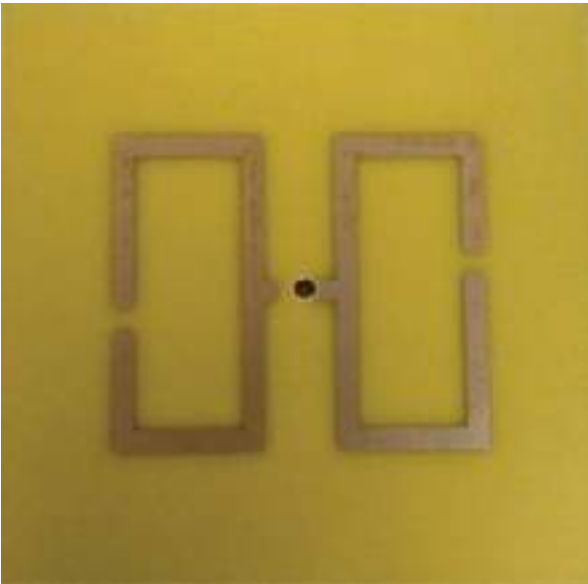
Figure 6.30 shows the simulated and measured return loss of the rectangular bi-loop antenna as well as the simulated axial ratio of the bi-loop antenna in the broadside direction. It can be seen that there is a good agreement between the simulated and measured  $S_{11}$ . The measured result shows that the bi-loop antenna has a 10-dB return loss bandwidth from 2.27–2.53 GHz. Meanwhile, the simulation result suggests that it has a 3-dB AR bandwidth from 2.34–2.6 GHz, which is better than 10% with central frequency of 2.4 GHz.



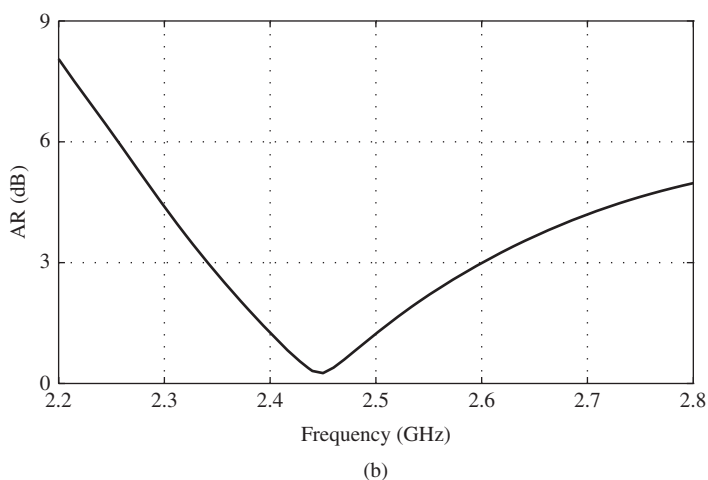
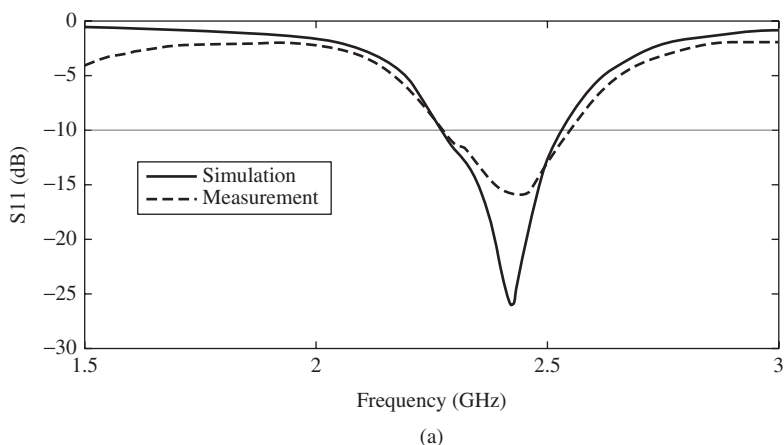
**Figure 6.27** The simulated  $S_{11}$  and AR bandwidth of the bi-loop antenna when varying the size of the gap ( $ds$ ). The rest of parameters for this study are:  $L = 53$ ,  $W = 25$ ,  $sw = 3.5$ ,  $d = 3$  (units are all millimetres)



**Figure 6.28** The current distribution of the bi-loop antenna at 2.4 GHz



**Figure 6.29** Top view of the fabricated bi-loop antenna



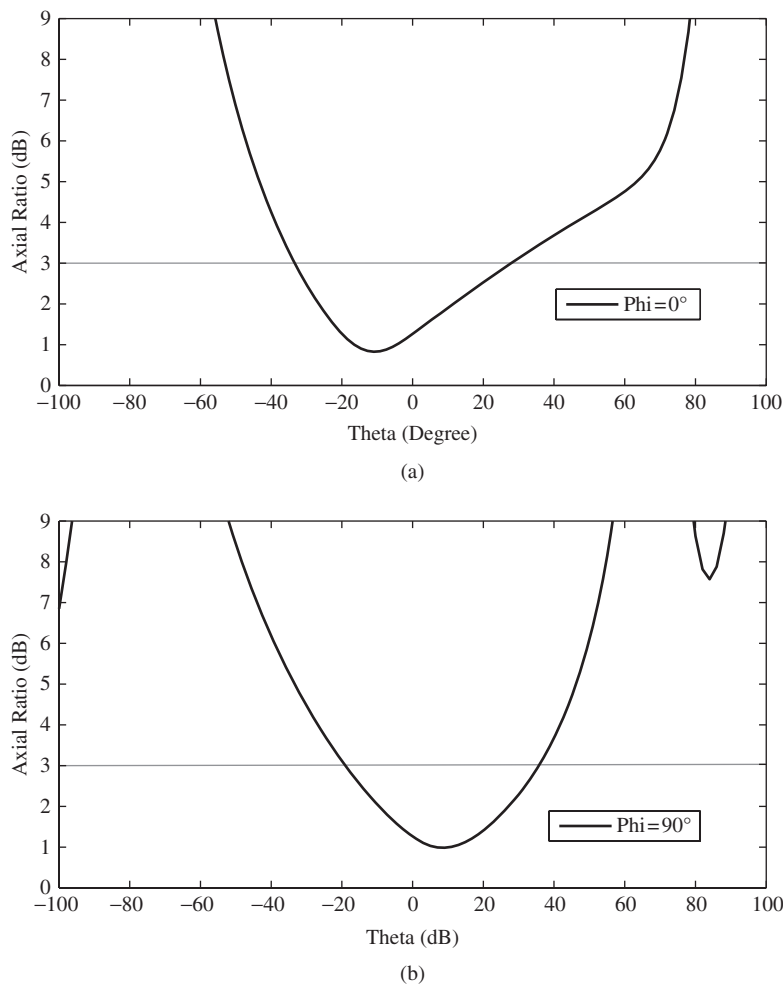
**Figure 6.30** Simulated and measured  $S_{11}$ , simulated axial ratio bandwidth of the rectangular bi-loop antenna

The simulated axial ratio at 2.4 GHz with different theta angle is presented in Figure 6.31. As shown in the simulation results, at the resonant frequency the bi-loop antenna has a 3-dB AR beamwidth over the vertical  $\theta$  angle of about  $60^\circ$ .

Figure 6.32 shows the simulated far-field radiation pattern of the bi-loop antenna at 2.4 GHz. As shown, it has a hemi-spherical radiation pattern and the simulated gain of this antenna is found to be about 9 dBi.

