An approach to achieve directional low-profile antenna of quintuple stable pattern band by utilising dipole with compound concave corrugated reflector

Chaofan Ren | Junping Geng | Han Zhou | Kun Wang | Xianling Liang | Weiren Zhu | Ronghong Jin

Department of Electronics Engineering, Shanghai Jiao Tong University, Shanghai, China

Correspondence
Junping Geng, Department of Electronics Engineering, Shanghai Jiao Tong University, No. 800 Dongchuan Road, Shanghai 200240, China.
Email: gengjunp@sjtu.edu.cn

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Abstract
An approach to achieve directional low-profile UWB antenna by omnidirectional source with a reflector, is proposed. The general theoretical model about the direction and the phase conditions between the upward and reflected field are developed. A bow-tie dipole source antenna is investigated first. To modify its radiation defect, proper phase equations between the original upward field and the reflected downward field are derived so that the aperture field becomes uniform enough to guarantee directional radiation. The concave-shaped reflector is primarily adopted to satisfy the phase conditions. Then the concave-shaped reflector is decorated with corrugated structure to realize good impedance matching within the wide band. Consequently, the split beams caused by multiple long linear currents with phase difference at high frequency is focussed while the directivity at low frequency is also improved. The prototype antenna was fabricated with a profile of 20.5 mm, about 0.156 λ_{max} (maximum wavelength in the work band), and the measured impedance bandwidth of VSWR ≤ 2.2 is over 2.27–26 GHz. Both simulated and measured results demonstrate the stable directional patterns within the quintuple band (2–10.5 GHz). The realised gain is over 7 dBi within the band (2–10.5 GHz) and peak is over 12 dBi.

1 | INTRODUCTION

Ultrawideband antennas have great potential and application prospects in wireless communication systems, wireless body area network, radar system, radio astronomy, medical imaging and MIMO applications [1–6].

There are many kinds of structures applied to UWB antenna designs. In general, the main beams of most UWB antennas are unstable within a wide band, and these antennas often have no ground [7, 8]. When the UWB antennas are mounted above a metal ground in the practice, their performance usually deteriorates seriously [9]. Therefore, developments of low profile UWB antennas with stable directional radiation performance attract a lot of attention.

To enhance the directional gain of UWB antenna, researchers firstly utilise flat reflectors or cavity reflectors to reflect the backward radiation [10–12]. The profile height is usually designed to be a quarter of wavelength at mid-band to balance the performance of the whole band. In Ref. [10], a reverse T-match dipole is mounted on a flat metal ground plane which achieves a bandwidth of 2.25:1 (2.89–6.5 GHz) with a profile of 0.17 \( \lambda_{\text{max}} \). Reference [11] presented a differential-fed folded antenna with a box-shaped reflector. A bandwidth of 4.88:1 (2.48–12.12 GHz) is achieved while its profile height climbed to 0.26 \( \lambda_{\text{max}} \). In Ref. [12], a simple parasitic top patch is used to act as the director which achieves a nearly 3:1 operating bandwidth (2.2–6.5 GHz) with a profile of 0.205 \( \lambda_{\text{max}} \) while its pattern begins to split at 6 GHz.

To lower the profile height of UWB directional antenna, several special reflectors are introduced to conventional UWB antenna such as planar spiral antenna and printed dipole antenna [13–15]. In Ref. [13], the directional radiation is achieved...
by three-stepped narrow rectangular slot antenna with a frequency selective surface (FSS) reflector. Its bandwidth is 3.75:1 (3.2–12 GHz) and its profiled height is 0.20 $\lambda_{\text{max}}$. In Ref. [14], a circular high-impedance surface (HIS) reflector is introduced to spiral antenna. It achieves a bandwidth of 3.3:1 (3–10 GHz) with a profile of 0.09 $\lambda_{\text{max}}$ but its main beam splits in 4.5–6.2 GHz. Besides the FSS and HIS reflector, artificial magnetic conductor (AMC) is also introduced to two crossed dipoles to design a low-profile circular polarisation directional antenna in Ref. [15]. Its profile is 0.08 $\lambda_{\text{max}}$ and the impedance bandwidth is 1.99:1 (1.19–2.37 GHz).

