Single-Feed Ultra-Wideband Circularly Polarized Antenna with Enhanced Front-to-Back Ratio

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Abstract—This paper presents a single-feed ultra-wideband circularly polarized (CP) antenna with high front-to-back ratio. The antenna is composed of two orthogonally placed elliptical dipoles which are printed on both sides of a substrate. It has a low profile, compact size and is demonstrated to be able to operate over a 3:1 frequency range. A simple but effective method of impedance matching is employed to broaden the bandwidth of the antenna. To realize an ultra-wideband CP antenna with high front-to-back ratio, a novel composite cavity is proposed and integrated with the presented crossed dipoles. The composite cavity effectively reduces the side-lobe and back-lobe of the antenna. Detailed analysis of the working mechanism of this cavity is provided. To verify the concept, a prototype covering the GNSS and S frequency bands has been designed and manufactured. Measured results are in good agreement with simulated results and both demonstrate an impedance bandwidth from 0.9 GHz to 2.87 GHz (3.19:1=104.5%) and a 3-dB axial ratio (AR) bandwidth from 1 GHz to 2.87 GHz (2.87:1=96.6%). The measured back-lobe level is below -15 dB and the front-to-back ratio (FBR) is over 25 dB across the whole GNSS band, showing the antenna has superior multi-path mitigation performance. The electric field distributions at different frequency points are given in order to analyze the ultra-wideband CP characteristics of the antenna. Compared to other reported single-feed CP antennas, the antenna has advantages such as a wider CP bandwidth, lower back-lobe radiation and better multi-path mitigation. It is a promising candidate as a multi-function antenna covering several wireless systems.

Index Terms—Crossed dipoles, single-feed antennas, circular polarization, wideband antenna, GNSS, satellite communication, back-lobe reduction.

I. INTRODUCTION

Circularly polarized (CP) antennas are widely used in global navigation satellite systems (GNSS) [1], satellite communication systems, RFID and wireless power transmission systems [2] due to their capabilities of reducing polarization mismatch and suppressing multipath interferences [3].

GNSS systems have found important applications globally in both the military arena and commercial and consumer markets [4]. However, there are several challenges in developing high-performance GNSS antennas. To increase the positioning speed, reliability and availability, the antenna should be able to receive signals from various navigation systems such as global positioning system (GPS), GLONASS, Galileo, Compass and even Indian Regional Navigation Satellite System (IRNSS). Consequently, the operation bandwidth for a high-performance GNSS antenna should extend from 1.1 GHz to 1.62 GHz and even to 2.5 GHz (for IRNSS) with good circular polarization properties.

Designing such a wideband circularly polarized antenna is always challenging. Spiral antennas exhibit interior broadband circular polarization characteristics and thus are extensively studied [5-7]. However, their bi-directional radiation properties make these conventional spiral antennas unsuitable for GNSS applications since a directional radiation pattern is needed to decrease the effects of the reflections from the ground [8]. Several directional spiral antennas including spiral-mode microstrip antenna (SMM) [9], conducting plane backed spiral [10], and cavity backed spiral [11] have been proposed. Helical antennas are another kind of broadband CP antenna, which have been extensively used in the GNSS scenario as well [12].

Another method to make the antenna fulfill wideband coverage requirement is to employ multi-band antennas. Novel multiband antennas, such as an aperture coupled stacked microstrip antenna [13] have been proposed to cover GPS L1 and L2 bands and a triple band CP antenna for GPS application is presented in [14].

Multi-path mitigation is another big challenge for high-performance GNSS antenna design. Since non-LOS (line of sight) multi-path signals could be generated by reflection of satellite signals, the system performance will be degraded once these multi-path signals are received by antennas [15]. Choke ring ground plane [16] is an effective way to mitigate multi-path signals through suppression of surface wave propagating along ground planes. Based on choke ring ground plane, a novel wideband non-cutoff ground plane has been
presented in [8]. Choke rings are, however, rather bulky and expensive. In order to reduce the complexity as well as the cost of choke ring ground plane, a compact-size cross-plate reflector ground plane (CPRGP) has been proposed [17]. The bandwidth of these ground planes is limited and none of them are able to cover such a wide frequency range as required for a high-performance GNSS antenna. One of the objectives of this presented work is to fill this gap.