## 6.5 CP Reflectarray for Ka Band Satellite Communications

Ka-band satellite communications can provide promising high-speed links between satellites and mobile terminals such as trains, airplanes, cars and ships. For this communication

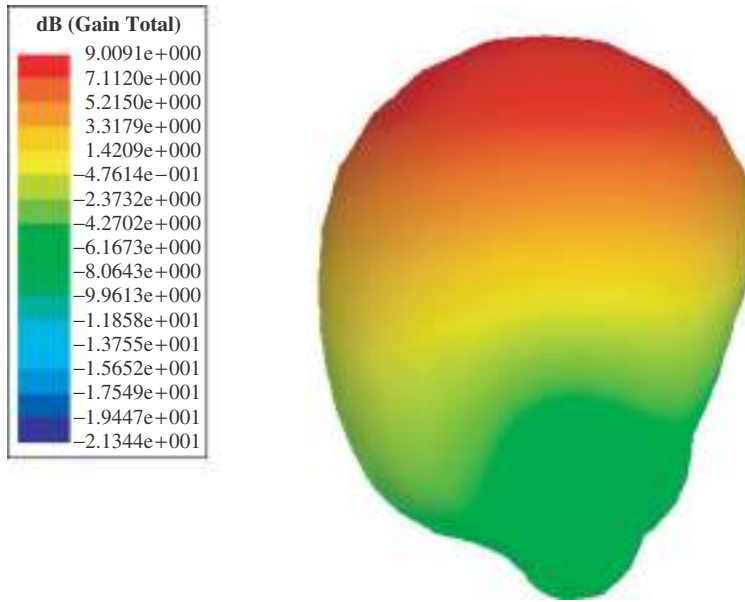


**Figure 6.31** Simulated axial ratio of the bi-loop antenna at 2.4 GHz for different  $\theta$  angles

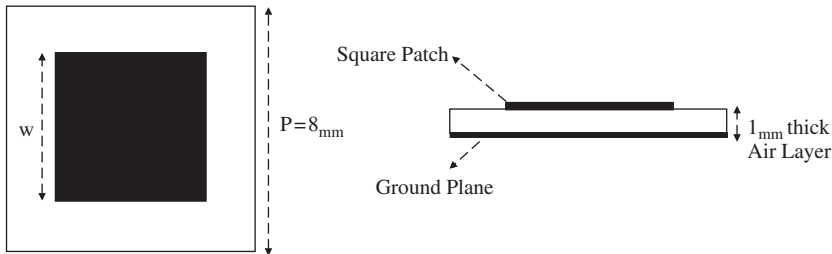
system, the antenna is a key component. Due to the highly dynamic environment for the mobile terminals and high data-rate requirements, the antenna is required to have a low profile, low mass, high gain and electrical beam scanning capability within a wide range of angles.

In this case study, one CP reflectarray using the square patch as the radiating element and one CP horn antenna as the feeding source is presented. The reflectarray consists of 400 patch antenna elements to produce twisted re-radiated field and provide phase compensation for beam focusing. The square patch is chosen in this study as it can provide equal phase delay for the horizontal and vertical components of the E-field decomposed from the CP incident wave:

$$\vec{E}_{cp} = E_x e^{j\Phi} + E_y e^{j(\Phi+90^\circ)} \quad (6.12)$$



**Figure 6.32** Simulated far-field radiation pattern of the bi-loop antenna at 2.4 GHz



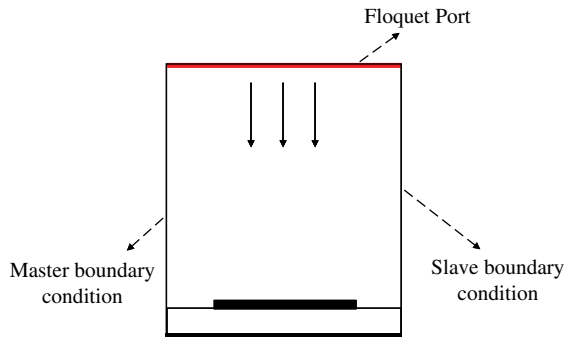
**Figure 6.33** Top and side view of the square patch antenna that is used to build the reflectarray

where  $\vec{E}_{cp}$  represents the E-field of the circularly polarized wave,  $E_x$  is the horizontal component and  $E_y$  is the vertical component of  $\vec{E}_{cp}$ .

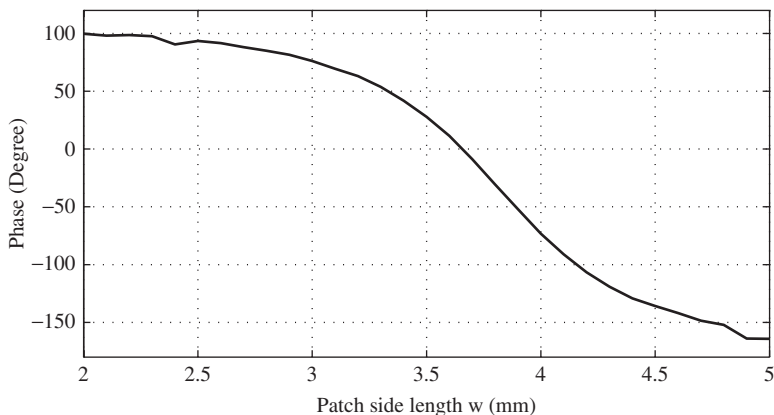
Although it has been reported that the cross polarization of the reflectarray can be suppressed if the linearly polarized reflection coefficients along the X- and Y- axis can have a phase difference of  $180^\circ$  [12], it is shown in this study that employing a simple square patch as the antenna unit cell can also reach a low level of cross polarization.

Figure 6.33 shows the top and side view of the square patch antenna. The square patch is designed on a 1-mm thick air layer and the distance between each antenna unit cell is chosen to be 8 mm, which is  $0.8\lambda_{30\text{ GHz}}$ , where  $\lambda_{30\text{ GHz}}$  represents the free space wavelength at 30 GHz. To resonate at 30 GHz, the side length of the square patch is set to be 3.8 mm.

This antenna unit cell is studied in Ansoft HFSS in an infinitely array condition, as depicted in Figure 6.34. The side length of the patch,  $w$ , are varied to calculate the phase response,



**Figure 6.34** Simulation model of the antenna unit cell in an infinite array condition



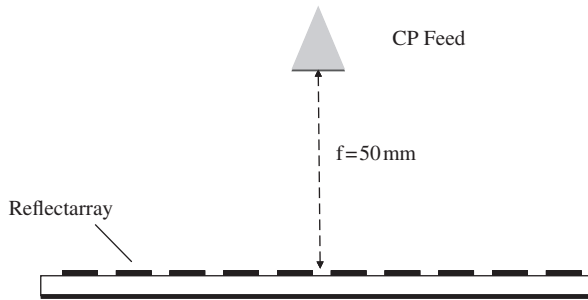
**Figure 6.35** Phase response of the square patch with varying side lengths

which is given in Figure 6.35. This patch antenna can provide a phase delay up to  $270^\circ$  when the side length of the square patch varies from 2–5 mm. Theoretically it is required to have a phase delay range as large as possible in order to provide larger operation bandwidth (e.g. 1 dB gain bandwidth). For the purpose of demonstration, in this case study the square patch is used to keep the reflectarray as simple as possible.