Although the HIS and AMC reflector can achieve directional radiation with a very low-profile height, its stable directional radiation pattern bandwidth (SDRPB) is restricted to 3:1. To break the limitation, some researchers proposed a few tridimensional structure and reflectors to broaden the bandwidth [16–19]. A bow-tie dipole with corrugated reflectors was proposed to realize low-profile directional UWB antenna [16]. Its whole profile is 0.193 $\lambda_{\text{max}}$. Its impedance bandwidth is 3.03:1 (2.75–8.35 GHz) while its SDRPB reduced to 2.61:1 (2.3–6 GHz). In Ref. [17], a modified horned reflector is utilised to achieve a bandwidth of 3.87 (2.81–10.92 GHz) with a profile of 0.30 $\lambda_{\text{max}}$. In Ref. [18], a tridimensional-folded bowtie-shaped element is presented to achieve a bandwidth of 2.02 (1.27–2.57 GHz) with a profile of 0.07 $\lambda_{\text{max}}$ while its gain in the whole band is around 5 dBi. Conventional Vivaldi antenna is a good choice of stable directional UWB antenna, but the dimension towards the radiation direction is quite large and has no ground. In [19], the Vivaldi antenna is bended to reduce the profile and a corrugated plate is added as a ground. It achieves an impedance bandwidth of 11.1:1 (0.9–10 GHz) with a profile of 0.216 $\lambda_{\text{max}}$ while its pattern begins to tilt when the frequency is over 6.5 GHz.

It can be summarised that the special reflector like FSS and AMC can achieve great impedance match performance with a very low profile. However, because of the reflection dispersion with frequency changing, the patterns of antenna with these reflectors usually distort in a very wide band which cause the SDRPB restricted to 3.75:1. In addition, the tridimensional metal reflector is non-frequency-dispersive which causes it to have a great potential to achieve UWB stable direction radiation. Thus, the general design method of tridimensional reflector design with a low-profile attracts more interests.

Generally, the low profile directional UWB antenna design is a trade-off among impedance bandwidth, SDRPB and profile. When the reflector is close to the source antenna, the coupling between the antenna and its image becomes strong, so that the impedance matching characteristics deteriorates rapidly in low frequency band. Once the distance between reflector and antenna increases, the coupling is reduced so that the impedance matching performance in the low frequency band becomes better. But the patterns become worse in high frequency band because the reflected field and the direct field are not in-phase.

The method to adjust the distance between reflector and source antenna is only effective for bandwidth below 3:1, and it is inapplicable for wider band directional antenna.

Here, the approach to design a stable directional UWB antenna with low profile is to find a reflector, which can reduce its coupling to source antenna for better impedance matching performance in lower frequency band, and improve the phase uniformity for better directional radiation performance in higher frequency band at the same time.

In this article, the general theoretical UWB directional radiation model and mathematical principles are constructed, and the corresponding approach to achieve directional low profile UWB antenna by omnidirectional source with the reflector is proposed. The bow-tie dipole acts as source antenna and the compound concave corrugated reflector (CCCR) is utilised to improve the phase uniformity on the antenna aperture. The corrugated structures are introduced to reduce the coupling between the source antenna and the reflector. In Section 2, a general model of directional radiation is introduced. Then the UWB bow-tie dipole is studied and its concrete equivalent model with the reflector, to realize directional radiation, is analysed. Moreover, the corrugated structures are studied. Thus, the CCCR is introduced and its performances are discussed. In Section 3, the simulated results by using computer simulation technology (CST) software and measured results are presented. In Section 4, the conclusion is presented finally.

The novelty of this article is described as follows. A concrete model to achieve directional radiation with the equations of phase condition is developed for the first time. Most of the geometric parameters of the antenna structure can be derived by the equations which caused less reliance on commercial electromagnetic software. Meanwhile, the concise expression of corrugated structure based on the equivalent depth is also given for the first time which can make the corrugated structure be easily utilised in the other arbitrary antenna design. Moreover, the comprehensive performance of the prototype antenna is better than the existing research results.