In this paper, an ultra-wideband CP antenna is presented. It is able to cover all GNSS frequency bands and a part of S-band. At first, a crossed elliptical dipole CP antenna is proposed and a simple but effective impedance matching method is given. Then, a novel composite cavity with unequal-length crossed fins is designed to obtain low backward radiation and high FBR. The proposed ultra-wideband CP antenna can work from 1GHz to 2.87GHz while maintain very low back-lobe level across the whole GNSS band.

The paper is organized as follows: section II introduces the antenna structure and explains the simple but effective impedance matching method for bandwidth enhancement; section III presents the novel composite cavity and its operation mechanism; section IV presents the measurement results and comparisons with simulation results as well as the analysis of electric fields above the cavity aperture; the conclusion is given in section V.

II. ULTRA-WIDEBAND CROSSED ELLIPTICAL DIPOLE ANTENNA

Single-feed crossed dipoles with integrated phase delay lines, producing circular polarization, have been initially presented in [18]. However, the input impedance of this antenna is about 122+j122 Ω and thus cannot be directly connected to a 50Ω feed line. To compensate the impedance mismatching as well as enhance the axial ratio (AR) bandwidth, parasitic open loops have been used [19]. Through changing the shape of dipole, this kind of antenna can be designed for wideband [20, 21] or multi-band [22, 23] operation.

In this section, the antenna configuration will be given at first and then the proposed simple but effective impedance matching method is explained.

A. Antenna Geometry

![Antenna Geometry](image)

![Geometry of proposed antenna: (a) top view, (b) bottom view, (c) side view.](image)

Fig. 1 shows the configuration of the proposed single-feed broadband CP antenna. The antenna consists of two pairs of double-sided elliptical dipoles which are placed perpendicularly to each other. The dipole arm 1 & 2 and 3 & 4 are both connected by ring-shaped phase delay lines, which introduce a 90° phase difference between the orthogonally placed dipole pairs to produce circularly polarized radiation.

The antenna is etched on a 0.817 mm thick Rogers RO4003 substrate with a relative permittivity of 3.55 and a loss tangent of 0.0027. The four elliptical dipole arms are all characterized by a major axis length R₁ and minor axis length R₂ while the phase delay line is a 3/4 ring with inner radius R₃ and line width W₂. A pair of partly overlapped rectangular patches with width W₁ on both sides of the substrate and are used for placing the coaxial feed line. Furthermore, the patches are employed to tune the impedance matching of the proposed antenna, which is discussed in detail in the following text.

Similar to other crossed dipole CP antenna, the proposed antenna radiates in bi-direction. As shown in Fig. 1, current phase on arm 1 is ahead of current phase on arm 2 and thus a RHCP wave is excited along broadside direction while a LHCP wave propagates along downward direction.

B. Impedance Matching

The impedance matching for a crossed dipole CP antenna is usually challenging. Methods such as using 75 Ω coaxial line [18], employing parasitic elements [19] and changing the shape of dipole [21] have been proposed to achieve good impedance matching. However, these approaches are either complex or space-consuming.

A simple but effective impedance matching method has been utilized in the proposed antenna, which is simpler and more effective than the aforementioned ones. By using this method, a 3:1 impedance bandwidth can be achieved without increasing complexity and space of the presented antenna.

A magnified picture of the partly overlapped rectangular patch is given by Fig. 2. As shown in the picture, the pair of
rectangular patches is printed on both sides of the substrate with an overlapped area of $W_1 \times L_1$. This overlapped structure of two patches results in a pair of parallel-plate capacitors, i.e. capacitor $C_1$ and $C_2$ shown in Fig. 2. It is found that these two capacitors can effectively enhance the impedance characteristics of proposed antenna and thus broaden the antenna’s bandwidth.

![Fig. 2. Configuration of overlapped rectangular patch: (a) top view, (b) bottom view.](image)

To analyze the proposed impedance matching mechanism, qualitatively, the equivalent circuit model of the antenna is provided in Fig. 3. The symbol $Y_A$ denotes the antenna admittance without the overlapped rectangular patches while $C_1$ and $C_2$ denote the parallel-plate capacitors and $Y_{in}$ represents the total input admittance of the proposed antenna.

