

Broad-Band Radial Slot Antenna Fed by Coplanar Waveguide for Dual-Frequency Operation

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Abstract—A novel design of a broad-band radial slot antenna fed by a coplanar waveguide for dual-frequency operation is presented. Various frequency ratios, within the range of about 1.3 to 2.1, of the two operating frequencies can be obtained by varying the included angle between the radial slots and/or by varying the length of the central slot pair. The bandwidth of the lower operating band lies in the range of 2.4% to 9.7%, while that of the upper band is much wider and ranges from 17.4% to 23.2%. Radiation patterns at the two operating frequencies are broadside and bidirectional. Details of the antenna design and the experimental results are presented and discussed.

Index Terms—Coplanar waveguides, multifrequency antennas, slot antennas.

I. INTRODUCTION

THERE have been growing research activities on dual-band and multiband conformal antennas for integrations of heterogeneous communication systems such as the global system for mobile communication (GSM), the general packet radio service (GPRS), the global position system (GPS), the wireless local area networks (WLAN), and the universal mobile telecommunications system (UMTS). Among them, the microstrip patch antenna and the slot antenna are two of the most prevailing antenna structures. Several different designs of dual-frequency microstrip patch antenna have been proposed previously, such as the stacked microstrip antenna [1], the reactively loaded microstrip antenna, the dual-mode (operated at TM_{10} , TM_{01} or other higher order modes) microstrip patch antenna [2]–[5], and the planar inverted-F antenna [6]–[8]. These designs, however, show high-Q characteristics in both operating bands with their bandwidths in the range of about 2% due to the microstrip configuration. Instead, slot antennas have broader impedance bandwidths than microstrip patch antennas. Moreover, slot antennas radiate bidirectionally while microstrip patch antennas radiate in one direction. In some wireless and spatial power combing applications, bidirectional antennas are more useful [9]. In spite of these advantages, there are fewer slot-based dual-band antennas proposed so far. Slot ring and slot loop antennas were commonly the fundamental structures used in [10]–[12], and the slot spiral antenna was proposed in [13].

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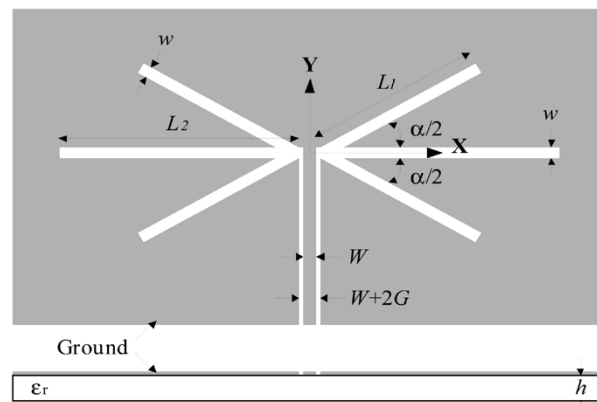


Fig. 1. Geometry of the CPW-fed radial slot antenna.

In this paper, a novel coplanar waveguide (CPW) fed radial slot antenna is proposed for dual-frequency operation. The frequency ratio of the two operating frequencies can be controlled by varying the included angle between the radial slots. Fine-tuning of the frequency ratio can be further obtained by varying the length of the central slot. The frequency ratio of the proposed design is tunable in the range of 2.10 to 1.30, while that of the slot antenna discussed in [12] ranges from 1.47 to 1.10. The two operating bands of the proposed design have similar radiation properties, including their polarizations and patterns, and the measured impedance bandwidths for the lower and upper operating bands are wide and range from 2.4 to 9.7% and 17.4 to 23.2%, respectively. In the upper operating band, not only the bandwidth is wider, but the cross-polarization radiation levels in both principle planes are far lower than those of the slot ring and the slot loop antennas proposed in [10] and [11], respectively. Also, the uni-planar and simple structure of the proposed design make it ease of mass production and more cost-effective than the cavity-backed slot-spiral antenna presented in [13]. Details of the design are presented. Experimental and simulated results of the obtained dual-frequency characteristics are also presented and discussed.