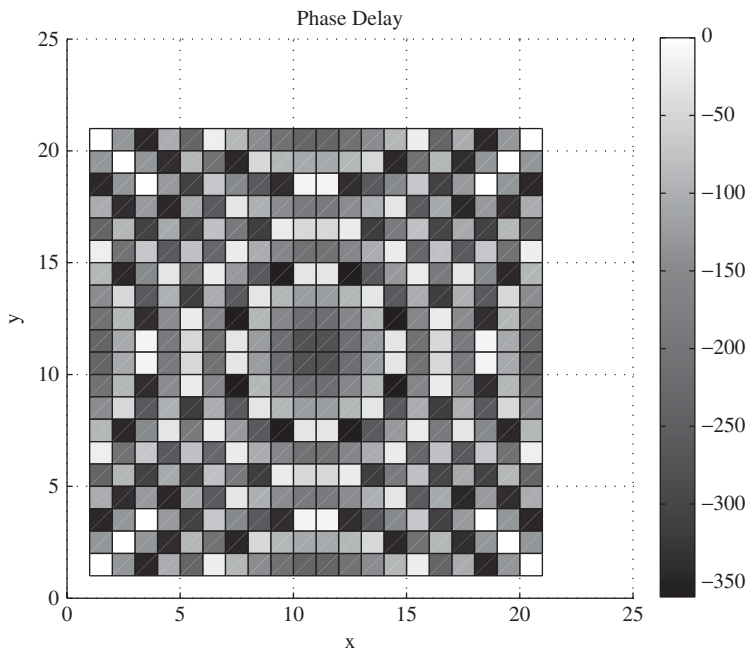
The design of the CP reflectarray starts with the calculation of the required phase for each radiating element based on the formula given in [13]:

$$\Phi_R(i) = k_0(d_i - (x_i \cos \varphi_b + y_i \sin \varphi_b) \sin \theta_b) \quad (6.13)$$

where  $\Phi_R(i)$  is the phase of the reflection coefficient of the antenna element  $i$ ,  $k_0$  is the propagation constant in vacuum,  $(x_i, y_i)$  are the coordinates of element  $i$ ,  $d_i$  is the distance from the phase centre of the feed to the antenna unit cell and  $(\theta_b, \varphi_b)$  is the expected scan angle of the beam. To design a reflectarray radiating at broadside ( $\theta_b = 0$ ,  $\varphi_b = 0$ ), first we choose



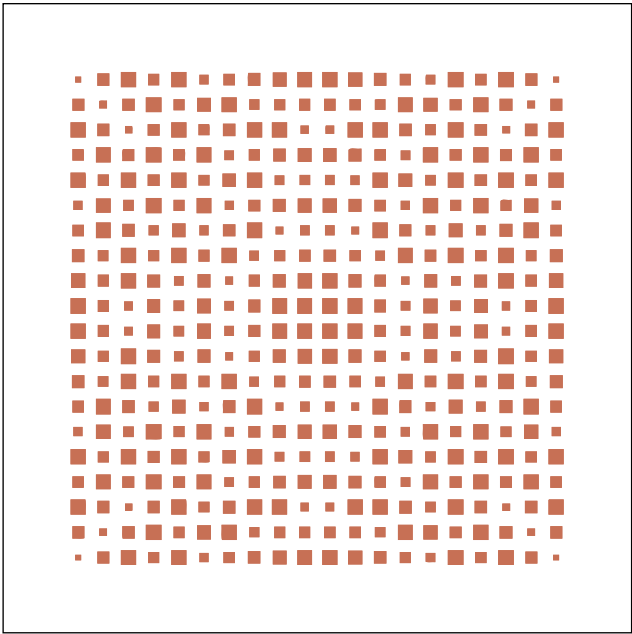
**Figure 6.36** Configuration of the CP reflectarray with a central CP feed



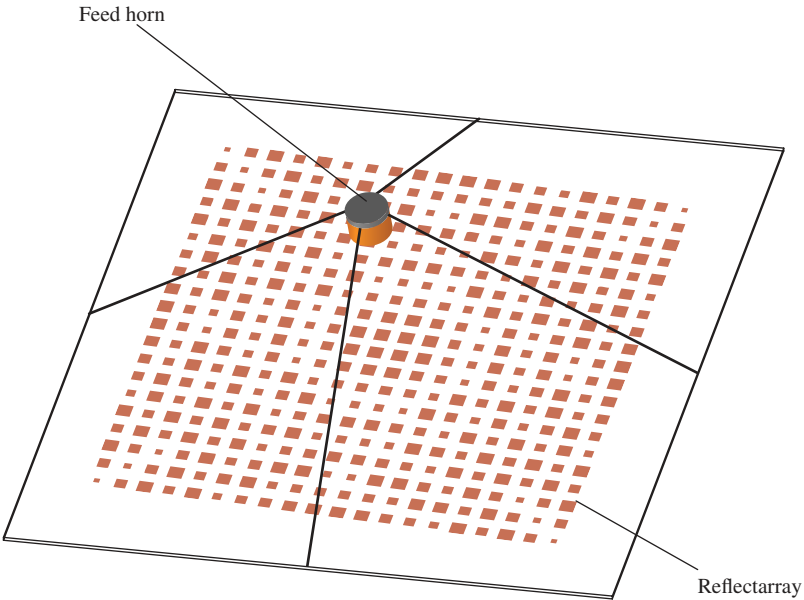
**Figure 6.37** Calculated phase distribution for each unit cell for broadside radiation

a square lattice with  $20 \times 20$  elements and a central CP feed, which is a circular waveguide with a radius of 7.1 mm, at a height of 50 mm. The configuration of the reflectarray is shown in Figure 6.36. Using equation 6.13, the required phase delay for each unit cell is calculated and plotted in Figure 6.37.

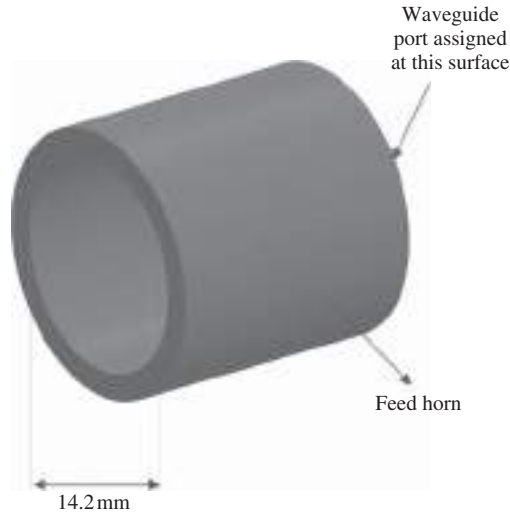
Figure 6.38 shows the layout of the reflectarray with 400 square patches. The side length of each square patch is decided based on the calculated phase distribution (Figure 6.37) and the phase response of the square patch (Figure 6.35). Figure 6.39 shows the 3D model of the reflectarray with a CP circular feed.