2 | DIRECTIONAL RADIATION ANALYSIS AND ANTENNA DESIGN

2.1 | General model

A simple approximate model is given to describe the radiation performance of the arbitrary reflector in Figure 1. An isotropic point source is located at origin and S is the point on the reflector. The downward field is reflected to a specific direction which is determined by the normal vector $\mathbf{n}$ of the reflector. It is obvious that the total field is the superposition of the upward field and reflected field in the upper half space.

Suppose the direction angle of $\theta \leq \theta_0$ is the goal radiation direction. The origin upward field is normalised. Thus,

$$E_{\text{upward}} = 1$$  \hfill (1)

As for the arbitrary point on the reflector, the radiation direction of the reflected field pointing to the goal direction when the normal vector meets the follow vector equation.
\[ |(\vec{R}/|\vec{R}| + 2) \cdot \vec{n}| \leq \sin \theta_0 \]  

(2)

\( \vec{R} \) is the position vector of the point on the reflector.

According to (2), when the point source is placed on the focus of a paraboloid, the paraboloid can reflect all the direction of the downward field to single direction which means the paraboloid may be applied in the design of the reflector.

The reflected field is the integral of all the angles satisfying (2). Compare with the upward field, the additional paths cause the phase difference of currents on dipole cannot be neglected.

When the length increases from half wavelength to three half wavelengths, the phase can be regard as typical sinusoidal distribution.

Moreover, when the dipole works at a very high frequency, the width of the dipole may reach one wavelength, which means the transverse currents will perform an important and even negative role in radiation.

For a better description about dispersion of UWB linear source, an actual UWB dipole antenna will be analysed in the follow section.

### 2.3 Analysis of bow-tie dipole antenna

As is known, printed bow-tie dipole antenna has wide impedance band. In order to ensure stable radiation patterns, the length and the width of dipole need to be less than \( \lambda_{\text{max}} \) and 0.8 \( \lambda_{\text{max}} \) respectively [16]. As a result, the bow-tie dipole becomes inapplicable because of its unstable radiation patterns in high frequency band although its impedance characteristics may perform excellently in very wide band. The bow-tie dipoles of \([12, 15]\) all have to fit this restricted condition to obtain stable patterns which causes these antennas working in the limited band. To broaden stable directional radiation pattern bandwidth of bow-tie dipole, the restricted condition must be broken.

As shown in Figure 3a, a bow-tie dipole antenna was designed on an Arlon AD255 substrate (\( \varepsilon_r = 2.55 \)) with a thickness of 0.508 mm. The values of parameters on the bow-tie dipole are shown in Table 1. The simulated return loss of the bow-tie dipole is shown in Figure 3b. Also, the impedance bandwidth of \( S_{11} < -10 \text{ dB} \) was more than 8.5 GHz over 2.5–11 GHz. It can be observed that there are two obvious valley points which may indicate that the dipole works in different modes.

The bow-tie dipole is placed in \( xoz \) plane. The simulated surface current distributions and patterns of the bow-tie dipole are shown in Figure 4. At 4 GHz, the bow-tie dipole is like a typical thin dipole. The main direction of surface current is parallel to \( y \) axis in Figure 4a, so that the far field beam is mainly distributed in \( xoz \) plane in Figure 4e. The similar results are shown in Figure 4b and 4f at 6 GHz. But in Figure 4c, at 8 GHz, the edge length of the bow-tie dipole is larger than half wavelength. There are current wave nodes at the edge, and the reverse currents distribute along the edges. In addition, the central currents and edge currents have reverse flow direction, and several beam lobes appear in \( xoy \) plane as shown in Figure 4g. The results shown in Figure 4d and 4h at 10 GHz are similar.

In Figure 4a–d, the surface currents mainly distribute at the edge of the dipole. In Figure 4i, the normalised phase of \( y \) axial currents on the red line in a–d is shown when the \( x \) axial...
currents are neglected. The phase of y axial currents on the middle of red line is normalised to 180°.