II. ANTENNA CONFIGURATION AND OPERATION

The configuration of the CPW-fed radial slot antenna is shown in Fig. 1. The central pair of the slots with dimension of $L_2 \times w$ is normal to the CPW feedline. All of the other four oblique slots having the same dimension of $L_1 \times w$ are oriented to have an angle $\alpha/2$ measured from the central slot. The widths of the strip and slot of the feeding CPW are W and G , respectively. As shown in Fig. 1, the geometry symmetric to

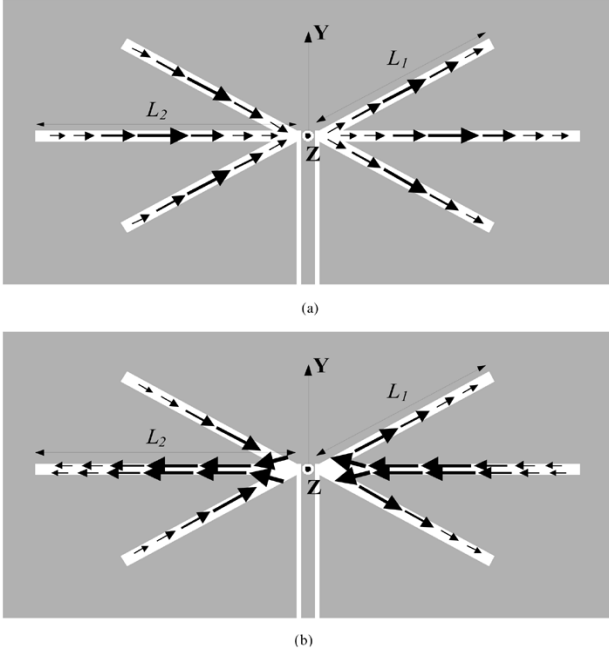


Fig. 2. Equivalent magnetic current distributions on radial slot antenna with $L_1 < L_2$ and operating at (a) f_H and (b) f_L .

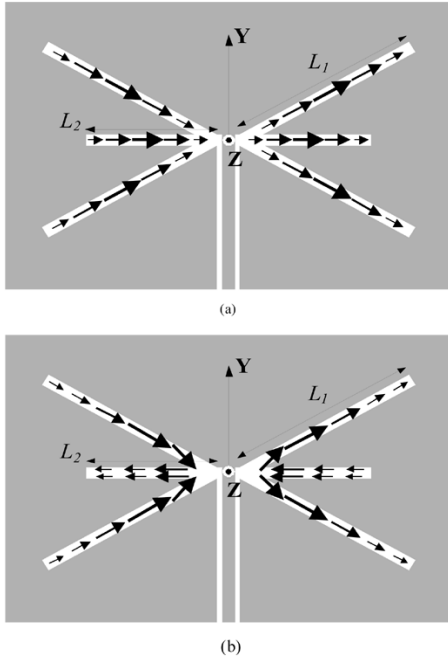


Fig. 3. Equivalent magnetic current distributions on radial slot antenna with $L_1 > L_2$ and operating at (a) f_H and (b) f_L .

the x - and y -axes results in less complexity during the design process, and the uniplanar structure makes it ease of fabrication.

The principles of the dual-frequency operation of this antenna can be observed via Figs. 2 and 3. These figures are obtained from the electromagnetic simulator Ansoft Ensemble.¹ The instantaneous equivalent surface magnetic current distributions of the radial slot antennas at resonance with $L_1 < L_2$ and $L_1 > L_2$

are shown in Figs. 2 and 3, respectively. As shown in Fig. 2 ($L_1 < L_2$), two resonant frequencies f_H and f_L exist, where f_H and f_L stand for the higher and lower resonant frequencies, respectively. When operating at f_H [Fig. 2(a)], more equivalent magnetic currents are flowing on the oblique slots and the resonant frequency f_H is mainly determined by L_1 . L_1 is slightly smaller than half the guided-wavelength $\lambda_g/2$ in the structure and L_2 is larger than $\lambda_g/2$. All the magnetic currents on the radial slots can be decomposed into x - and y -components. In Fig. 2(a), the in-phase x -components produce a broadside bidirectional radiation pattern at resonance, while the y -components cancel out each other along the x and y -axes which result in a low cross polarization radiation level in both principle planes ($\phi = 0^\circ$ and $\phi = 90^\circ$). The cross-polarization level will reach its maximum in the plane of $\phi = 45^\circ$ but remain acceptable. When operating at f_L [Fig. 2(b)], L_1 plus L_2 equals to $\lambda_g/2$, and more magnetic current will distribute on the central slot pair for L_2 longer than L_1 . Although the x -components of the magnetic current distributed on the oblique slots are out-of-phase with those of the central slot pair, the amount of these components can only slightly lower the antenna gain but will not affect the radiation pattern enormously. Thus, the pattern distortion of the co-polarization components in other oblique planes different from E and H planes is negligible.