**Figure 6.38** The layout of the reflectarray for broadside radiation



**Figure 6.39** The 3D model of the CP reflectarray with a circular feed



**Figure 6.40** Structure of the feed horn used in the CP reflectarray design

The feed of the reflectarray is chosen to be a circular waveguide with diameter of 14.2 mm and length of 3 mm. The radius of the circular waveguide is decided by calculating its cut-off frequency:

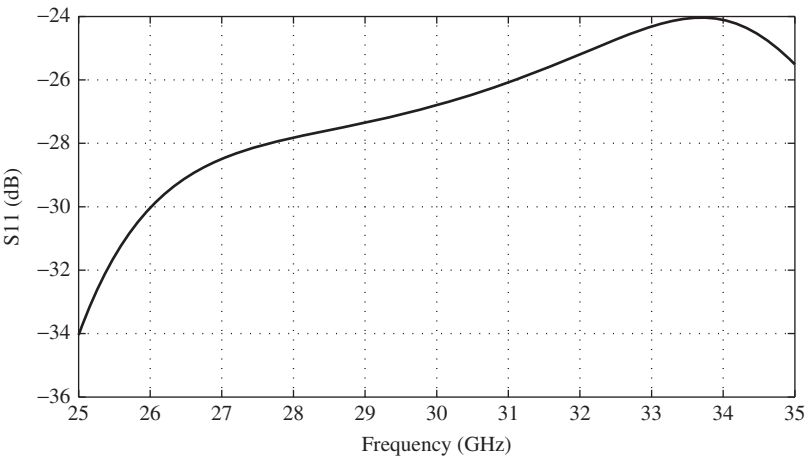
$$f_c = \frac{1.84 \cdot c}{2\pi r} \quad (6.14)$$

where  $r$  is the radius of the circular waveguide and  $c$  is the speed of light. The feed is designed and simulated in HFSS V14. Figure 6.40 shows the structure of the feed horn and Figure 6.41 shows the simulated S11 and radiation pattern at 30 GHz of this feed.

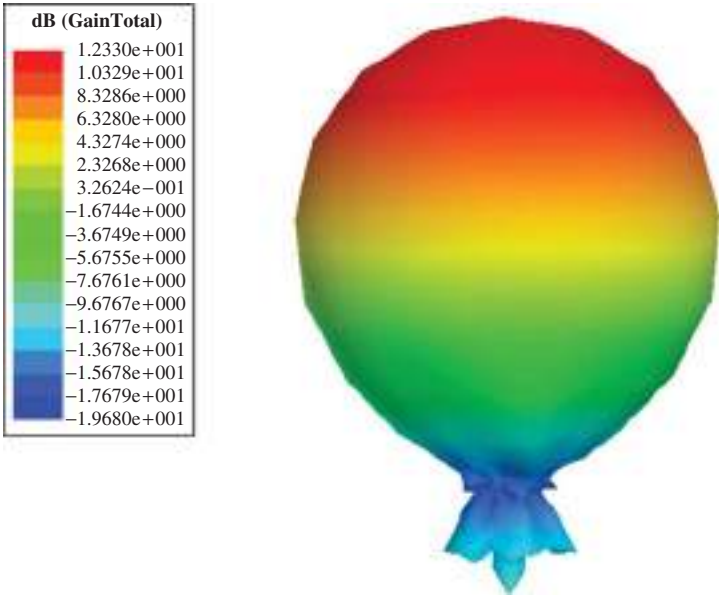
The simulated radiation patterns of this CP reflectarray at 30 GHz are shown in Figure 6.42. The simulation results show that this reflectarray has RHCP due to the use of a RHCP feed. The reflectarray has a directivity of 26 dBi and the cross-polarization is always 20 dB less than co-polarization within the main beam. Moreover, at the resonant frequency, the axial ratio of the reflectarray is less than 0.5 dB.

Figure 6.43 presents the simulated peak gain of this CP reflectarray against frequency. It can be seen that this reflectarray has 1-dB gain bandwidth from 29–30.5 GHz, which represents 5% with central frequency of 30 GHz.

Using the same configuration, it is also possible to let the CP reflectarray radiate toward other directions. The only issue is that this reflectarray cannot steer the beam to wide ranges due to the increase of surface wave propagation at large scan angle. For more information about how to achieve large scan angles, readers can refer to [14]. With a small scan angle, for example  $10^\circ$ , the presented reflectarray can still operate properly. Figure 6.44 shows the calculated phase distribution for the reflectarray has the main beam at  $10^\circ$  elevation angle using the equation 6.13. The resulting layout of the reflectarray is also presented in this figure.

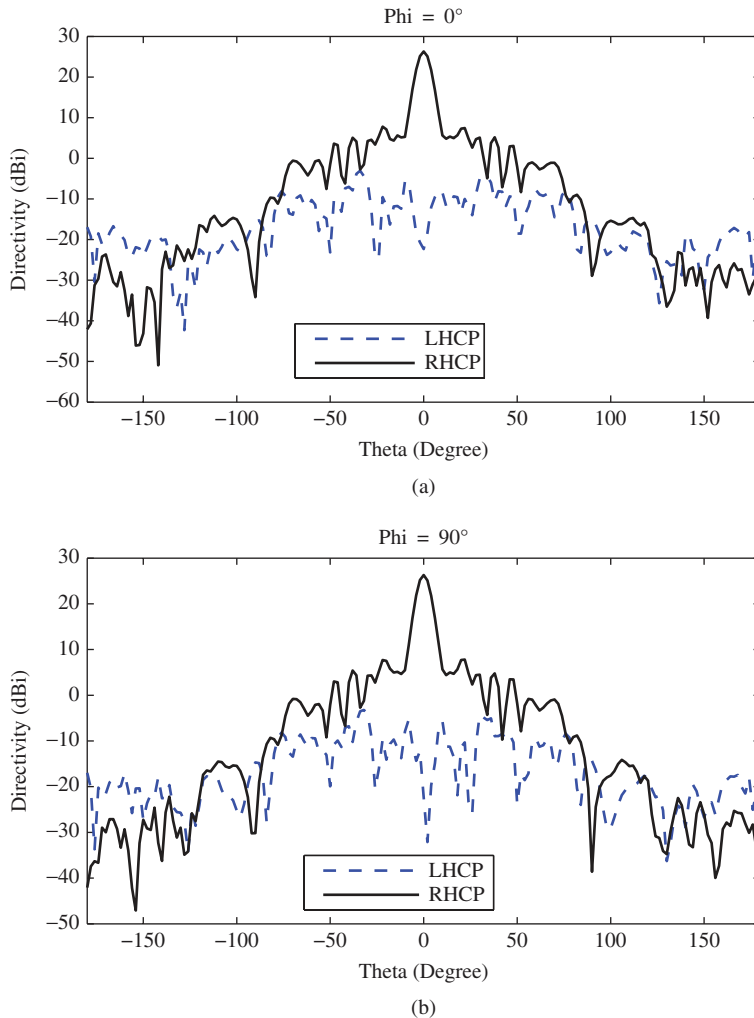


(a)



(b)

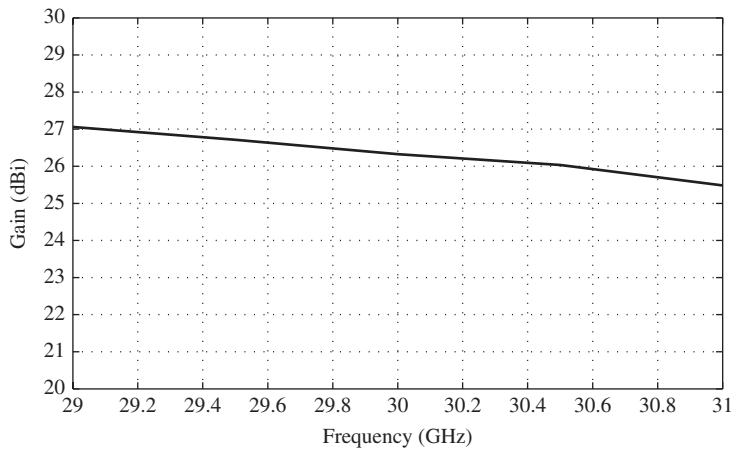
**Figure 6.41** Simulated return loss and radiation pattern at 30 GHz of the CP feed for the reflectarray



**Figure 6.42** Simulated radiation patterns of this CP reflectarray at 30 GHz

The simulated directivity of this CP reflectarray is presented in Figure 6.45. As can be seen from the simulation result, this reflectarray is able to have the main beam at  $10^\circ$  from the elevation plane with a slightly increased cross-polarization level, which is still at least 20 dB lower than the co-polarization in the beam peak. Simulation results when this CP reflectarray is designed to operate at larger scan angles are not provided because it is found that when the scan angle is increased to more than  $20^\circ$ , the reflectarray fails to operate properly.

