

A Novel Differentially-Fed Dual-Polarized Filtering Antenna for Base Station

Xuekang Liu

School of Engineering and Digital Arts
University of Kent
Canterbury, UK
xl255@kent.ac.uk

Steven Gao

School of Engineering and Digital Arts
University of Kent
Canterbury, UK
s.gao@kent.ac.uk

Benito Sanz-Izquierdo

School of Engineering and Digital Arts
University of Kent
Canterbury, UK
b.sanz@kent.ac.uk

Haiwei Zhang

Huawei Technologies Ltd
China
zhanghaiwei1@huawei.com

Abstract— The theory, design and experiments of a wideband dual-polarized filtering antenna are presented. A novel technique, which combines the use of open loops, parasitic loop, and extended ground plane, is proposed to realize two upper band radiation nulls and a lower band radiation null. Thanks to these radiation nulls, the proposed antenna obtains a sharp cut-off at the edges of the operating band. To achieve wideband performance, another novel technique of exciting multiple resonant modes of the crossed dipoles and parasitic loop is proposed. The measured results demonstrate that this antenna achieves a wide impedance bandwidth from 1.7 to 3.01 GHz (56%), high isolation of 38dB, low cross-polarization within the working frequency band and also filtering performance. At 3.17 GHz, the gain of the antenna will drop rapidly to -18.1 dBi, which is 26.4 dB lower than the average in-band gain. These advantages make the proposed antenna a good candidate for base station applications.

Keywords—Dual-polarized, base station, differentially-fed, filtering antenna

I. INTRODUCTION

Base stations must support an increasing number of antennas operating on diverse frequency bands as wireless communication develops. The interference between these antennas have a big influence on antenna performance. Filters and antennas are often designed independently and then connected to obtain out-of-band rejection and to reduce antenna coupling [1]. This increases the total size of the antenna system in an already limited base station's space. To overcome this challenge, many strategies for designing integrated filtering antenna have been proposed [2]-[10].

In [2], a differentially fed dual-polarized filtering antenna is designed by etching slots on the radiator and introducing integrated resonant filters on the feeding transmission lines. The out-of-band rejection level of this antenna is higher than 17.7 dB. However, the impedance bandwidth of this antenna is relatively narrow. By employing stack patch structure, a radiation null appears at the upper out-of-band of the filtering antenna in [3]. Then, shorted patches and cross-shaped strips are utilized to obtain another two additional radiation nulls. As a result, a wide bandwidth of 20% has been realized. In [7], by etching defected ground structures (DGSs) and introducing shorting pins under the rectangular patch, a lower band radiation null can be realized. Then, to enhance the upper out-of-band rejection level, cross slots are etched on the radiator to introduce a new radiation null. The out-of-band radiation level is 20dB lower than the in-band gain. Recently, a new

method of designing filtering antenna is presented in [8]. By exciting resonant modes at the out-of-band, three radiation nulls are obtained at lower and upper out-of-band. This antenna has an impedance bandwidth of 49% (1.7 GHz - 2.8 GHz). By simply utilizing U-shaped parasitic elements, a dual-polarized filtering antenna with a wide impedance bandwidth of 63% is proposed in [9] based on the traditional crossed-dipoles.

An innovative wideband dual-polarized antenna with filtering response is designed in this paper. Unlike earlier approaches, open loops, parasitic loops, and extended ground plane are used to produce the three radiation nulls in this design. It is worth noting that all the radiation nulls in this design are controllable. This is crucial when designing a filtering antenna. Furthermore, the antenna achieves a 56 percent impedance thanks to the innovative usage of multiple resonant modes. It signifies that the proposed antenna can cover all of the 2G/3G/LTE communication systems' frequency bands. Apart from the advantages listed above, the proposed antenna has a simple structure which is simple to be fabricated.

II. ANTENNA THEORY AND DESIGN

A. Configuration

Fig. 1 shows the proposed differentially-fed filtering antenna layout. As depicted, the proposed antenna includes two Rogers 4003 substrates (relative dielectric permittivity $\epsilon_r = 3.55$, loss tangent $\tan\delta = 0.0027$). The thickness of the substrates are 0.813mm. The radiator is printed on the upper layer of the substrate1. The ground plane is printed on the lower layer of the substrate2. The radiator has three parts: the transformed crossed-dipoles, the parasitic loop, and the extended ground plane. As illustrated in Fig.1, the outer conductors of the coaxial cables connect the extended ground plane to the main ground plane on substrate 2. The crossed-dipoles are connected to the inner conductor of the coaxial cables. The S-parameters of the differentially-fed antenna are calculated in this paper by employing:

