A Novel Broadband High-Isolation Cross Dipole Utilizing Strong Mutual Coupling

Zengdi Bao, Xianzheng Zong and Zaiping Nie
University of Electronic Science and Technology of China
NO.4, Section 2, North Jianshe Road
Chengdu, Sichuan 610054 China

***Abstract*-A novel broadband cross dipole with high isolation and low cross polarization is presented in this paper. By skillfully utilizing the strong mutual coupling between the two dipoles, very good impedance match, sufficiently high isolation and low cross polarization can be achieved simultaneously. To get a directional radiation pattern, a metal reflector is placed under the antenna. The simulated measured results show that a more than 56% impedance bandwidth for SWR<1.5 from 1.66GHz to 2.96GHz is achieved, within which the port isolation is larger than 30dB. As an important advantage, the input impedance of the antenna can be changed within certain limits by properly adjusting the configuration, thus the antenna can be fed directly by 50-ohm coaxial cables without using impedance transformers. Particularly, the mechanisms for wide impedance bandwidth and high port isolation are discussed in this paper.**

I. Introduction

With the space resources for base stations decreasing, it is very important to demand that base station antennas can cover a wider frequency range and accommodate to more communication standards so as to avoid the repetitive constructions of telecom equipment. In recent years, the ±45° dual-polarized antennas were widely used to increase the communication capability and quality. Though significant development in the design of ±45° dual-polarized antennas has been observed [1]-[6], it is still difficult to achieve wide impedance bandwidth, high isolation, and low cross polarization simultaneously.

In this paper, we present a new design for dual-polarized antenna and discuss the mechanisms for wide bandwidth and high isolation from new points of view. The proposed antenna works in the frequency range from 1.66GHz to 2.96GHz, which can cover the DCS, PCS, UMTS, and part of LTE bands. And the input impedance of the antenna can be changed by properly adjusting its configurations, thus it can be fed directly by coaxial cables without using impedance transformers. From the structure point of view, the antenna is simple and compact and can be used to form an antenna array for base stations in wireless communications.

II. Antenna Configuration

As shown in Fig. 1 (a), the proposed antenna consists of two orthogonally situated dipoles, a dielectric post, and a metal reflector. If this antenna is used as an element to form an array antenna for base stations, a certain number of elements would be linearly arranged in a reflector which should be some longer than the array itself. In this paper, only one antenna with a truncated reflector is studied. The configuration details of the dipoles and the dielectric post are shown in Fig. 1 (b). The dielectric post is made of Teflon with relative permittivity ε_r=2.1 and is just used to support and fix the two dipoles above the reflector. Two through holes are drilled in the dielectric post, through which coaxial cables pass and then feed the dipoles respectively. On the metal reflector plane, there are also two holes, corresponding with the two above mentioned through holes, which are used to guide the cables behind the reflector. For either dipole, the outer conductor of its feeding coaxial cable is connected to one arm, and the inner connector is connected to the so-called copper connector, the far end of which is jointed with the other arm. The E-planes for dipole1 and dipole2 are φ=45°and φ=135° respectively according to the coordinate system shown in Fig. 1 (a), and correspondingly the H-planes are φ=135°and φ=45° respectively.

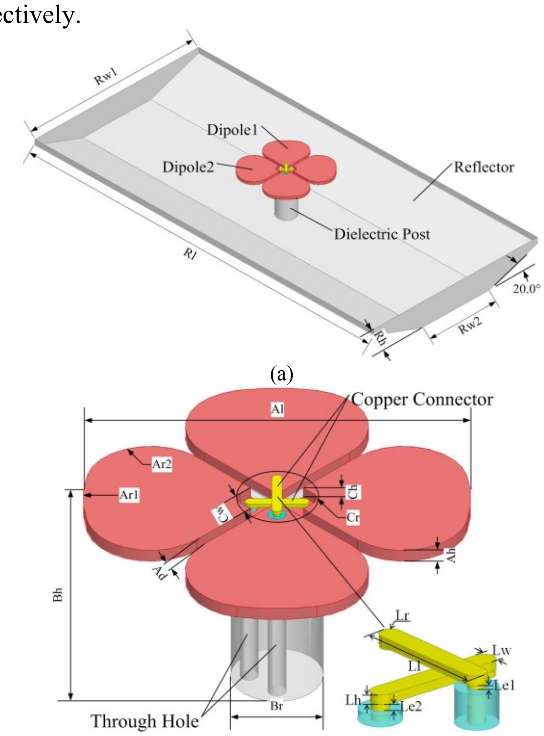


Figure 1. Configuration of the antenna: (a) overall diagram and (b) detail view; *Rl*=300, *Rw1*=145, *Rw2*=60, *Rh*=20, *Ah*=2.2, *Ad*=1.8, *Al*=62.8, *Ar1*=31.4, *Ar2*=10.5, *Br*=7.5, *Bh*=33.4, *Ch*=1.8, *Cw*=4.2, *Cr*=1.8, *Ll*=10.4, *Lw*=1.6, *Lh*=0.7, *Le1*=0.3, *Le2*=0.5, *Lr*=0.8, *Ld*=0.9 (Units: mm).