In brief, in this case study we demonstrated the methodology to design a CP reflectarray with a simple configuration. As can be seen from this design example, the reflectarray can achieve good CP performance with high gain and low cost.



**Figure 6.43** Simulated peak gain of the CP reflectarray against frequency

## 6.6 Circularly Polarized Logarithmic Spiral Antenna with a Wideband Balun

### 6.6.1 The Logarithmic Spiral Antenna

The logarithmic spiral antenna belongs to the class of frequency independent antenna and it has a stable radiation pattern over the operating frequency band, which makes it attractive for wideband applications needing a CP antenna. Typically, the fractional bandwidth of the spiral antenna can reach 5:1 with a gain around 5 dBi. The input impedance of the logarithmic spiral antenna is about  $188 \Omega$ ; as a result, a wideband balun with impedance transformer is required to feed the antenna.

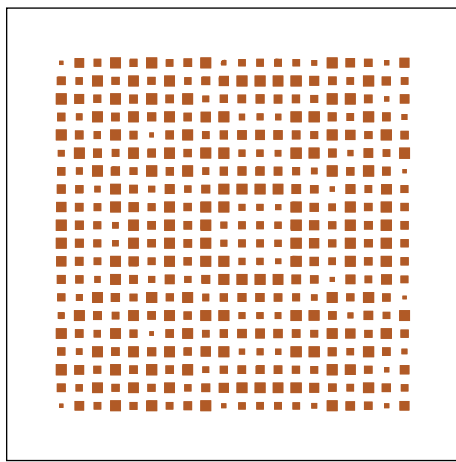
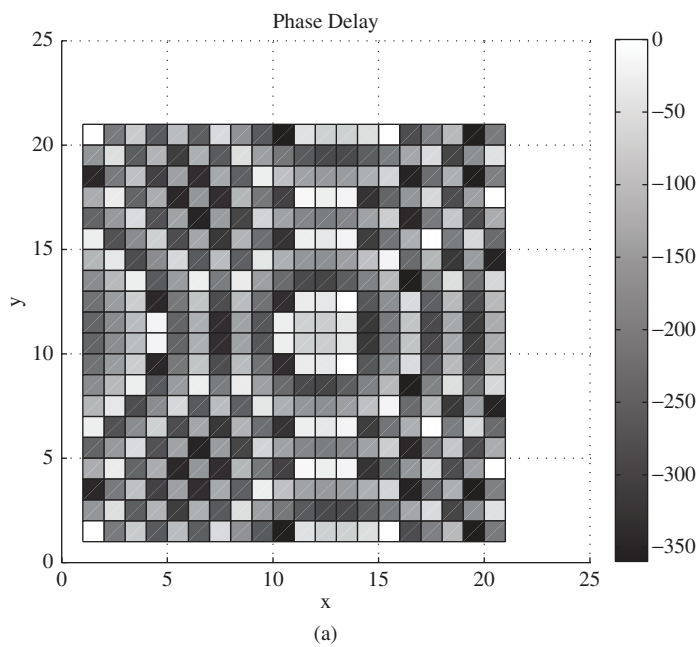
The radiation pattern of the logarithmic spiral antenna is bi-directional and on different side of the antenna it has different polarization, for example, RHCP or LHCP. The radiation pattern of the spiral antenna is typically broadside and with a peak radiation direction perpendicular to the plane of the spiral. Generally speaking, its HPBW is approximately  $70-90^\circ$ .

Figure 6.46 shows the layout of the logarithmic spiral antenna. The spiral curve can be defined by the following formula [15]:

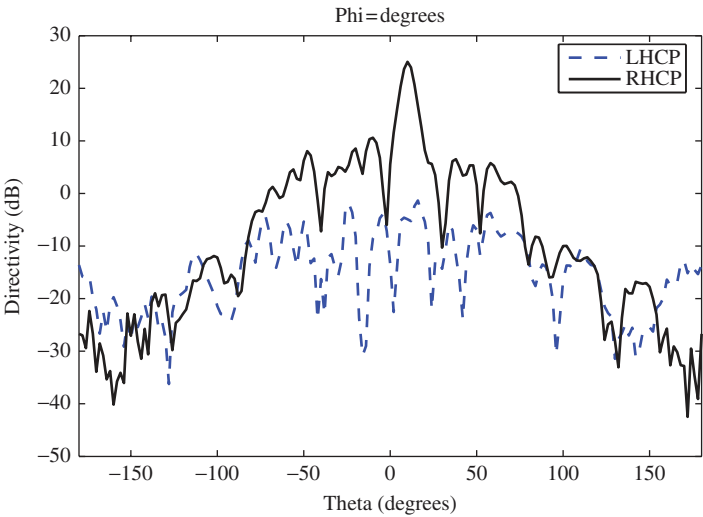
$$r = \frac{D_i}{2} \times \exp \left[ \frac{\ln \left( \frac{D_0}{D_i} \times \theta \right)}{2\pi N} \right] \quad (6.15)$$

where  $D_0$  is the diameter of the outer spiral,  $D_i$  is the diameter of the inner spiral,  $N$  is the number of turns,  $\theta$  and  $r$  is the angle and radius in the polar coordinate, respectively.

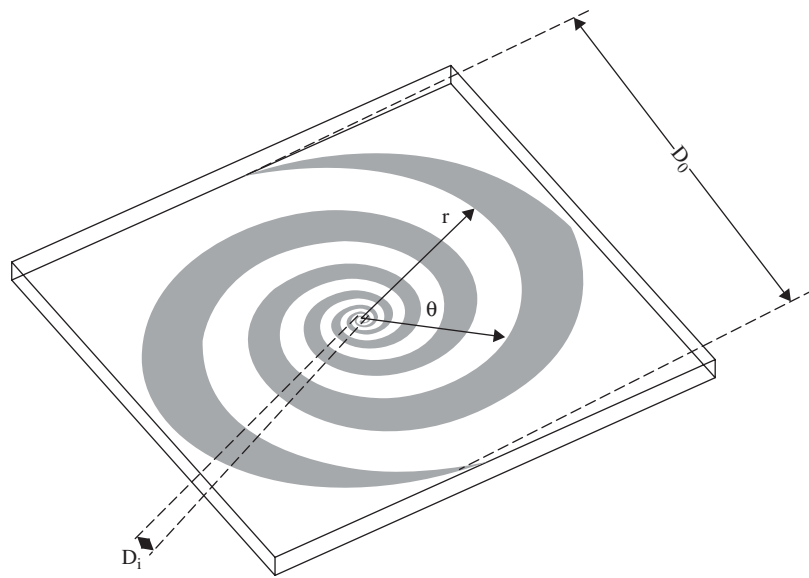
The minimum operating frequency of the logarithmic spiral antenna can be decreased by increasing the outer diameter of the spiral while the maximum operating frequency can be increased by reducing the inner diameter of the spiral. Figure 6.47 shows the simulation