![Fig. 3. Equivalent circuit model of proposed antenna.](image)

Using the formula provided in [24], the capacitance of parallel-plate capacitors $C_1$ and $C_2$ can be calculated by the following equation when the electric charge density on the plates is uniform and the fringing fields at the edges can be neglected.

$$C_1 = C_2 \approx \frac{\varepsilon_0 \varepsilon_r W_1 L_1}{h}$$

(1)

The approximation derives from the small slot on bottom rectangular patch and thus the input admittance $Y_{in}$ of antenna can be denoted by:

$$Y_{in} = Y_A + 2j\omega \frac{\varepsilon_0 \varepsilon_r W_1 L_1}{h}$$

(2)

It is shown in equation (2) that the antenna input admittance $Y_{in}$ is affected by the rectangular patch width $W_1$ and length $L_1$. Therefore, the impedance characteristic of the antenna can be tuned by these two parameters to achieve a good impedance matching.

In order to demonstrate, intuitively, the effect of changing parameters $W_1$ and $L_1$, the simulated input impedances under different pairs of $W_1$ and $L_1$ are studied while keeping the other antenna geometry parameters fixed, as in TABLE I.

<table>
<thead>
<tr>
<th>$L_1$</th>
<th>$W_1$</th>
<th>$R_1$</th>
<th>$R_2$</th>
<th>$R_3$</th>
<th>$h$</th>
</tr>
</thead>
<tbody>
<tr>
<td>6.2mm</td>
<td>1.5mm</td>
<td>23.4mm</td>
<td>18mm</td>
<td>5.2mm</td>
<td>0.817mm</td>
</tr>
</tbody>
</table>

The simulated input impedance of the proposed antenna for various values of $L_1$ is shown in Fig. 4. It is worth noting that the rectangular patch width $W_1$ is kept constant as 7.5mm while the length $L_1$ is varied. As can be seen from Fig. 4, the input impedance without the overlapped patch (case $L_1 = 0$ mm) is large inductively and the majority of the impedance loci falls outside the VSWR = 2 circle, indicating that the bandwidth at this state is very narrow. By increasing $L_1$, the impedance loci begin to move along the admittance circle due to the increasing capacitance of $C_1$ and $C_2$. This phenomenon can be well explained by equation (2) as well.

![Fig. 4. Simulated input impedance of proposed antenna with different patch length $L_1$.](image)

Fig. 5 depicts the input impedance loci under different patch widths, $W_1$, when patch length $L_1$ is 2.5 mm. Similar to the phenomenon observed in Fig. 4, the impedance loci will move along the admittance circle when $W_1$ becomes larger. This impedance matching procedure is similar to the method using lumped LC elements to tune the impedance bandwidth. However, the proposed method is more advantageous in terms of antenna efficiency and complexity due to the absence of a lossy $\pi$-type or M-type LC circuit.

![Fig. 5. Simulated input impedance of proposed antenna with different patch width $W_1$.](image)
C. Results

The prototype of the proposed antenna is shown in Fig. 6. As shown in Fig. 6, the antenna is etched on a 115mm × 115mm Rogers RO4003 substrate and is fed directly by a 50Ω SMA connector.

Fig. 6. Prototype of proposed antenna: (a) top view, (b) bottom view.

Fig. 7 shows the simulated and measured VSWR of the proposed broadband circularly polarized antenna. It can be seen from the measured result that the antenna can achieve a VSWR < 2 bandwidth from 0.96GHz to 3.02GHz (3.14:1 = 103.5%), which is much wider than aforementioned crossed dipole antennas [18-23]. Meanwhile, it is also worth noting that the antenna size is only 106mm × 106mm (0.34λm × 0.34λm, λm denotes the free space wavelength at the minimum working frequency), which is smaller than the regular crossed dipole antenna as shown in TABLE II.