Since the two dipoles are orthogonally situated, the feeding structures for them have to be properly arranged in order to avoid intersecting. The feeding points of the dipole2 are lowered with a height of *Ch* comparatively with that of the dipole1. This is the only difference between the two dipoles. Since the configurations of the two dipoles are highly symmetrical, the port and radiation characteristics of them are highly symmetrical too.

III. Discussion of Working Mechanism

A. Impedance Bandwidth Enhancement by Coupling

In order to explain the reasons why the antenna we proposed has such a wide impedance bandwidth clearly, a comparison between the dual-polarized antenna (DPA) and the linear-polarized antenna (LPA) was made. The DPA is the antenna we proposed, the LPA is obtained simply by deleting the dipole2 (shown in Fig. 1 (b)) from the DPA while remaining all other configurations unchanged.

Fig. 2 shows the simulated S11 curves of the LPA and the DPA: in the frequency range from 1.65GHz to more than 3.4GHz, the DPA has a much better impedance match performance. Fig. 3 shows the current distribution on the DPA at center frequency 2.31GHz when only one dipole is fed and the other is terminated with a 50-ohm load. It can be found that current distribution with large magnitude appears on the unfed dipole. So there is strong mutual coupling between the two dipoles. As mentioned above, the only difference between the LPA and the DPA is that the LPA has only one dipole, while the DPA has two orthogonal dipoles. Therefore, the reason why the DPA has a much wider impedance bandwidth becomes clear. That the orthogonally situated dipole2 works as a parasitic element for dipole1 greatly extends the impedance bandwidth of the latter: due to the existence of parasitic element, another resonant frequency is introduced. And vice versa, the dipole1 will also extend the impedance bandwidth of the dipole2.

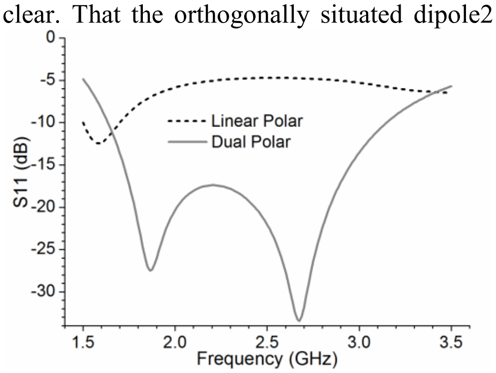
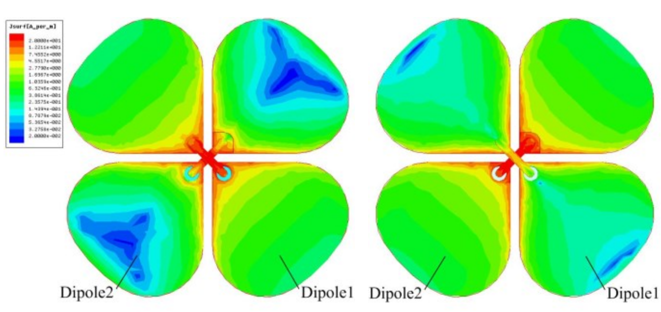


Figure 2. Comparison of the simulated S11 curves of the LPA and DPA.



(a) (b)

Figure 3. Current distribution on the two dipoles at 2.31GHz when (a) only dipole1 is fed and (b) only the dipole2 is fed and the other dipole is terminated with a 50-ohm load.

In broadband antenna design, various bandwidth enhancement techniques based on passive parasitic structures are widely used. However, in this paper, both of the two dipoles have active feed-port, and they are parasitic to each other simultaneously. The effective utilizing of the mutual coupling between the dipoles can improve the bandwidth without the aid of other passive parasitic structures, which is helpful to make the design more compact.