**Figure 6.44** Calculated phase distribution and the layout of the reflectarray to radiate at  $10^\circ$  in the principle plane ( $\varphi = 0^\circ$ )



**Figure 6.45** Simulated co- and cross polarization of the reflectarray with a scan angle of 10°



**Figure 6.46** Layout of the logarithmic spiral antenna



**Figure 6.47** Simulation model of the logarithmic spiral antenna in Ansoft HFSS

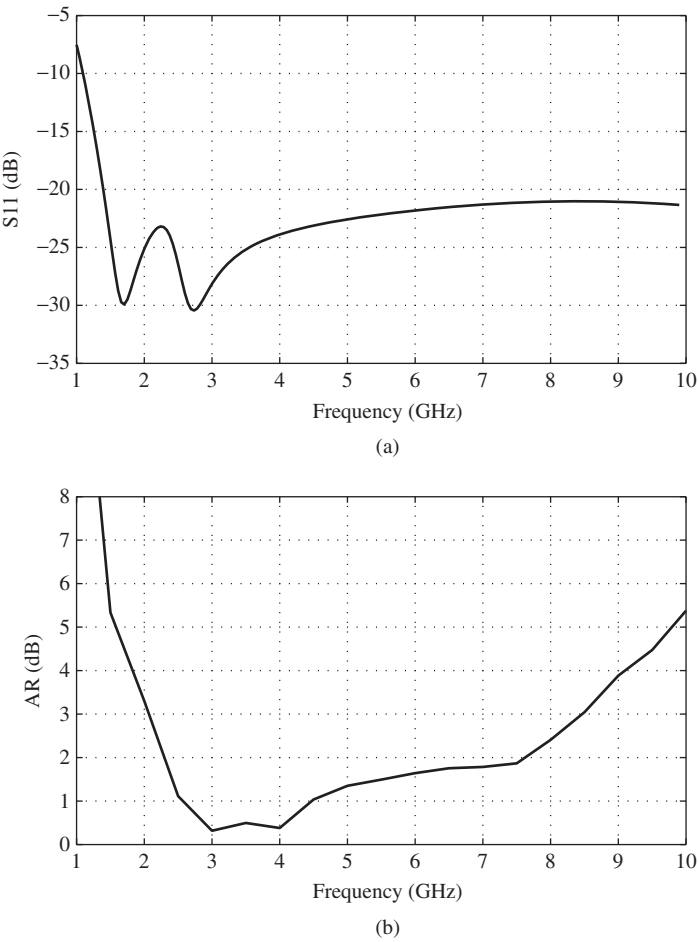
model of the logarithmic spiral antenna in Ansoft HFSS [11]. A lumped port is assigned as the excitation for the logarithmic spiral antenna. To ensure that the antenna can operate from 1–10 GHz, the inner diameter of the spiral is chosen to be 12 mm and the outer diameter of the spiral is selected to be 118 mm. The number of turns is decided to be 1.25, to give proper spacing between each turn and thus give a good AR bandwidth. At this stage, we consider the use of ideal excitation source with input impedance of  $188\ \Omega$  and there is no balun used in this simulation model.

Figure 6.48 shows the simulated return loss and axial ratio of the logarithmic spiral antenna. It can be seen that the logarithmic spiral antenna has both a wide 10-dB return loss bandwidth and a wide 3-dB AR bandwidth. With the configuration shown in Figure 6.47, the logarithmic spiral antenna radiates RHCP waves in the upper plane.

As mentioned, the inner and outer diameter of the spiral is related to the maximum and minimum operating frequency of the logarithmic spiral antenna. To demonstrate this, Figure 6.49 presents the simulated  $S_{11}$  of this logarithmic spiral antenna with varied diameter of the outer spiral. It can be seen that with a larger size of outer diameter, the logarithmic spiral antenna can be made to resonate at lower frequencies. Similarly, the maximum operating frequency can be adjusted by changing the diameter of the inner spiral.

### 6.6.2 Design of a Wideband Balun

In the antenna simulation model presented in Figure 6.47, it is assumed that an ideal excitation source is used. However, in real applications a balun is required to feed the logarithmic spiral antenna. Moreover, this balun needs to transform the input impedance of the spiral from  $188\ \Omega$  to  $50\ \Omega$ . A microstrip-to-parallel strip balun is proposed for the application of

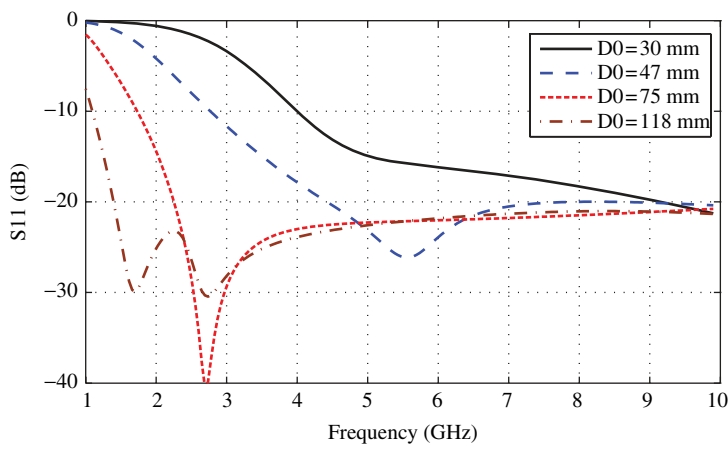


**Figure 6.48** Simulated S11 and axial ratio of the logarithmic spiral antenna

spiral antenna feeding by [16]. In this study, a similar balun is designed to feed the logarithmic spiral antenna.

Figure 6.50 shows the top band back view of the designed balun. The balun is designed by using the Roger/Duriod 5880 substrate with  $\epsilon_r = 2.2$  and thickness of 1.27 mm. The exponential taper is employed to do the impedance transform from 188  $\Omega$  to 50  $\Omega$ . The shape of the ground plane is also modified with the aid of the numerical simulation to transform the microstrip line to a parallel strip line.

Based on the optimized values provided in [16], the tapering of the microstrip lines on both sides of the substrate is divided into several sections, such as the one demonstrated in Figure 6.51. To simplify the design, instead of having a smoothing transition between each section as the one presented in [16], in this study, stepped transition is employed between each section. The dimensions for each section of both top microstrip line and bottom ground plane are given in Table 6.5.