![Simulated and measured VSWR of proposed antenna](image)

**Fig. 7.** Simulated and measured VSWR of proposed antenna

### TABLE II

<table>
<thead>
<tr>
<th>Ref. No.</th>
<th>Antenna Size (mm)</th>
<th>Minimum Working Frequency fwa (GHz)</th>
<th>Antenna Size (in wavelength)</th>
<th>Impedance Bandwidth (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[18]</td>
<td>120 × 120</td>
<td>2.01</td>
<td>0.8λm × 0.8λm</td>
<td>30.7%</td>
</tr>
<tr>
<td>[19]</td>
<td>120 × 120</td>
<td>1.97</td>
<td>0.79λm × 0.79λm</td>
<td>38.2%</td>
</tr>
<tr>
<td>[20]</td>
<td>60 × 60</td>
<td>0.877</td>
<td>0.18λm × 0.18λm</td>
<td>11.8%</td>
</tr>
<tr>
<td>[21]</td>
<td>55 × 55</td>
<td>1.99</td>
<td>0.36λm × 0.36λm</td>
<td>50.2%</td>
</tr>
<tr>
<td>This Paper</td>
<td>106 × 106</td>
<td>0.96</td>
<td>0.34λm × 0.34λm</td>
<td>103.5%</td>
</tr>
</tbody>
</table>

III. NOVEL COMPOSITE CAVITY FOR MULTIPATH MITIGATION APPLICATION

Multipath interference, caused by the deleterious superposition of signals received at a user’s antenna via multiple paths from a satellite, is one key source of error in relative positioning of GNSS systems [25]. For GNSS receiving antennas, the ability of multi-path mitigation is critical for achieving high performance. Theoretically, a good multipath mitigating GNSS antenna should be able to have a cross-polarization rejection ratio (or polarization isolation) ≥ 15 dB for signals incoming at any positive elevation angle and back radiation less than -10 dB to decrease the effects of the reflections from the ground [8]. In addition, the pattern roll-off from zenith to horizon should be between 8 and 14 dB, while phase center variation should be less than 2 mm [17].

The aforementioned requirements are very difficult to achieve simultaneously and thus certain compromises need to be made within the antenna design procedure. Moreover, considering that the proposed antenna in section II is a ultra-wideband antenna, further compromise is expected to be made. Besides, since the proposed crossed elliptical dipole antenna is bi-directional, making the antenna radiate in one direction, i.e. improving front-to-back ratio (FBR), is regarded as the main goal during the antenna optimization procedure.

A. Design Concept

The proposed novel composite cavity is shown in Fig. 8. It can be seen from Fig. 8 that the ground plane is composed of a cylindrical cavity with crossed unequal length fins. The composite cavity is designed in such a way to reduce the back-lobe of the antenna maximally.

![Geometry of proposed composite cavity](image)

**Fig. 8.** Geometry of proposed composite cavity: (a) top view, (b) prospective view.

To explain the working mechanism of the proposed composite cavity, similar analysis methods to [8, 17] are adopted. It is well known that the surface wave propagating along the ground plane will reradiate at the ground plane edge and thus increase the side-lobes as well as the back-lobe of the antenna. Therefore, suppression of surface waves can reduce the back-lobe of the antenna.
Considering the two TEM plane waves I and II shown in Fig. 9, they can be denoted by the following equations respectively.

\[ E_1(y, t) = E_{01}\cos(ky - \omega t + \varphi_1) \]  
\[ E_2(y, t) = E_{02}\cos(ky - \omega t + \varphi_2) \]  

The boundary condition of flat conductor ground plane imposes that no tangential electric field can exist, i.e.

\[ E_t = 0 \]  

This equation indicates that plane wave I cannot propagate along the circular ground plane while wave II can exist. Considering the situation in Fig. 9 (b), wave I cannot propagate along the cavity bottom either due to the boundary condition. In contrast to the circular ground plane, wave II cannot propagate along the cavity either since there is a mandatory boundary condition imposed by the crossed fins. Therefore, it can be expected that the proposed composite cavity can offer better ability of surface wave suppression compared to regular circular flat ground plane.

As indicated above, the proposed antenna radiates in bi-direction, RHCP towards the broadside and LHCP backward. Conventionally, a large ground plane or a cavity is required to reduce the back-lobe of such kinds of antenna by reflecting the downward waves into the upward direction [19, 22]. The proposed composite cavity placed underneath the antenna can afford better back-lobe reduction ability by minimizing diffracted waves; the mechanism can be explained as follows.