B. Isolation Analysis

As analyzed above, with the aid of the strong mutual coupling between the two dipoles, the impedance bandwidths of both of them are greatly extended. In traditional designs, strong mutual coupling between the two dipoles will lead to poor isolation between them. However, high isolation still can be achieved in this design. See Fig. 4, apparently, the current distribution on the dipole2 is coupled from the dipole1. For convenience, we call the currents on dipole1 *Excited Current Vectors* (ECV), and call the currents on dipole2 *Coupled Current Vectors* (CCV). The distribution is extracted at t=T/8, where T is the length of one oscillation period. As illustrated in Fig. 4, the CCV distribution on the two arms of the dipole2 are strictly symmetrical with respect to the center of the dipole2, which means that there is no potential difference between the two feeding points of the dipole2. Therefore, the induced currents on the dipole2 cannot enter the feeding port. Vice versa, when the dipole2 is fed, the induced currents will not enter the feeding port of the dipole1 for the same reasons.

Though only the current distributions at t=T/4 is presented, distributions at other time are similar: the CCV distribution on the dipole2 is strictly symmetrical with respect to the center of the dipole2 all the time. So, high isolation between the dipole1 and the dipole2 can be achieved in the entire period. Though only the case at 2.31GHz is discussed, current distributions at other frequencies are similar. Hence, sufficiently high isolation can be achieved in the entire working frequency range.

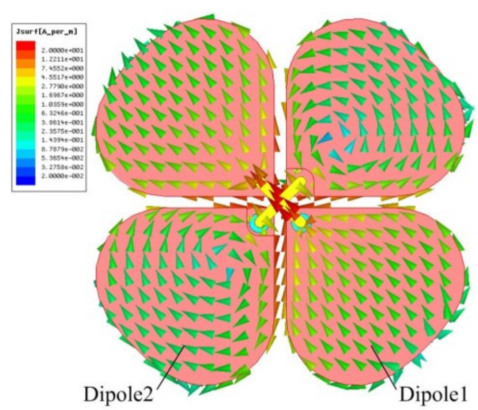


Figure 4. Vector-form current distribution on the two dipoles at 2.31GHz when only the dipole1 is fed and the dipole2 is terminated with a 50-ohm load.

IV. PARAMETER STUDY

Since dipole1 and dipole2 have highly symmetric port and radiation characteristics, only dipole1 is studied. When one parameter is studied, all the other uncorrelated parameters keep constant. The results of these studies offer a useful guideline for practical design.

A. *Fillet Radius Ar2*

Shown in Fig. 1 (b), *Ar2* is the Fillet radius on the arms of the two dipoles. *Ar2* is more sensitive in effects on impedance match than the others parameters. As shown in Fig. 5, it influences the port isolation slightly, but can make the impedance match change remarkably. The upper resonant frequency moves from 2.5GHz to 3.05GHz as *Ar2* increasing from 9.0mm to 12.0mm and a wider impedance bandwidth can be achieved. However, the impedance match in the middle part of the frequency band becomes worse at the same time, which must be taken into consideration when adjusting *Ar2* to get a wider impedance bandwidth.

B. *Arm Spacing Distance Ad*

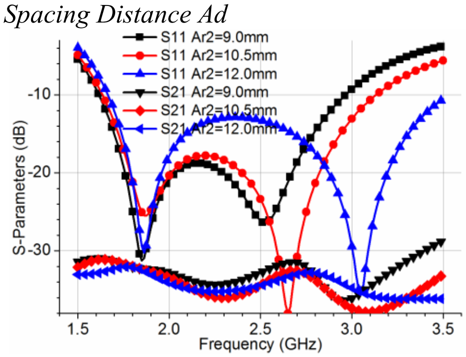


Figure 5. Effects of *Ar2* on S11 and S21 of the antenna.

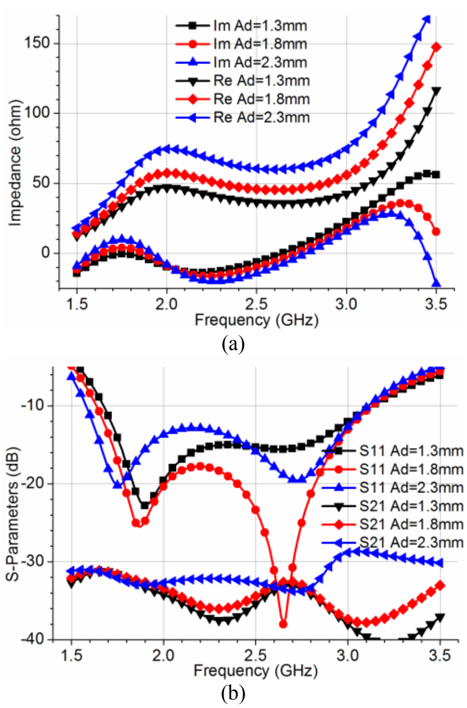


Figure 6. Effects of *Ad* on (a) input impedance and (b) S11 and S21 of the antenna.