To obtain a simple description of the current on the dipole, the spatial sampling of currents is taken. Because the maximum phase variation range is less than 180° in the whole width of the dipole, according to the spatial Nyquist sampling theorem [20], three sampling currents are enough to analyse the currents on the dipole. Thus, a simple theoretical model of the bow-tie dipole is shown in Figure 5. Two edge currents (I₁ and I₆) and one central current (I_c) are sampled to characterise all the y axial currents on the dipole. Because the edge currents flow along the edge of dipole, the additional length (Δl) causes phase difference between the edge currents and central current.

When the bow-tie dipole operates at low frequency band such as 4 GHz, the phase difference between edge currents and central current is less than 45°, so the pattern is approximate omnidirectional. As shown in Figure 4b, when the operating frequency increases, and the distance between I₁ and I₆ approaches half wavelength, the radiation field of I₁ and I₆ cancels each other out in x-axis direction while the far field in the z-axis direction is the superposition of field radiated by three currents. Consequently, the pattern changes from omnidirectional to bidirectional as shown in Figure 4f.

When the frequency reaches around 10 GHz, the distance between the edge currents and central current approaches half wavelength so that the edge currents and the central current have reversal phase which causes their radiation field cancelled in z-axis direction. In the x-axis direction, I₁ and I₆ have the same phase, and the feeding phase difference between the edge currents and the central current is compensated by the space phase difference, which causes in-phase far field superposition.

To obtain a directional antenna, dipole is often placed above a reflector which can be regarded as aperture antenna. The directivity is determined by the distribution of the fields at the aperture. To achieve high directivity in the goal direction, the aperture fields need to be as uniform as possible. Compared with amplitude, it is easier to adjust the phase of aperture fields by designing the shape of the reflector. There are two key principles of the reflector design listed as follows.

- The phase of the upward field directly radiated from the bow-tie dipole at the aperture must be the same as that of the reflected field from the reflector in the goal orientation.
- The radiation fields generated by the edge currents and central currents must have the same phase in the goal orientation.

2.4 Compound concave corrugated reflector

According to the key principles of directional antenna mentioned above, the design schematic of dipole with concave reflector is shown in Figure 6. To obtain high directivity, a parabolic section reflector is firstly used to reflect the downward fields to the upward direction. And a flat plate is inserted into the bottom of the parabolic reflector to lower the profile. Depending on the different incidence angle, the concave reflector can be divided into three regions, I, II and III, when the incident wave is linearly polarised.

Region I is just below the bow-tie dipole, so it is considered to be vertical incidence. When the incident point moves to the edge of region II, the wave vector of resultant reflected fields mainly points to the upward direction. To achieve in-phase reflection, the phase difference between the reflected fields and the original upward fields needs to be less than π/2 at the aperture. Thus, b needs to satisfy the follow equation.

\[
2k\pi - \frac{\pi}{2} \leq \frac{2\pi}{k} 2b - \pi \leq 2k\pi + \frac{\pi}{2}
\]  

Select 10.5 GHz as the reference frequency, and it is obvious that \(k = 1\) at 10.5 GHz, so b needs to be larger than 17.85 mm.

In region II, with the increase of the oblique incidence angle, the wave vector of resultant reflected downward fields deviates from the upward goal direction. Because of the short wavelength at high frequency band, the phase difference between the reflected fields and original upward fields can reduce to zero, and thus the side lobe appears. However, the amplitude of the side lobe is relatively small compared with the main lobe.
In region III, with the reflection of paraboloid, the oblique downward fields can be reflected to the goal direction. At the same time, the wave path difference between $L_C$ and $L_R$ needs to compensate the phase difference between central currents and edge currents.

$$
\frac{2\pi}{k} (L_R + \Delta l - L_C) \leq \frac{\pi}{2}
$$

Where $\Delta l$ is the additional path length of edge currents compared with central currents on the bow-tie dipole in Figure 5. According to (6), because $\Delta l$ is assumed to be constant, the junction of region II and III satisfies hyperbolic...
equation. Thus, the boundary of region III is the intersection of a hyperbolic curve and the parabolic curve.