Fig. 10 depicts the propagation of downward LHCP wave excited by the antenna under different situations. When the proposed antenna is placed above a large flat ground plane, the majority of the LHCP wave will be reflected by the ground plane. However, considerable downward waves can still propagate across the ground plane due to the diffraction of these waves and result in large back-lobe. This diffraction effect can be decreased by replacing the flat ground plane with a cylindrical cavity shown in Fig. 10 (b) as the peripheral vertical wall can stop part of the diffracted waves.

A LHCP wave depicted in Fig. 10 can be expressed by the following equation.

\[ \vec{E}(z, t) = \vec{E}_x \cos(-kz - \omega t + \pi/2) + \vec{E}_y \cos(-kz - \omega t) \]

\[ |\vec{E}_x| = |\vec{E}_y| \]  

As a circularly polarized wave, the electric field vector \( \vec{E} \) will rotate on xy plane. To analyze the downward wave around the peripheral wall of the proposed composite cavity, two transient plane waves III and IV whose electric field are along the +x and +y direction, respectively, have been plotted in Fig. 10 (c). These two waves cannot propagate along the vertical fins due to the boundary condition given by equation (5). Actually all the 16 vertical fins can be used to block the diffracted waves when the direction of the transient \( \vec{E} \) vector is parallel to these fins. Therefore, the proposed composited cavity is more advantageous for back-lobe reduction than the regular cylindrical cavity.

B. Comparison of Back-lobe Reduction

It has been shown in the last section that the proposed composite cavity can offer better back-lobe reduction ability than a flat ground plane and regular cylindrical cavity. This performance enhancement is mainly from the suppression of surface waves and the blocking of diffracted waves around the edge of the ground plane. To verify this analysis, a comparison between a flat plane ground, regular cylindrical cavity and the proposed composite cavity, in terms of the back-lobe reduction, is given.

The geometry dimensions of proposed composite cavity have been given in TABLE III.

<table>
<thead>
<tr>
<th>TABLE III</th>
<th>COMPOSITE CAVITY DIMENSIONS (MM)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W_3</td>
<td>W_4</td>
</tr>
<tr>
<td>35</td>
<td>25</td>
</tr>
</tbody>
</table>

For the flat ground plane and regular cylindrical cavity, the most critical dimensions for back-lobe reduction are the ground plane size as well as the height from antenna to ground plane. To make an unprejudiced and reasonable comparison, the proposed antenna in section II is placed at a height of \( h_1 = 50 \) mm to the ground plane for all the three cases. Meanwhile, the radius of the ground plane is kept at 110mm.

The simulated back-lobe levels for the three ground plane reflectors are shown in Fig. 11. As can be seen from this figure, the back-lobe is very high when a flat ground plane is placed underneath the broadband CP antenna. By using the regular cylindrical cavity, back-lobes smaller than -12 dB can be achieved across the whole GNSS band (1.1 - 1.62 GHz). Further back-lobe reduction has been observed through
utilizing the proposed composite cavity. It can be noted that a minimum 8 dB and 2 dB enhancement in back-lobe reduction can be achieved compared to flat ground plane and regular cylindrical cavity respectively when the proposed composite cavity is used as a reflector of ultra-wideband CP crossed elliptical dipole.

IV. RESULTS AND ANALYSIS

In this section, the simulated and measured results of the proposed composite cavity backed ultra-wideband CP antenna are given to verify the overall radiation performance. The electric field distribution in the aperture is provided to demonstrate its circularly polarized characteristics intuitively.

A. VSWR

The prototype of the proposed ultra-wideband composite cavity backed CP antenna is shown in Fig. 13. The rim of this cavity is made by gluing two semi-circle shaped conductors together. Therefore, the rim is not in a perfect circular shape, which is expected to result in some differences between simulated and measured results.

B. Axial Ratio

Axial ratio (AR) is an important index to evaluate a
circularly polarized antenna. The frequency bandwidth within which the broadside AR is smaller than 3dB is named as the AR bandwidth of an antenna. The AR bandwidth is generally narrower than the impedance bandwidth for a CP antenna and thus the bandwidth of a CP antenna is usually referred as the AR bandwidth.