$$S_{dd11} = (S_{11} + S_{33} - S_{13} - S_{31})/2 \quad (1)$$

$$S_{dd22} = (S_{22} + S_{44} - S_{24} - S_{42})/2 \quad (2)$$

$$S_{dd21} = (S_{21} + S_{43} - S_{23} - S_{41})/2 \quad (3)$$

$$S_{dd12} = (S_{12} + S_{34} - S_{14} - S_{32})/2 \quad (4)$$

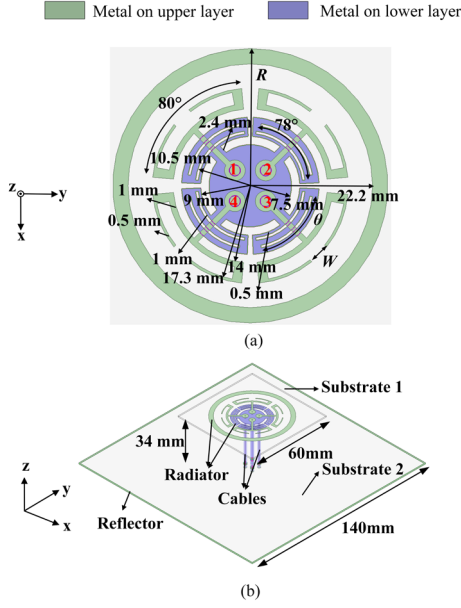


Fig. 1. (a) Top view of the proposed antenna. (b) 3D view of the proposed antenna. Dimensions are $W = 3.8$, $R = 25$, $\theta = 70^\circ$ (unit: millimeter).

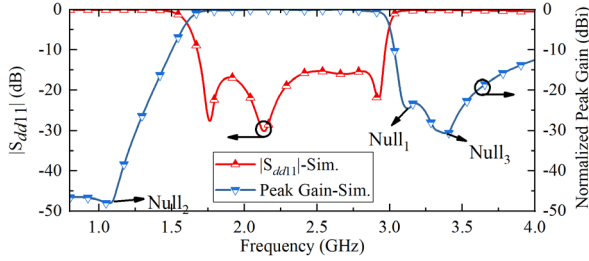


Fig. 2. Simulated reflection coefficient and peak gain of the proposed antenna.

where S_{ij} ($i = 1, 2, 3, 4$; $j = 1, 2, 3, 4$) are the single-ended S-parameters.

B. Radiation Nulls and Resonant Modes

Fig.2 plots the simulated reflection coefficient and peak gain of the proposed antenna. As can be observed, the proposed antenna has a wide impedance bandwidth and good out-of-band rejection level.

Simulated input impedance of the traditional crossed-dipoles (Ant.1) is given in Fig.3 (a). The two resonant modes of Ant.1 resonate at 1.8 GHz and 2.6 GHz. To introduce radiation null₁, the arms of the Ant.1 are simply substituted by the open loops. After the replacement, a miniaturised antenna (Ant.2) with a radiation null in the upper out-of-band can be obtained, as illustrated in Fig.4. It can be observed in Fig. 3(b) that the replacement has nearly no effect on the resonant frequencies of the original resonant modes of the Ant.1. However, a null will appear at 3.2 GHz on the input resistance curve of Ant.2. To further illustrate the rationale of this radiation null, the simulated current distribution of Ant.2 at 3.2 GHz is shown in Fig. 5(a). It can be observed that the current distributions of the open loops on dipole 1 and dipole 2 are shown to be completely different. At this frequency, the open loops of dipole 1 can be seen as feeding structures, and their counterparts on dipole 2 act as half-wavelength resonators. So, the open loops on dipole 2 will resonate at 3.2 GHz when dipole 1 is excited. All the input power will be reflected back to the input port instead of radiating to the space at this frequency. The resonant frequency of the null₁ can be approximately calculated by utilizing:

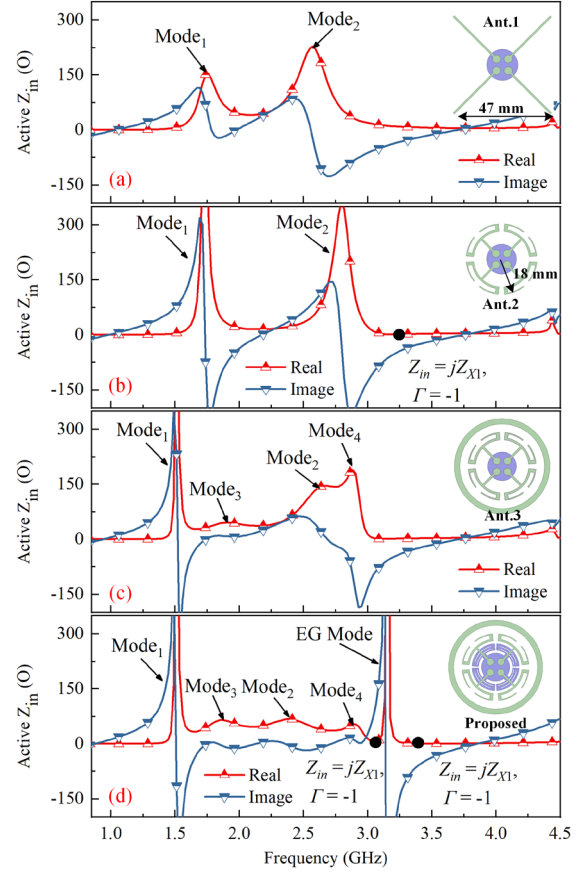


Fig. 3. Simulated input impedances of (a) Ant.1, (b) Ant.2, (c) Ant.3, and (d) the proposed antenna.

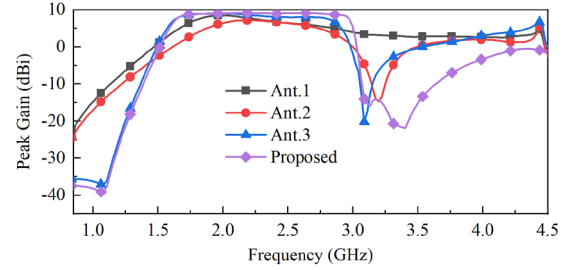


Fig. 4. Simulated peak gains of (a) Ant.1, (b) Ant.2, (c) Ant.3, and (d) the proposed antenna.

$$f_{null,1} \approx \frac{c}{2L_{eff}} \quad (5)$$

Where L_{eff} is the effectively length of the open loop. c is the speed of light in free space.

A parasitic loop is then used to suppress the radiation in lower out-of-band while also expanding the impedance band. As shown in Fig. 4 and 3 (c), a radiation null and two new resonant modes can be obtained at 1.1 GHz, 1.9 GHz and 2.9 GHz, respectively. Fig. 5 (b), (c), and (d) show the simulated current distributions of Ant. 3 at these frequencies. At 1.1 GHz, the current vector on the parasitic loop is in the opposite direction of the dipole arms. As a result, the dipoles' far-field radiation will be neutralised, resulting in a radiation null. As observed in Fig. 5(c), due to the exciting of the loop mode of the parasitic loop, a new resonant mode occurs at 1.9 GHz. For the resonant mode at 2.9 GHz, it is caused by the coupling

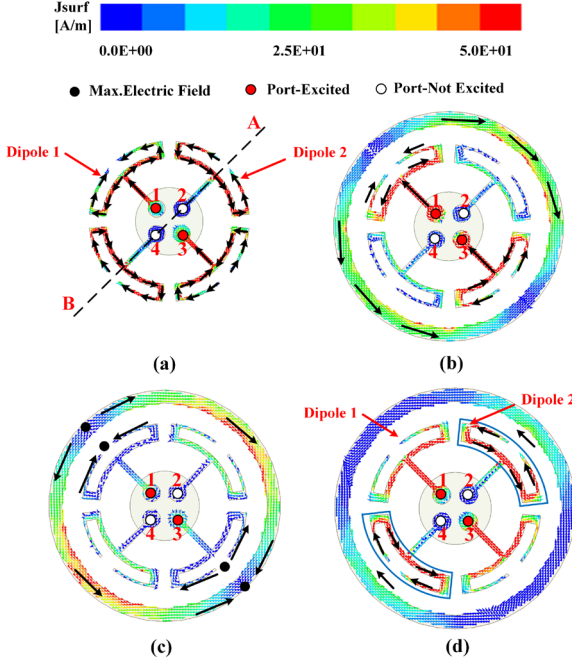


Fig. 5. Simulated current distributions of (a) Ant.2 at 3.2 GHz, (b) Ant. 3 at 1.1 GHz, (c) Ant. 3 at 1.9 GHz, and (d) Ant. 3 at 2.9 GHz.

between the open loops and parasitic loop. Thanks to the coupling, the open loops can be equal to unbroken rings resonating at 2.9 GHz. The resonant frequency of the loop mode (Mode₃) can be calculated by utilizing:

$$f_L \approx \frac{c}{\pi R_{eff}} \quad (6)$$

where R_{eff} and is the effective radius of the parasitic loop.