In Fig. 3(a), where $L_1 > L_2$ and the antenna is operating at f_H , L_1 is slightly longer than $\lambda_g/2$ and L_2 is shorter than $\lambda_g/2$. The resonant frequency f_H is mainly determined by L_1 . The radiation pattern resembles that of the case in Fig. 2(a). On the other hand, Fig. 3(b) shows that, when operating at f_L , where L_1 plus L_2 equals to $\lambda_g/2$, there is more magnetic current distributed on the oblique slots than on the central slot pair, and the x -component of the magnetic current on the oblique slots will cancel out with that on the central slot pair and results in a poor radiation pattern. Therefore, in the present proposed design, L_2 is recommended to be longer than L_1 to ensure satisfactory radiation patterns at both resonant frequencies. Furthermore, the case with $L_1 \approx L_2$ is a broad-band design with L_1 and L_2 both approximately equal to half the guided-wavelength at resonance. The x -components of the magnetic current on all slots are in phase to have a broadside and bidirectional radiation pattern, while the y -components cancel out each other to ensure a low cross polarization level. Note that the wide bandwidth nature of this case with $L_1 \approx L_2$ can be further increased by letting $L_1 < L_2$ or $L_1 > L_2$. However, as described above for Figs. 2 and 3, there would be an additional lower frequency band occurs when the difference between L_1 and L_2 becomes significant. L_1 is approximately equal to half the guided-wavelength at the inherent resonant frequency of the design with $L_1 \approx L_2$, and this is also the case for the upper resonant frequency f_H in Figs. 2(a) and 3(a). In addition, the magnetic current distributions at resonance resemble those illustrated in Figs. 2(a) and 3(a). Thus, the upper operating band near f_H of the dual-frequency operation in Figs. 2(a) and 3(a) corresponds to the inherent operating band of the broad-band design with $L_1 \approx L_2$, hence the bandwidth of the upper operating band near f_H is much wider than that of the lower band near f_L . It is worth mentioning that the enhanced bandwidth of the upper band of the proposed design

¹Ansoft Ensemble is a registered trademark of Ansoft Corporation, Pittsburgh, PA 15219-1119 USA.

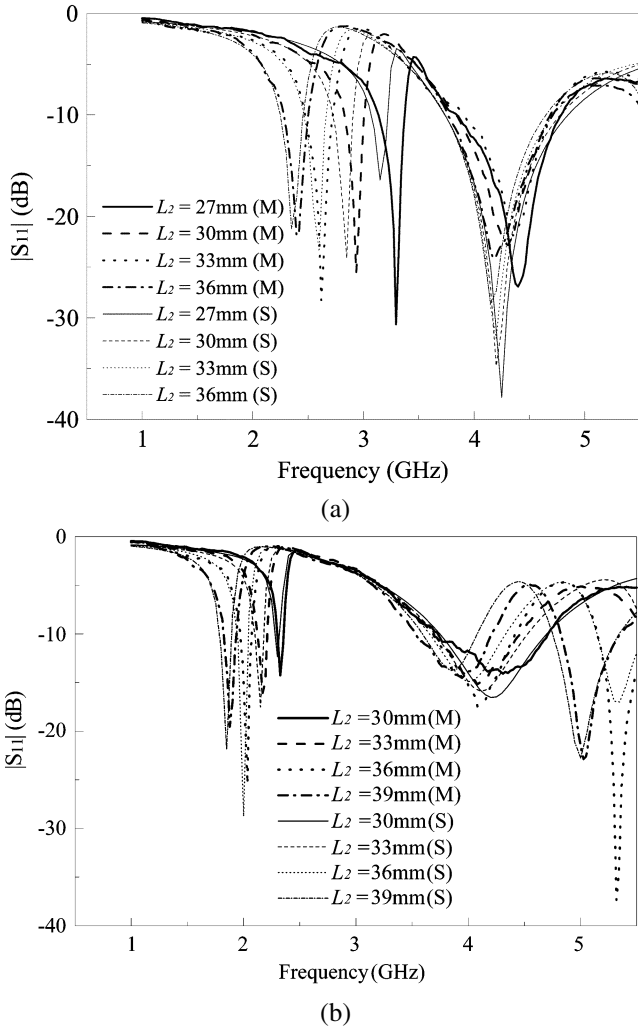


Fig. 4. Measured and simulated input return losses of the radial slot antenna with (a) $\alpha = 30^\circ$ and (b) $\alpha = 60^\circ$ for different L_2 . (M: measured, S: simulated).

as shown in Fig. 2 ($L_1 < L_2$) is satisfactory for most wireless applications.