By changing *Ad*, *Ah*, or the angel of the two parallel adjacent sides of two arms, the input impedance of the antenna can be changed. For instance, Fig. 6 (a) shows the input impedance of the dipole1 with varied *Ad*. Shown in Fig. 1 (b), *Ad* is the spacing distance between the two parallel adjacent sides. See Fig. 6, the impedance real-part curve moves upward remarkably when *Ad* increases. For *Ad*=1.8mm case, in the working frequency band from 1.66GHz to 2.96GHz, the impedance real-part is closer to 50 Ohms and the imaginary-part is closer to zero, which is helpful to get a better match with general coaxial cables. By properly adjusting the configuration details, the input impedance can be changed conveniently, which indicates the antenna can be fed directly by coaxial cables without using impedance transformers. For this reason, the antenna has the advantage of compactness. Fig. 6 (b) shows S11 and S21 of the antenna corresponding to Fig. 6 (a). The S11 curves vary remarkably when *Ad* is changing. The upper resonant frequency around 2.7GHz disappears when *Ad* is too small, which should be avoided.

V. SIMULATED AND MEASURED RESULTS

To verify the design, a prototype of the proposed antenna was fabricated and measured. The photograph of the fabricated antenna is shown in Fig. 7. Limited by the manufacture conditions, the manufacture precision of the antenna is not very good. The port characteristics and the radiation characteristics of the antenna are measured by Agilent E5071C network analyzer and SATIMO StarLab antenna near field measurement system respectively.

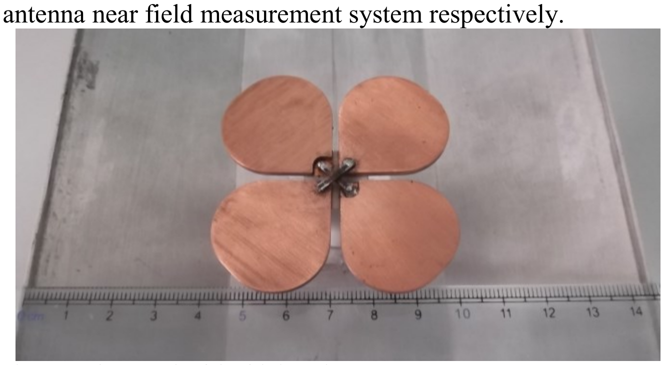


Figure 7. Photograph of the fabricated antenna

A. *Port Characteristics*

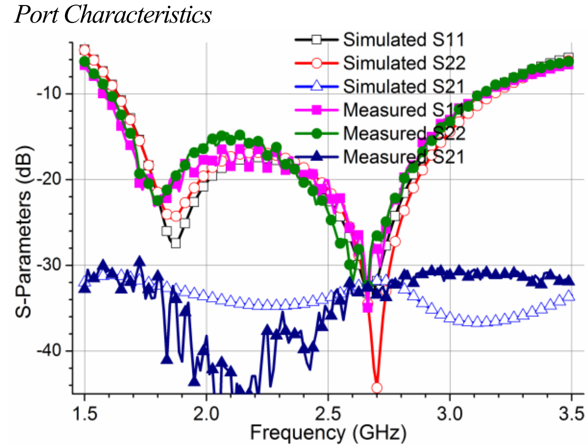


Figure 8. Simulated and measured S-Parameters of the antenna.

Fig. 8 shows the simulated and measured S-Parameters of the antenna. It can be observed that the simulated and measured impedance matches agree very well. The simulated and measured S21 do not agree very well, but both of them are less than -30dB across the working frequency band. The measured results show that more than 56% impedance bandwidth from 1.66GHz to 2.96GHz with SWR<1.5 and isolation>30dB (i.e., S11, S22<-14dB and S21<-30dB) is achieved for both dipole1 and dipole2 of the antenna. Very good impedance match and sufficiently high isolation are achieved.

B. Port Characteristics

The dipole 2 is terminated with a 50-ohm load when the radiation patterns of the dipole 1 are being tested, and vice versa. Since the radiation patterns of the two dipoles are highly symmetrical, only the radiation patterns of the dipole1 are shown here. As illustrated in Fig. 9 (a)-(f), the measured and simulated radiation patterns agreed well except for the cross polarization patterns and backward co-polarization patterns at 1.66GHz. The peak gain of the antenna at 2.31GHz is about 8.3dBi and varies within $\pm$0.5dBi across the working frequency range. The 3dB beamwidth varies from 58.2° to 72.6° in E-plane, and varies from 75.2° to 84.3°in H-plane across the working frequency range. Relatively stable radiation patterns are achieved. Larger than 25dB front-to-back ratio is achieved at 2.31GHz and 2.96GHz. And at 0° direction, the cross polarization at 2.31GHz is lower than -28dB and lower than -24dB at 2.95GHz. Low cross polarization is achieved.

