S-bend waveguides. The performances may be further improved by using a more sophisticated coupling design of the S-bend waveguides.

3. CONCLUSION
A novel TMI wavelength filter with a 1D photonic crystal holey structure has been proposed and verified using the 2D FDTD simulation method. Two wavelength channels on different sides of the PBG at 1.55 μm were separated by a coupler length of 76.8 μm. The channel widths were about 40 nm and the channel spacing was about 35 nm. Channel contrasts of 20 dB and insertion losses of 0.8 dB were achieved. It is expected that the channel spacing and channel widths in a 3D device will be much narrower than those calculated in a 2D simulation. A numerical method more efficient than the FDTD method will be helpful and necessary for further 3D performance studies and device manufacturing.

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LOW-PROFILE OMNIDIRECTIONAL CIRCULARLY POLARIZED ANTENNA FOR WLAN ACCESS POINTS

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ABSTRACT: A novel low-profile omnidirectional circularly polarized antenna comprising a top-loaded cylindrical monopole and four printed arc-shaped dipoles is presented. The cylindrical monopole and four printed dipoles share a common feeding point and generate two orthogonal polarizations with a 90° phase difference, resulting in omnidirectional circular-polarization (CP) radiation in the azimuthal plane. The proposed antenna is also arranged in a compact structure with a low profile (less than 0.1 wavelength of the center operating frequency). A constructed prototype suitable for applications in a wireless local area network (WLAN) access point in the 2.4-GHz band is demonstrated. © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 46: 227–231, 2005; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.20952

Key words: omnidirectional antennas; circularly polarized antennas; low-profile antennas; WLAN antennas

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1. INTRODUCTION

Omnidirectional antennas are very attractive for applications in wireless communications, such as WLAN systems, that need to cover a large service area. For such applications, a variety of promising omnidirectional antennas have been reported recently. To name a few, related omnidirectional antenna designs include the use of cylindrical patch antennas [1], planar dipole antennas [2–6], planar collinear antennas [7–9], and so on. These designs, however, generate only linearly polarized (LP) radiation. For achieving omnidirectional CP antennas should find great applications for wireless communications, such as WLAN systems, that need to cover a large service area. For such promising applications, we propose a novel omnidirectional CP antenna with a low profile of less than 10% of the wavelength of the center operating frequency in this paper. A prototype of the proposed antenna suitable for applications in a WLAN access point in the 2.4-GHz band (2400–2484 MHz) is constructed and experimentally studied.

2. ANTENNA DESIGN

Figure 1(a) shows the geometry of the proposed omnidirectional CP antenna. Top and side views of the antenna are shown in Figures 1(b) and (c). The antenna consists of two main parts: (i) a top-loaded cylindrical monopole and (ii) four printed arc-shaped dipoles. The top-loaded cylindrical monopole comprises a hollow metal cylinder (diameter $S$, height $h$) and a loading circular patch of diameter $T$ centered on top of the cylinder. The bottom end of the cylinder is also centered on a circular grounded substrate, on which four identical arc-shaped dipoles (length $d$ and width 2 mm) are printed along the perimeter of the circular substrate. Note that a circular patch of diameter $D$ is printed on the bottom side of the substrate, which serves as the ground plane of the cylindrical monopole and also as the ground plane for four 50Ω microstrip lines feeding the four printed dipoles. Also note that each dipole is spaced away from the circular ground plane at a distance $t$, and one dipole arm is printed on the top side of the substrate while the other arm is on the bottom side of the substrate.

Both the top-loaded cylindrical monopole and the four printed dipoles share a common feeding point at the center of the circular substrate and are fed using a coaxial probe feed through a via-hole in the substrate. Good impedance matching for the antenna can be achieved by tuning the diameter $S$ of the cylindrical monopole. The cylindrical monopole in this design operates as a quarter-wave structure and generates an omnidirectional LP wave with vertical polarization ($E_z$). On the other hand, the four printed dipoles operate as half-wave structures and together generate an omnidirectional LP wave with horizontal polarization ($E_x$).