In addition, the phase difference between reflected downward fields and upward fields needs to be less than $\pi/2$.

$$2k\pi - \frac{\pi}{2} \leq \frac{2\pi}{k} (L_R + b) - \pi \leq 2k\pi + \frac{\pi}{2}$$  \(7\)

The phase difference between the downward fields generated by $I_C$ and $I_R$ also needs to be less than $\pi/2$.

$$2k\pi - \frac{\pi}{2} \leq \frac{2\pi}{k} (L_L - L_R) \leq 2k\pi + \frac{\pi}{2}$$  \(8\)

According to the theory of wire antenna, the phase of current on dipole in $y$-axis direction can be considered as a continuous sinusoidal distribution. Thus, a parabolic section of reflector in $zoy$ plane is also introduced to improve the phase uniformity at the aperture. Combining two geometry sections of reflector in $zox$ plane and $zoy$ plane, a concave cavity reflector is designed in Figure 7. The vertical height between the bow-tie dipole and the bottom of the cavity $h_p$ is the key parameter that influences the impedance bandwidth, SDRPB and the profile of antenna. It is determined by the phase equation (5)–(8). The following results in Figure 8 are given to show how the performance of bow-tie dipole with concave reflector is affected by $h_p$.

In Figure 8, with the increase of $h_p$, the gain of dipole with the concave reflector within 7–10 GHz is improved obviously. But the performance of impedance matching is not improved significantly within 2.5–3 GHz, which indicates that the concave reflector behaves not well in low frequency band.

To improve impedance matching, a composite corrugated reflector composed of grooves with hollowed slots was proposed to replace flat reflector in [16]. The size of the corrugated reflector was only optimised without theoretical computation. Here, the equivalent height was introduced to evaluate the composite corrugated reflector.

Firstly, the corrugated reflector is divided into several unit cells in Figure 9a and its reflection phase is simulated in Figure 9b.

To cut off all TE modes in grooves, the width of grooves $d$ need to be less than $\lambda_{11}/4$ [16]. The reference frequency is 10.5 GHz, and $g$ is set to 14 mm. If all TE modes are cut off, it can be equivalent to a ground plane inside the grooves. In another side, the vertical slots are hollowed on the grooves, which means some surface currents flow in the vertical direction, and the vertical currents on the corrugated unit cell are orthogonal to the horizontal currents on the dipole, so that the coupling effect to the above antenna decreases, and deteriorates less compared to the impedance band of the source antenna.

An appropriate value of the duty cycle $(d_r = a_r/p_r)$ is assumed to be one-third. And the cell period $(p_r)$ is suggested to be less than $\lambda_{11}/4$. The height of simulation port is set the same as antenna height $(h_r)$.

Figure 9b shows the reflection phase diagram of the unit cell under the incidence of a $y$-axis directional polarised TE wave illumination. For a given length of hollowed slot $(d_s)$, the reflection phase varies linearly with the frequency. Consequently, the corrugated ground can be equivalent to a PEC ground plane whose equivalent height $(h_r)$ changes with $d_s$.

Furthermore, the relationship between the reflection phase $(p_r)$ can be determined by the electromagnetic wave through double equivalent height and $\pi$ phase shift generated by the equivalent PEC plane.

$$p_r = \frac{2\pi}{\lambda} \cdot 2h_e + \pi$$  \(9\)

Figure 9c shows how $h_e$ varies with the frequency for different $d_s$. For a given $d_s$, $h_e$ is almost frequency independent.
It demonstrates that the corrugated unit cell can be equivalent to a PEC plane. Then, it can be supposed that the relationship between \( h_c \) and \( d_i \) is linear. Thus, a concise expression is obtained by fitting the above results in Figure 9c. As shown in Figure 9c, changing \( d_i \) in corrugated unit cell, the equivalent plane moves between the top plane and the bottom plane.

\[
h_c = 0.9 \cdot d_i + 10
\]

Through the expression of equivalent height of corrugated unit cell, the concave cavity reflector is replaced with corrugated cells. Then a compound concave corrugated reflector (CCCR) is developed in Figure 10.