**Figure 6.49** Simulated S11 of the logarithmic spiral antenna with different diameters of the outer spiral

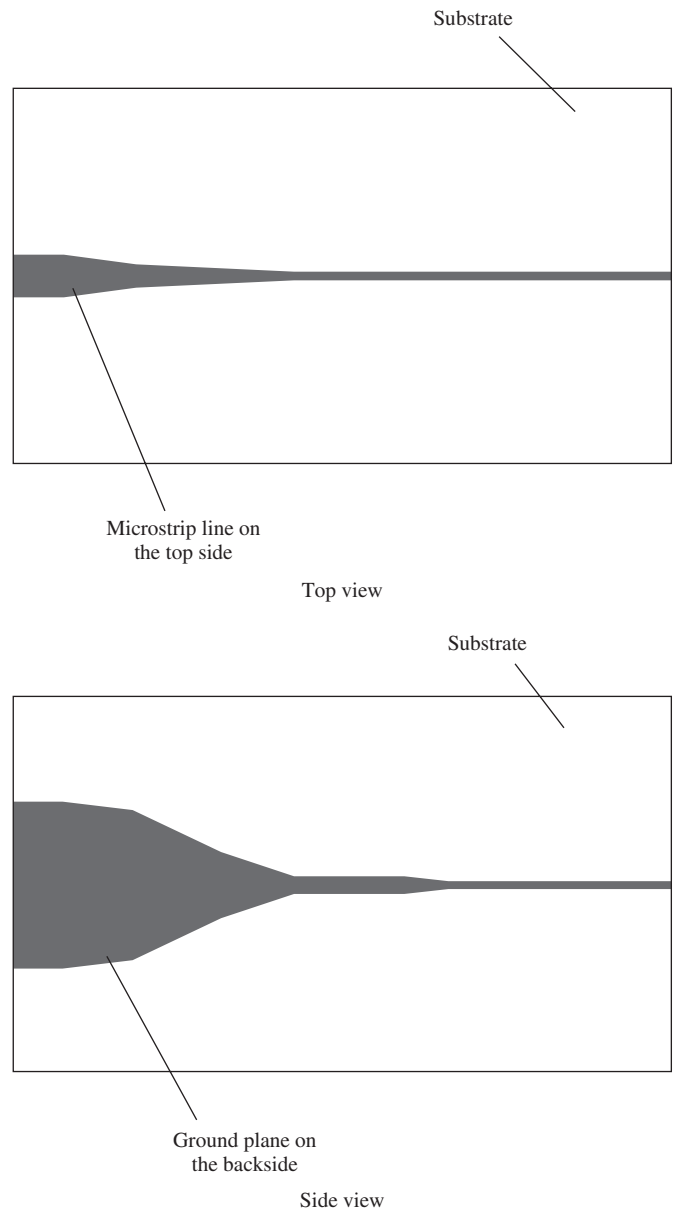
**Table 6.5** Dimensions of the tapered structure

| Part | Width for the microstrip line | Length for the microstrip line | Width for the bottom ground plane | length for the bottom ground plane |
|------|-------------------------------|--------------------------------|-----------------------------------|------------------------------------|
| 1    | 4.8                           | 5                              | 18                                | 5                                  |
| 2    | 2.6                           | 7.7                            | 16                                | 7.7                                |
| 3    | 1.76                          | 9.5                            | 7                                 | 9.5                                |
| 4    | 1.2                           | 7.6                            | 2                                 | 7.6                                |
| 5    | 1.1                           | 11.7                           | 1.8                               | 11.7                               |
| 6    | 0.92                          | 4.7                            | 0.98                              | 4.7                                |
| 7    | 0.88                          | 3.8                            | 0.88                              | 3.8                                |
| 8    | 0.88                          | 20                             | 0.88                              | 20                                 |

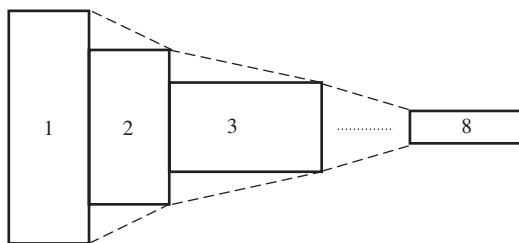
6.6.3 *Logarithmic Spiral Antenna with Wideband Balun*

Then, instead of using an ideal excitation source, the wideband balun is used in the simulation model to feed the logarithmic spiral antenna. Figure 6.52 shows the simulation model of the logarithmic spiral antenna with the wideband microstrip to parallel strip balun.

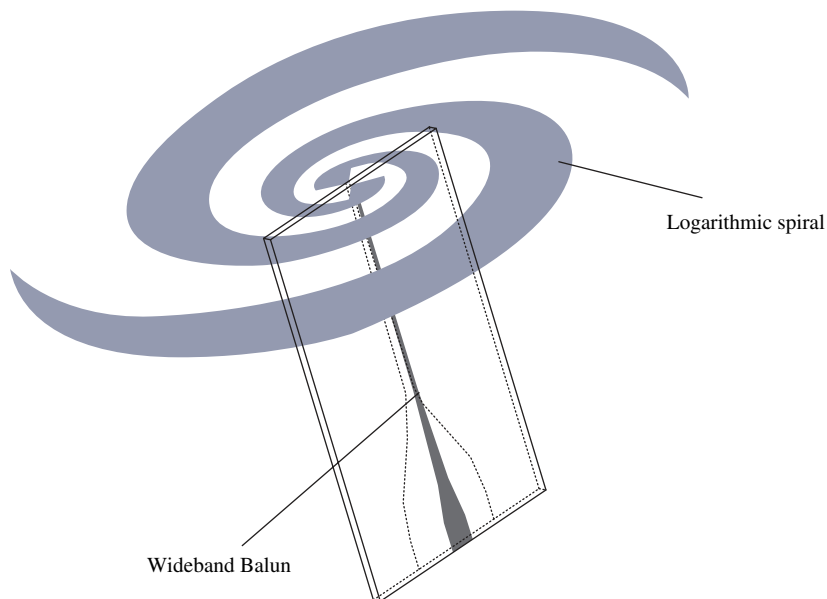
Figure 6.53 shows the simulated return loss and AR of the assembled antenna. It is observed that the return loss of the overall antenna is as good as the case when an ideal excitation source is used. However, the 3-dB AR bandwidth (3.4–7.7 GHz) is about 20% narrower than the ideal case. From this point, it can be concluded that the design of a wideband balun for this type of application is critical.



**Figure 6.50** Top band back view of the microstrip to parallel strip balun



**Figure 6.51** Configuration of the tapering of the microstrip line

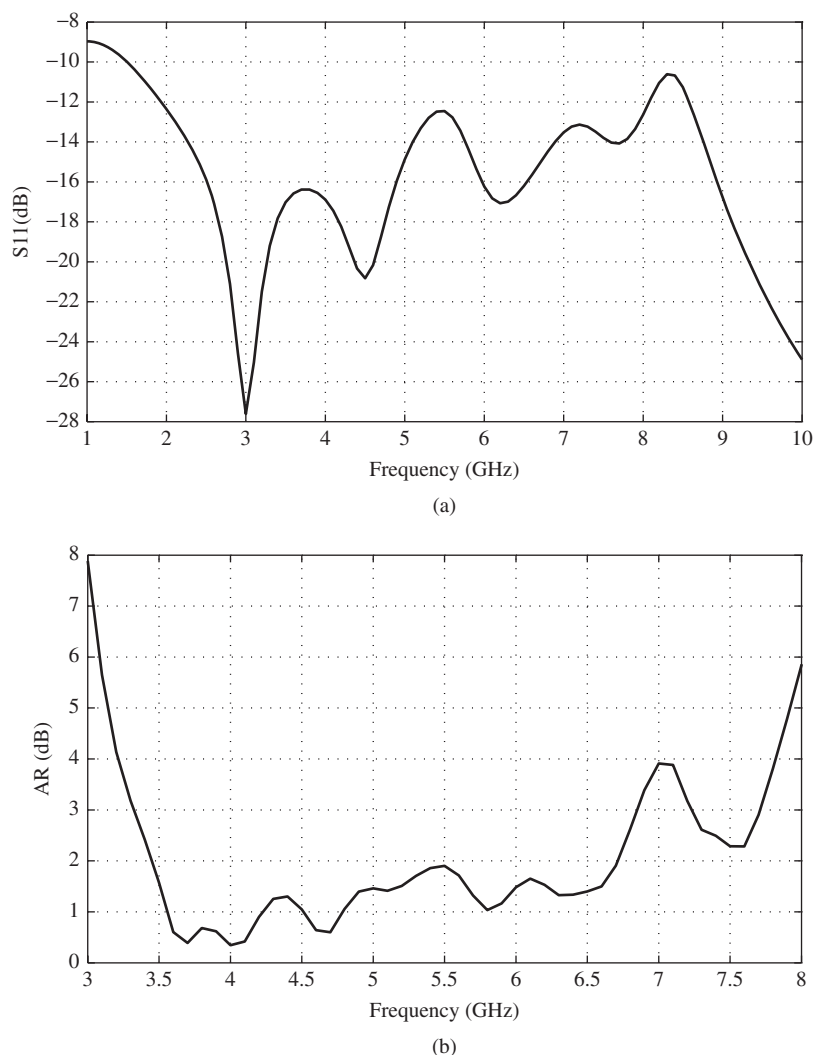


**Figure 6.52** Simulation model of the logarithmic spiral antenna with the wideband microstrip-to-parallel strip balun in HFSS