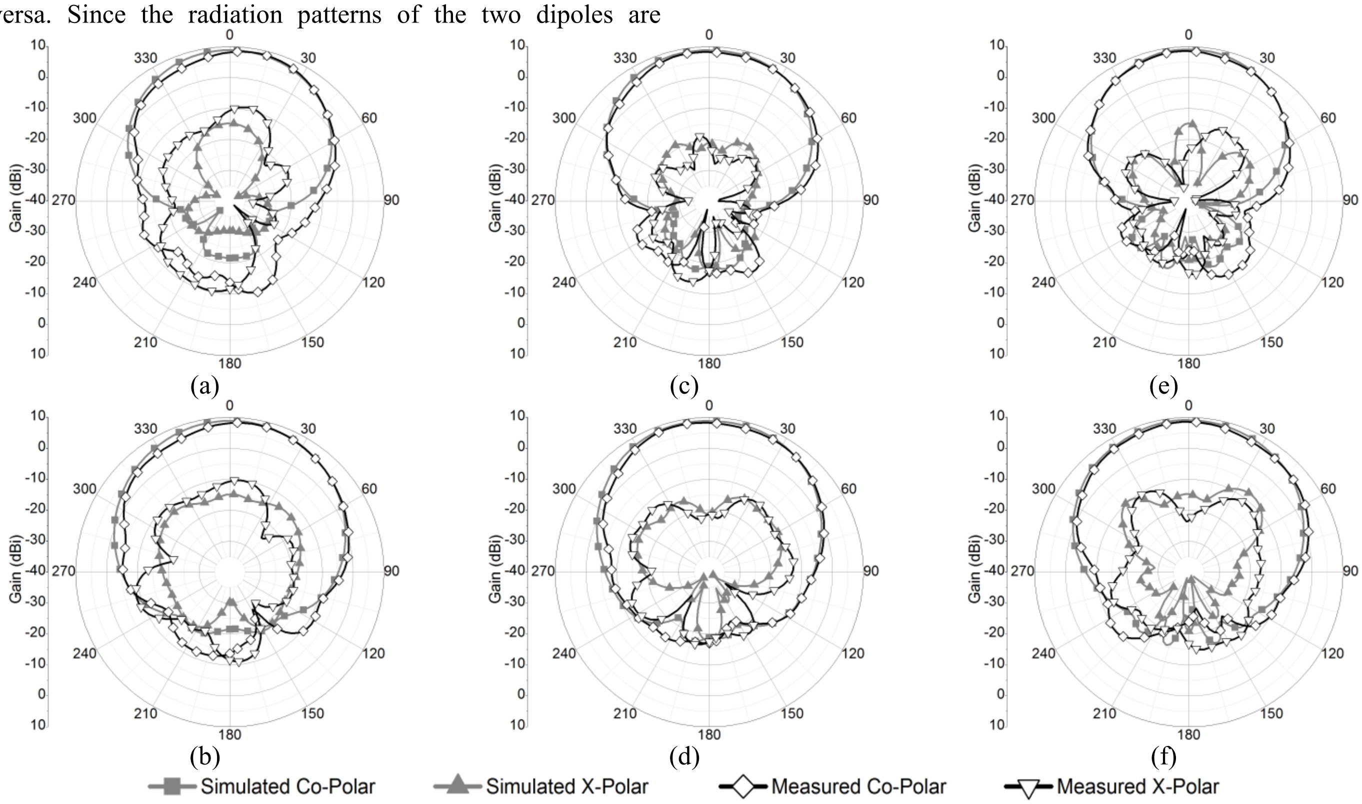


Fig. 9. Simulated and measured radiation patterns of dipole1 at (a) 1.66GHz E-plane, (b) 1.66GHz H-plane, (c) 2.31GHz E-plane, (d) 2.31GHz H-plane, (e) 2.96 E-plane, (f) 2.96GHz H-plane.

VI. Conclusion

In this paper, a broadband cross dipole with high isolation and low cross polarization is proposed. It has a compact and simple structure and can be fed directly by coaxial cables without using impedance transformers. The presented design can be used to form an antenna array for base stations in mobile wireless communications. The wide impedance bandwidth and high port isolation mechanisms are also discussed in this paper, which are also appropriate for other antennas with similar patterns.

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