The performances of the bow-tie dipole with different reflectors are compared in Figure 11. It can be observed that the impedance bandwidth of the bow-tie dipole with the CCCR is close to the dipole without reflector at lower band. In addition, the gain of the dipole with the CCCR is 5dB larger.
than the dipole without reflector, and more uniform within the whole band.

Compared with flat reflector, both CCCR and concave reflector perform well in the whole band especially in 7–10 GHz. It can be concluded that the CCCR can achieve better radiation pattern performance within high frequency band and better impedance matching characteristics within low frequency band. The final optimised parameters of the CCCR are summarised in Table 2.

2.5 | Comparison with different reflectors

To better evaluate the performance of different reflectors, the simulated results of the same dipole with flat reflector (FR), AMC reflector (AMCR), corrugated reflector (CR) in [16] and CCCR in this work, are analysed. And these antennas are designed to have the same profile height of 20.5 mm. Because the benefits to stable directional pattern of CCCR can be shown in high frequency band especially in 6–10 GHz, the 3D patterns are given as shown in Figure 12. Figure 13 shows the reflection coefficient and gain at z-axis direction in 2–10 GHz.

In Figure 12, the patterns of all reflectors perform well in 2–4 GHz. However, the pattern of flat reflector begins to split at 6 GHz. When it works at 8–10 GHz, the main lobe points to two flanks of z-axis. As for AMC reflector, it can achieve excellent reflection at 6 GHz while the beam splits in two lobes at 8 GHz and more lobes appear at 10 GHz. Moreover, the main lobe of CCCR always points to z-axis and its level of side lobe is much lower than the level of split lobes of other reflectors. The detailed analysis of CCCR will be given in the follow section.

According to Figure 13, these four reflectors all perform similarly in impedance matching while the radiation pattern performances are different. It can be summarised that the impedance characteristic of CCCR performs better than the corrugated reflector in 3–4 GHz while the performances of flat and AMC reflector are awfully close because of their identical profile height. Furthermore, the gain of CCCR is much higher and more stable than other reflectors while the gain of other reflectors drops sharply in high band. It further indicates that CCCR performs better than other reflectors to achieve a stable directional radiation.

2.6 | Analysis of aperture E-field

Once the bow-tie dipole is placed above the reflector, it can be regarded as the aperture antenna. Suppose a virtual aperture plane (VAP) above the bow-tie dipole with CCCR, and the directivity of the antenna is determined by the field phase distribution on the aperture.

Since the co-polarisation direction is parallel to y-axis, the phase distributions of the y-component of electric field parallel to the dipole axial direction in zoy plane are shown in Figure 14a. The black dot line represents the VAP. Column I, II and III represent without reflector, flat reflector and CCCR respectively. The rows represent different frequencies.

In Figure 14a, at 2–6 GHz, the phase on the VAP is similar for different cases in column I, II and III respectively. At 8 GHz, the central green region on the VAP in column III is much larger than other two cases. Especially, at 10 GHz, the VAP in column III is almost all located in the green region which means it is an equiphase plane while the phase on the VAP in column I and II periodic changes.

The phase distributions on the VAP in zox plane at different frequencies are shown in Figure 14b, which are similar to the cases in Figure 14a. At 2 GHz, 4 GHz, and 6 GHz, the phase distributions on the VAP central region are close. At 10 GHz, the phase on the VAP in column III is more uniform than that in column I and II.