III. ANTENNA DESIGN AND EXPERIMENTAL RESULTS

Several typical designs of the proposed antenna with various lengths of the central slot pair L_2 and included angles α were constructed and investigated. All of the test pieces were fabricated on the dielectric substrate FR-4 with dielectric constant $\epsilon_r = 4.3$, substrate thickness $h = 1.6$ mm, and loss tangent $\tan \delta = 0.02$. The widths of the strip and slot, W and G , of the 50 Ω CPW feedline are fixed to be 3.4 and 0.3 mm, respectively, and the length of the oblique slots L_1 is fixed to be 23.1 mm, which is approximately half of the guided-wavelength in the structure at 4.5 GHz. The width of the oblique slots and the central slot pair w is one of the key parameters to tune for the input match of the antenna because it has a negligible effect on the resonant frequencies and the frequency ratio. In the test pieces, w is chosen to be 2.4 or 2.6 mm for a compromised matching at resonant frequencies. A series of designs with various L_2 (27, 30, 33, 36, and 39 mm) and α (30° and 60°) were implemented and tested.

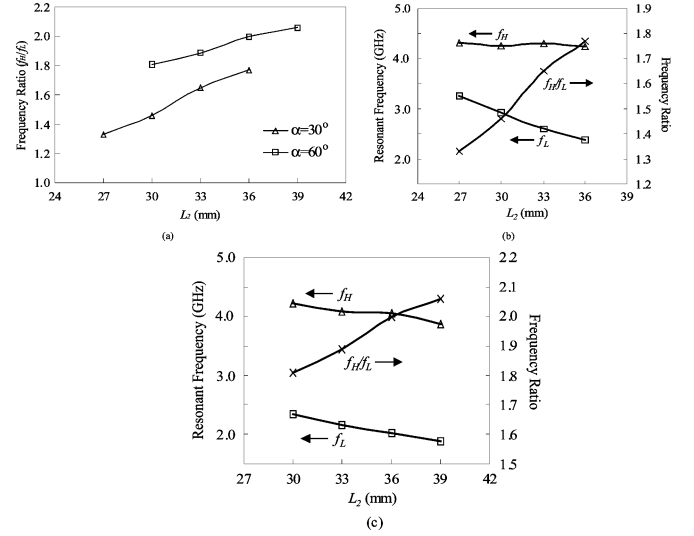


Fig. 5. Measured frequency ratio (f_H/f_L) and resonant frequencies (f_H and f_L) against L_2 . (a) f_H/f_L versus L_2 , (b) f_H , f_L , and f_H/f_L versus L_2 for $\alpha = 30^\circ$, and (c) f_H , f_L , and f_H/f_L versus L_2 for $\alpha = 60^\circ$.

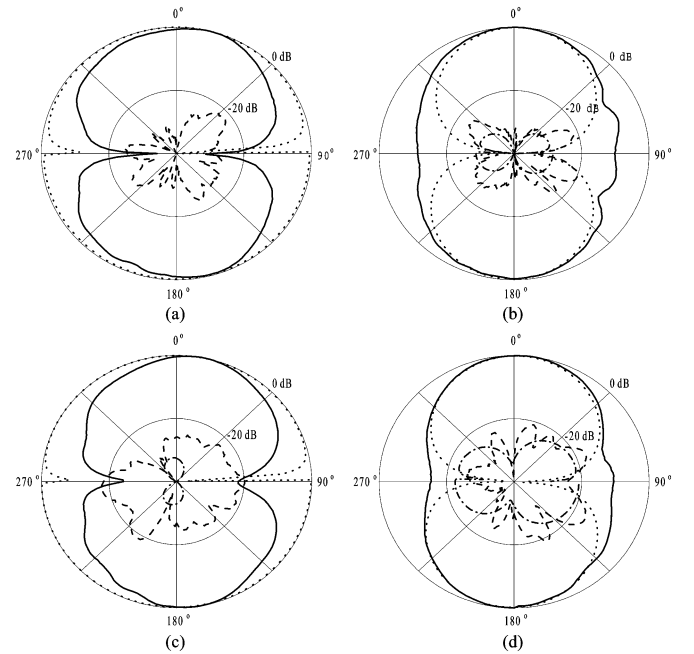


Fig. 6. Measured and simulated radiation patterns with $L_2 = 27.0$ mm and $\alpha = 30^\circ$. (a) E-plane pattern at f_H , (b) H-plane pattern at f_H , (c) E-plane pattern at f_L , and (d) H-plane pattern at f_L . — Measured co-pol, - - - Measured cross-pol, ····· Simulated co-pol, - · - · Simulated cross-pol.