To further suppress the radiation in upper out-of-band, a new radiation null (Null₃) is realized by extending the ground plane under the dipole arms and etching slots on it. As can be observed in Fig. 3 (d), a new null appears at 3.4 GHz on the curve of input resistance (refer to input ports). The reflection coefficient at this frequency is equal to -1. As shown in Fig. 6, the simulated current concentrates near the slots on the extended ground plane, it is impossible to transfer transmission power to open space.

C. Parametric Study

To have a better understanding of the proposed filtering antenna's principles of operation. In this part, some crucial variables are examined. While one parameter is being investigated, the others are remain constant.

Fig. 7 depicts the proposed antenna's simulated peak gains under various W (width of the gap of the open loop) values. As the width increases, the radiation null₁ moves to a higher frequency range. When $W = 1.8$ mm, the radiation null₁ appears at 2.93 GHz. There will be a peak between the radiation null₁ and null₃. However, when $W = 5.8$ mm, the frequency of radiation null₁ will equal to the one of radiation null₃. So, $W = 3.8$ mm is chosen in this design.

The simulated peak gains of the proposed antenna at various parasitic loop outer radius (R) values are given in Fig. 8. The outer radius of the parasitic loop has a substantial influence on the frequency of the radiation null₂, as can be observed. The radiation null₂ shifts from 1.17 GHz to 0.97

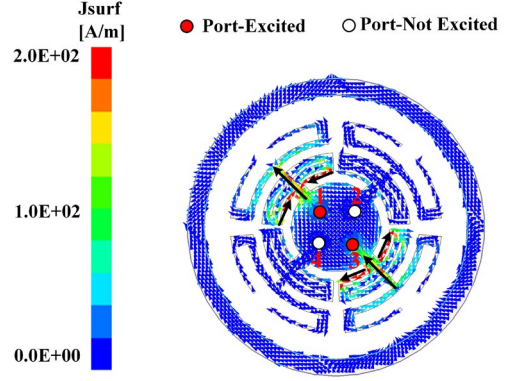


Fig. 6. Simulated current distribution of the proposed antenna at 3.4 GHz.

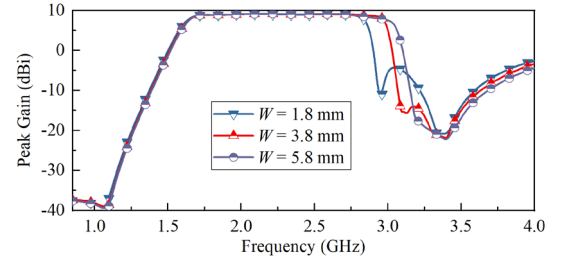


Fig. 7. Simulated peak gains of the proposed antenna under different W .

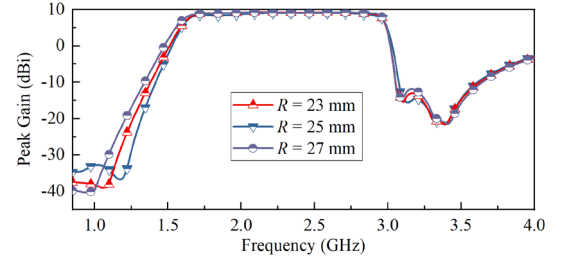


Fig. 8. Simulated peak gains of the proposed antenna under different R .

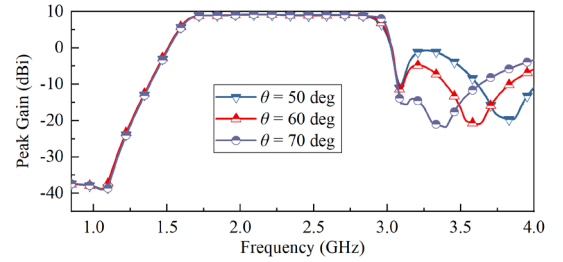


Fig. 9. Simulated peak gains of the proposed antenna under different θ .

GHz when R increase from 23 mm to 27 mm. $R = 25$ mm is the final decision after considering the out-of-band rejection level and roll-off rate.

Fig. 9 depicts the proposed antenna's simulated peak gains of under three θ values. As can be seen, the frequency of radiation null₃ can be altered by changing the length of the slots. When $\theta = 50^\circ$, the radiation null₃ appears at 3.85 GHz. With the increase of the length of slots, the frequency of radiation null₃ shifts towards lower frequency band. When $\theta = 70^\circ$, the two upper band radiation nulls are close enough to achieve a satisfactory out-of-band rejection level.