It is obvious that the field phase uniformity on the VAP in column III is better than column I and II in Figure 14. It means the CCCR can surely improve the phase uniformity on the aperture above the source dipole.

| TABLE 2 | Parameters of the CCCR |
| --- | --- | --- | --- | --- | --- |
| Parameters | $a_1$ | $a_2$ | $b_1$ | $b_2$ | $H$ |
| Value (mm) | 132 | 72 | 108 | 53 | 20.5 |
| Parameters | $g$ | $h_a$ | $w_a$ | $p_1$ | $a_t$ |
| Value (mm) | 32 | 48.8 | 12.5 | 5.5 | 1.8 |

**FIGURE 11** Simulated realised gain and VSWR of the bow-tie dipole with different reflectors.
amplitude of side lobes become larger while the main lobe is the principal part and points to the z-axis direction. When the frequency rises to 10 GHz, the number of side lobes increases as Figure 15c shows. The amplitude of the main lobe rises more significantly than the side lobes. As a result, the shape of pattern at 10 GHz performs better than that at 8 GHz.

The schematic how the CCCR reflects the downward side lobes to the upward direction in high frequency is shown in Figure 16a. Here, assume the coupling between the bow-tie dipole and the reflector is neglected. Thus, the far field radiated by the dipole can be divided into the upward radiation field and downward radiation field. In Figure 16b, the near field on a big enough plane above the dipole is simulated to calculate the equivalent source of upward radiation. The downward equivalent source is simulated in the same way.

Then, the downward equivalent source is placed above the CCCR at the same height as that of the actual dipole. The simulated result of reflected pattern shows that the downward side lobes are reflected and focussed on z-axis direction. Most of the radiation power is concentrated in the main lobe. Furthermore, the final resultant radiation pattern is the superposition of upward radiation pattern and reflected downward radiation pattern. It can be observed the superposition pattern is similar to the simulated pattern in Figure 15c. Moreover, the simulated pattern has lower side lobes than the superposition pattern which indicates the coupling between the bow-tie dipole and the reflector may benefit the performance of radiation patterns.

2.7 Analysis of patterns

To show the improvement of patterns by using the CCCR, the three-dimensional radiation patterns of antenna are analysed in this section. Figure 15 shows the simulated radiation patterns of the dipole with CCCR in 2–10 GHz. In the band of 2–6 GHz, the directional patterns are stable and have no side lobes in Figure 15a–c. When it operates at 8 GHz, the

3 EXPERIMENTAL RESULTS ANALYSIS

3.1 Balun design

In order to feed the bow-tie dipole over a quintuple band, an exponential tapered transformer from a 50 Ω impedance
microstrip line to a 100 Ω impedance double lines is designed. Due to the profile height restriction, the balun needs to be transverse to increase length. The feeding point is connected to horizontal lines by one-fourth arc double lines.

In addition, if the transformer is located in the CCCR, the beam direction of patterns will appear deflection in high frequency. Thus, transformer needs to be placed outside the reflector. To balance the coupling between the bow-tie dipole and double lines with the coupling between double lines and reflector, the radius of one-fourth arc is set to 15 mm.

Although the whole length of balun may be slightly long, the balun can be folded as Figure 17b shows. Thus, the length of balun will not affect the overall size of antenna.

To evaluate the influence of the balun on radiation, the radiation efficiency of the antenna with and without balun is simulated and the results is shown in Figure 17d. It is remarkable that the radiation efficiency of the antenna without balun is approximately equal to 1. The reason is that the whole antenna is almost metal structure except a very thin dielectric substrate of the printed bow-tie dipole. Meanwhile, the dielectric loss in this band is low and the ohmic loss can be also

**FIGURE 14** Simulated E-field phase distribution in zoy and zox planes of the bow-tie dipole with different reflector. “---” represents the virtual aperture plane in each case. (a) Plane zoy. (b) Plane zox.
ignored. Once the long balun is added to achieve impedance match, because of the relatively long electric length at 10 GHz, the dielectric loss of the high frequency band increases compared with the low frequency band which cause the radiation efficiency to drop slightly with the frequency increasing while it is all above 0.9.

3.2 | Reflection coefficient

The measured and simulated reflection coefficient results of the bow-tie dipole with and without CCCR are shown in Figure 19a. It can be observed that the measured impedance band is 2.27–26 GHz, which coincides with the simulated
result 2.34–26 GHz respectively. In fact, the reflection coefficient of the dipole with CCCR is generally below −10 dB over the whole band except a few frequency points such as 4.7 and 9.2 GHz. Meanwhile, the reflection coefficient performs badly at 9.2 GHz which may be generated by the length of double lines being several times of wavelength in high band which causes slight impedance mismatch appears at some frequency points.