Measured return losses of the test pieces and the ones simulated by Ansoft Ensemble 6.0 are shown in Fig. 4. They are in good agreement. The summary of the experimental frequency response results of Fig. 4 is depicted in Fig. 5. The measured impedance bandwidth, defined as a return loss level of 10 dB, of the upper band falls in the range from 17.4% to 23.2%, while that of the lower band ranges from 2.4% to 9.7%. The bandwidth of the lower band decreases as the included angle α increases. It is worthy mentioning that, in the course of simulation, we found that the effective length that the equivalent magnetic current actually flows at f_L along the slots elongates for

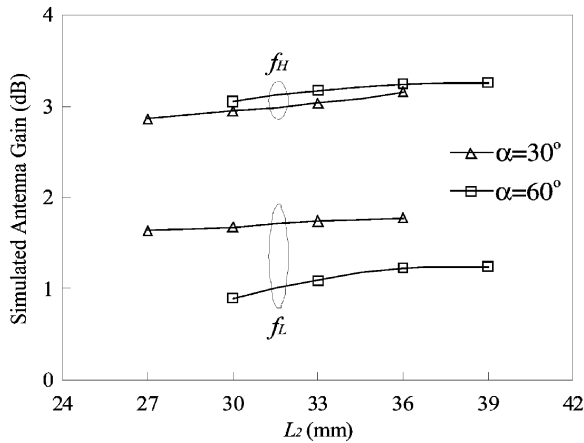


Fig. 7. Antenna gains simulated at f_H and f_L against L_2 for various α .

larger included angle α even though the physical slot lengths remain unchanged. Thus, the lower resonant frequency f_L decreases and the frequency ratio increases monotonically as the included angle α increases. Consequently, there is a tradeoff between the frequency ratio and the bandwidth of the lower band, which sets an upper limit of the frequency ratio for the design. The obtained frequency ratio for the present design varies in the range of about 1.3 to 2.1 and increases monotonically with the increasing L_2 . On the other hand, the resonant frequencies f_H and f_L decrease as L_2 increases, and f_L is more sensitive to the variation of L_2 than f_H .

The difference between radiation patterns of the antennas with various L_2 and α is insignificant. Only one case of the radiation pattern will be shown. The measured and simulated radiation patterns for the radial slot antenna with $L_2 = 27.0$ mm, $\alpha = 30^\circ$, and operating at f_H and f_L are shown in Fig. 6. Note that the cross polarization level at f_H is lower than that at f_L , which agrees well with the descriptions in Section II. The radiation patterns at f_H and f_L are of the same polarization. The antenna gains measured at f_H and f_L are 4.4 and 3.1 dBi, respectively, while those simulated at f_H and f_L are 2.86 and 1.64 dB, respectively. The antenna gains simulated at both resonant frequencies against L_2 for various α are shown in Fig. 7. The antenna gain at f_L is lower than that at f_H is due to the field cancellation between the central slot pair and the oblique slots.

IV. CONCLUSION

A novel broad-band, dual-frequency radial slot antenna fed by a CPW has been presented in this paper. This antenna simultaneously possesses a wider bandwidth of the upper band, satisfactory radiation patterns, high radiation efficiency, and compact size. In addition, the uniplanar and simple structure makes it ease of fabrication and design. Distributions of the equivalent magnetic current on the radial slot antenna have been shown

and demonstrated the principle of the dual-frequency operation. The behavior of the antenna has been studied intensively and investigated experimentally. The frequency ratio of the two resonant frequencies is tunable from 1.3 to 2.1 by varying α and L_2 . The bandwidth of the lower band ranges from 2.4 to 9.7%, and a much wider bandwidth of the upper band ranges from 17.4 to 23.2%. This antenna may find applications in many wireless communication systems.

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