Figure 6.54 shows the simulated radiation pattern of the logarithmic spiral antenna at different frequencies. As can be seen from this figure, the radiation pattern of the spiral antenna is bi-directional and has stable radiation patterns over the operation frequency band.

Figure 6.55 shows the simulated axial ratio of the logarithmic spiral antenna at 5 GHz with a different  $\theta$  angle. The simulation results show that the logarithmic spiral antenna has a broad 3-dB AR beamwidth.

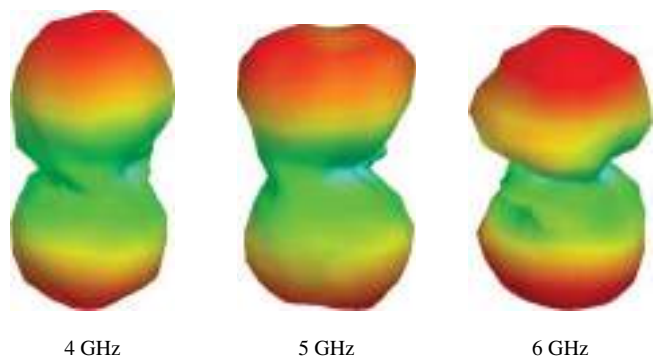
The simulated peak gain of the logarithmic spiral antenna with the balun within the desired frequency band is shown in Figure 6.56. It is found that the logarithmic spiral antenna has a peak gain of 6.8 dBi at 7.7 GHz. Within the 3-dB AR bandwidth, it has an average gain of about 5 dBi.



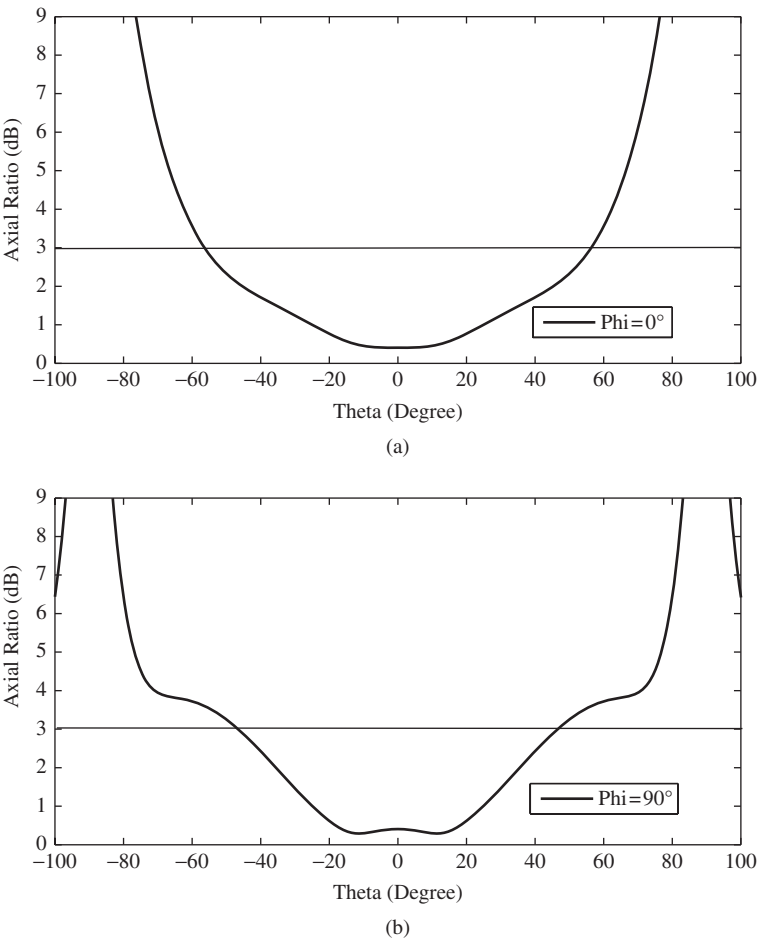
**Figure 6.53** The logarithmic spiral antenna with the wideband microstrip-to-parallel strip balun. (a) Simulated return loss and (b) axial ratio

## 6.7 Summary

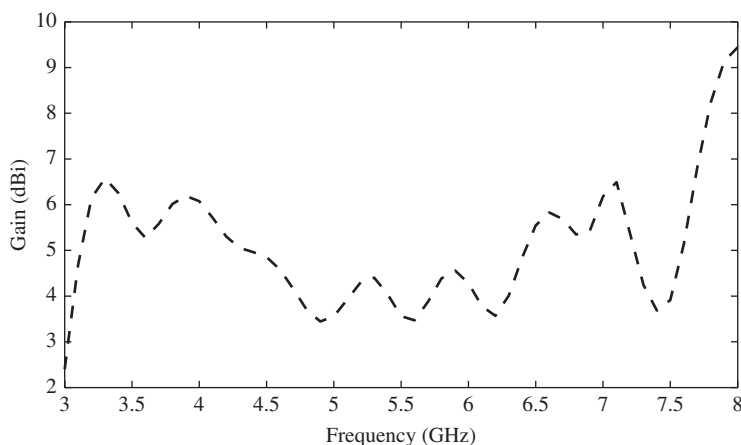
This chapter presents five case studies including three different types of CP antenna elements and two CP arrays. In each case study, design considerations are explained and details of step-by-step design procedures are described. The results of antenna designs are also shown and the whole process from design requirements to each step of practical antenna design is illustrated. These will be useful for readers who want to get started with real-world CP antenna designs. The PQHA is small and has moderate gain but the operation bandwidth is narrow. The logarithmic spiral antenna has a wideband performance but a broadband



**Figure 6.54** Far-field radiation pattern of the CP spiral antenna at different frequencies



**Figure 6.55** Simulated axial ratio of the logarithmic spiral antenna at 5 GHz



**Figure 6.56** Simulated peak gain of the logarithmic spiral antenna with the balun against frequency

balun is required, which actually increases the overall antenna complexity. The printed CP rectangular bi-loop antenna for RFID reader has the most simple configuration due to its single band operation and planar structure. The broadband performance is achieved by using a thick and low permittivity material as the substrate. The dual-band CP array can achieve good CP performance, high gain, low profile and impedance matching at both L1 and L2 bands, thus suitable for small satellite GNSS reflectometry applications. The reflectarray can avoid the use of a feed network and the antenna can be fabricated using PCB technology. However, the downside is that it has an inherently narrow bandwidth.

These antennas exhibit different radiation performances (e.g. dualband operation, high gain or broadband) and they can be applied to different applications based on the system requirements.

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