It also can be observed that in low band, the measured reflection coefficient rises slightly compared with the simulated results while it is still all below −8 dB, which illustrates that CCCR has excellent performance in impedance matching.

Moreover, because of the excellent performance of balun, the impedance band can even cover 2.27–26 GHz. The simulation results show that When the frequency exceeds 10.7 GHz, the main beam appears to split which makes it unable to be applied to directional application. Thus, the gain and pattern are not measured over 10.5 GHz.

3.3 Radiation patterns

Measured and simulated co-polarisation and cross-polarisation radiation patterns in x0z plane (θ = 90°) of the dipole with the balun on CCCR are both presented in Figure 18a–c. It can be observed that the antenna has directional radiation patterns over the whole band of 2–10 GHz and the measured and simulated results of the co-polar show a good agreement. In 2–10.5 GHz, the main lobe all direct the z-axis direction and there is no beam splitting. Moreover, in the direction of main lobe, the cross-polar amplitude is less than −20 dB.

The beamwidth decreases from low band to high band, because the relative electric size of the whole antenna increases from 0.8λ × 0.72λ to 4.4λ × 3.6λ. Although the side lobe appears peak around 8 GHz, the main lobe points to z-axis over the whole band (2–10.5 GHz).

Figure 18f–j shows the measured and simulated co-polar radiation patterns in x0z plane (θ = 0°). Side lobes appear in x0z plane within 8–10 GHz, because the reflector only reflects the downward side lobes to target direction, while the upward side lobes generated by the bow-tie dipole still exist.

Figure 19b shows the measured and simulated gain of the bow-tie dipole with CCCR. It can be observed that the measured gain agrees well with the simulated gain, and little difference in high frequency band may be influenced of the machining error, installation tolerances and test error. The peaks of gain appear at 5.5 and 10 GHz, because the uniformity of electric field on aperture plane performs better in these two frequency points. In the whole band, the measured gain is larger than 7 dBi and peaks at 12 dBi. Although the gain variation may be slightly large around 7 GHz, it can be improved by adjusting the phase distribution of the aperture plane in the future work.

Table 3 compares the performance of different UWB directional antenna. \( H_2 \) is the profile height of antenna which refers to maximum wavelength. \( B_1 \) represents the relative impedance bandwidth of antenna. \( B_p \) represents the relative bandwidth of stable directional patterns. \( R_{IH} = B_1/H_2 \). \( R_{IH} = B_p/H_2 \). To better assess the performance of antenna, the relative stable directional pattern bandwidth and the ratio of the relative bandwidth to profile height are proposed. It is
**FIGURE 17** Configuration of the balun, photo of the antenna with folded balun and radiation efficiency of the antenna with and without balun. All units in mm. (a) Balun (b) The antenna with folded balun (c) Photo of the antenna (d) Radiation efficiency

**FIGURE 18** Patterns of the bow-tie dipole with CCCR in yoz plane and xoz plane (a)–(e) plane yoz, (f)–(j) plane xoz
with compound concave corrugated reflector (CCCR) has been presented in this article. Because of the inhomogeneous phase of currents on dipole, the radiation patterns of the dipole perform poorly in the high band. The CCCR can achieve more uniform electric field phase on the aperture plane above the dipole to obtain better directional radiation patterns. In addition, the CCCR can also achieve better impedance matching characteristics. According to the measured results, the impedance bandwidth of the antenna is 2.27–26 GHz and the stable directional patterns are achieved from 2 to 10.5 GHz. The gain is over 7 dBi within the band (2–10.5 GHz) and peaks at 12 dBi in 5.5 and 10 GHz. It can be foreseen that this approach can be applied in other UWB directional antenna design and this antenna structure may be applied to various conformal and high gain directional applications.

ORCID
Chaofan Ren https://orcid.org/0000-0003-2056-9504
Junting Geng https://orcid.org/0000-0001-6501-3169

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