To evaluate the CP performance of the proposed antenna, the AR characteristics versus frequency relationship is given in Fig. 15. It can be seen from this figure that the measured 3dB AR bandwidth is from 1 GHz to 2.87 GHz (2.87:1=96.6%), which can cover the whole GNSS bands (including higher band of IRNSS) and S band as well.

![Fig. 15. Simulated and measured AR of proposed antenna.](image)

**C. Radiation Pattern**

Since the proposed antenna is measured using a linearly polarized standard horn antenna, only the two linear components $E_\theta$ and $E_\phi$ have been obtained.

Fig. 16 shows the radiation patterns of the proposed ultra-wideband CP antenna in two main planes (XoZ and YoZ planes) at 1.15 GHz, 1.4 GHz, 1.7 GHz, 2.1 GHz and 2.4 GHz. As shown in this figure, the proposed antenna can maintain smaller than -15dB back-lobe and larger than 25dB FBR across the whole GNSS frequency band.

![Fig. 16. Simulated and measured radiation patterns: (a) 1.15GHz, (b) 1.4GHz, (c) 1.7GHz, (d) 2.1GHz, (e) 2.4GHz.](image)

**D. Analysis of the CP Characteristic**

As stated in [26], the radiation of a cavity backed antenna is determined more directly by the electric field distribution in the aperture than by the current on the exciter. To investigate the ultra-wideband CP characteristic of the proposed antenna, the electric field distributions at 10 mm height above the aperture at 1 GHz, 1.8 GHz and 2.6 GHz are given in Fig. 17.
As can be seen from the figure, the distribution of electric field is varied at different frequencies. From Fig. 17 (a), it can be noticed that the maximum electric field at phase angle 0° and 90° generate at the end of elliptical dipole arm. This indicates that the proposed antenna works at a half-wavelength dipole mode at 1 GHz. When the frequency increases to 1.8 GHz, the strongest electric field occurs at the side edge of the elliptical dipole arm (see phase angle 45° and 135° in Fig. 17 (b)). Therefore, it can be concluded that the resonance length of proposed antenna is decreased, which makes the antenna work at higher frequency band. At 2.6 GHz, the strong electric field no longer appears just around the antenna (see phase angle 90° in Fig. 17 (c)) and there are evident strong fields between the antenna and the cavity. Comparing this field distribution with those at lower frequencies, it is found that a higher mode is excited, which results in the deterioration of radiation pattern shown in Fig. 16 (e).

From the different electric field distributions at phase angle 0°, 45°, 90° and 135° in Fig. 17, it can be seen that a right-hand rotated electric field has been produced at all given frequencies. This indicates that the antenna can work as a circularly polarized antenna in an ultra-wideband frequency range.

In addition, the electric fields are almost all bounded by the cavity and little fields can propagate over the cavity as shown in Fig. 17 (a) and (b). This phenomenon also verifies the back-lobe reduction ability of proposed antenna, which has been analyzed in section III. On the contrary, it is shown in Fig. 17 (c) that the electric fields can propagate over the cavity at 2.6 GHz and thus the back-lobe and side-lobe reduction ability of the cavity is limited. This conclusion is in good agreement with the simulated and measured radiation patterns at higher frequencies.

V. CONCLUSION

A single-feed ultra-wideband circularly polarized antenna has been proposed and investigated in this paper. A simple but effective impedance matching method is employed to enlarge the impedance bandwidth and a composite cavity is presented to reduce the back-lobe of the antenna. Good agreement between simulation results and measurement results proves that this antenna can work in a nearly 100% bandwidth region as a CP antenna. Analysis of the electric field above the cavity, as given in the paper, has explained the ultra-wideband CP operation of the proposed antenna. With a lower than -15dB back-lobe level and a larger than 25dB FBR character, this antenna is very promising for high-performance GNSS application as well as S band applications.

ACKNOWLEDGMENT

The authors would like to thank for the funding from EC FP7 GaNSat project under the contract number 606981. The authors also thank Mr. Simon Jakes and Mr. Clive Birch for their help in antenna fabrication.

REFERENCES