III. RESULTS AND DISCUSSION

To verify the design principle, the proposed filtering antenna was fabricated and measured. The antenna's S-

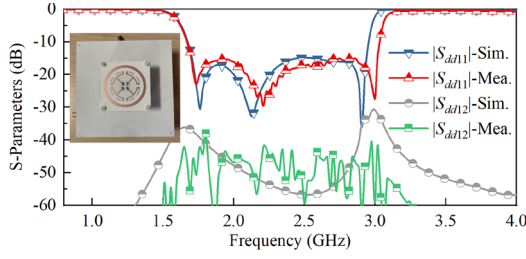


Fig. 10. Measured and simulated S-parameters.

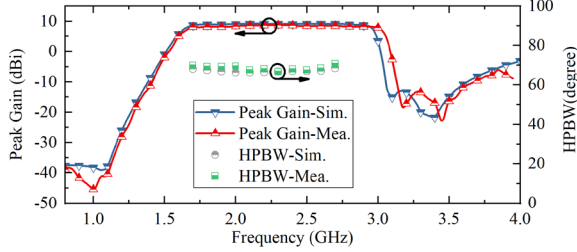


Fig. 11. Measured and simulated peak gain and HPBW.

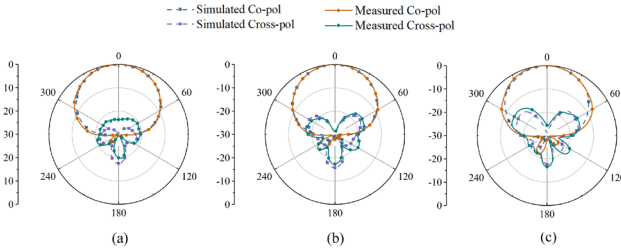


Fig. 12. Measured and simulated normalized radiation patterns at (a) 1.7 GHz, (b) 2.2 GHz, and (c) 2.7 GHz.

parameters are measured using a vector network analyser with 24 ports. As shown in Fig. 10, the measured S-parameters are in good agreement with the simulated values. The measured results demonstrate that the proposed antenna has a 56 percent impedance bandwidth and a 38 dB isolation. The far-field performance of the filtering antenna was measured in the anechoic chamber and is given in Fig. 11. As depicted, the average peak realized gain of the fabricated antenna is about 8.3 dBi. In the out-of-band, the radiation nulls can be clearly observed and the peak gains have been effectively restrained. The results in Fig. 12 are an comparison of measured and simulated normalized radiation patterns of the proposed antenna at 1.7, 2.2, and 2.7 GHz. As can be seen, the proposed antenna has very stable radiation patterns within the operating band.

In table I, the proposed antenna is compared to various dual-polarized filtering antennas. In comparison to the antennas in [4] and [7], the proposed antenna achieves a wider bandwidth, higher gain and higher isolation under a smaller size. The antenna in [9] obtains a wide bandwidth of 69.3% for $|S_{11}| < -10$ dB. However, the gain variation of this antenna is 1.1 dB, which is 0.7 dB higher than the proposed antenna.

IV. CONCLUSION

An innovative wideband dual-polarized filtering antenna is designed. The working principle of the proposed antenna is explained and demonstrated in the paper. As illustrated, the three radiation nulls are mainly introduced by the open loops, parasitic loop, and extended ground plane. By optimizing the value of the critical parameters, these controllable radiation nulls can be properly allocated to realize a good out-of-band

TABLE I
COMPARISON OF THE PREVIOUSLY PRESENTED DUAL-POLARIZED
FILTERING ANTENNAS

Ref.	Imp. BW	Size (mm)	Filtering Technique	Gain (dBi)	Iso. (dB)
[4]	1.65-2.75 (50%) ($ S_{11} < -15$ dB)	55×55	Modified balun	~8.1	26
[8]	1.7-2.8 (49%) ($ S_{11} < -10$ dB)	61×61	Inherent modes	~8.0	35
[10]	1.81-3.73 (69%) ($ S_{11} < -10$ dB)	60×60	Slots	8.1±1.1	30
This Work	1.70-3.01 (56%) ($S_{d11} < -15$ dB)	50×50	Open loops, parasitic loop, and extended ground	8.3±0.4	38

rejection. The measured results show that the antenna not only achieves a high level of out-of-band suppression, but also a wide bandwidth. Furthermore, the coupling between the input ports is lower than -38 dB. A dual-polarized antenna with such a bandwidth, high gain rejection level, and high isolation is a good candidate for base station use.

ACKNOWLEDGMENT

The authors are grateful to Professor Xiaodong Chen, Mr. Long Qian from Queen Mary University of London and Mr. Antonio Mendoza from University of Kent for their help in antenna measurement.

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