By further adjusting the spacing $t$ (the spacing between the dipoles and the circular ground plane), the amplitude of the horizontal LP radiation can be tuned to be the same as that of the vertical LP radiation. (Related effects are discussed in section 3). As for tuning the phase difference between the horizontal and vertical LP radiation to be 90°, it is most effective to adjust the diameter of the circular ground plane, which changes the length of the four microstrip feedlines and in turn changes the phase difference of the signals fed into the printed dipoles referenced to that into the cylindrical monopole. (The effects of the ground-plane size on the phase difference of the two orthogonal polarizations are also be explored in Table 1.) With equal amplitude and 90° phase difference, the horizontal and vertical omnidirectional LP waves can then be formed into an omnidirectional CP wave.

Also note that the dipole arms of the four printed dipoles on the top surface of the substrate are arranged along the $−\phi$ direction and the cylindrical monopole is in the $+z$ direction. In this case, the resulting electric field of the CP wave in the azimuthal plane will rotate in a counterclockwise direction from the $+z$ direction to the $−\phi$ direction, thereby leading to a right-handed CP wave. Conversely, when the dipole arms of the four printed dipoles on

![Diagram](image)

**Figure 1** (a) Geometry of the proposed low-profile omnidirectional CP antenna: (b) top view; (c) side view

| TABLE 1 Comparison of the Phase Difference (Calculated Data*) Due to the Microstrip Feedline for the Printed Dipoles and the Measured Phase Difference of the $E_x$ and $E_y$ Components in the Azimuthal Plane (Other Parameters are as in Fig. 2) |
|---------------------------------|-------|-------|-------|
| Ground-Plane Diameter $D$      | 44 mm | 50 mm | 56 mm |
| Phase difference due to the microstrip feedline for dipoles | 100°  | 116°  | 132°  |
| Measured phase difference of $E_x$ and $E_y$ ($\angle E_x − \angle E_y$) | 84°   | 93°   | 112°  |

* Phase lag $= 360° \times$ (microstrip feedline length/guided wavelength in the FR4 substrate)
the top surface of the substrate are arranged to be along the $+\phi$ direction, the resulting electric field of the CP wave in the azimuthal plane will rotate in a clockwise direction from the $+z$ direction to the $+\phi$ direction, thereby leading to a left-handed CP wave. The two cases described above are fabricated and tested, and both the measured and simulated results confirm the predicted rotation of the CP wave. For brevity, only the right-hand CP case is demonstrated in section 3.

The design procedure is also summarized as follows. The first step is to select the lengths $h$ and $T$ of the top-loaded cylindrical monopole ($h + 0.5T$ is close to 0.25 wavelength of the desired center frequency, with $h$ chosen to be less than 0.1 wavelength of the center frequency in order to achieve a low profile). The second step is to choose the length of the four printed dipoles ($d$ is about 0.5 wavelength of the center frequency, when the effect of the substrate is not considered). The third step is to select an optimal diameter $S$ of the cylindrical monopole for achieving good impedance matching over a desired band. The final step is to adjust the spacing $t$ and ground-plane diameter $D$ in order to achieve equal amplitude and 90° phase difference for the two orthogonal polarizations. The final step can be aided using a simulation software such as the Ansoft High-Frequency structure simulator (HFSS) and the optimal or near-optimal design dimensions of the antenna can then be obtained.

It should also be noted that the cylindrical monopole and the four printed dipoles, respectively, generate $E_\theta$ and $E_\phi$ radiation with similar monopole-like radiation patterns (see Figs. 5 and 7), which are then formed into omnidirectional CP radiation by performing the final step described above. In addition, as described in the third step above, simply by selecting a proper diameter of the cylindrical monopole, without varying the parameters of the four printed dipoles, the impedance matching for the antenna can be easily achieved (note that varying the diameter has a very small or negligible effect on the radiation pattern of the cylindrical monopole). In sum, the cylindrical monopole in this design has two functions: (i) to generate omnidirectional $E_\theta$ radiation and (ii) to achieve good matching for the antenna.

3. EXPERIMENTAL RESULTS AND DISCUSSION

The proposed antenna for WLAN operation in the 2.4-GHz band was constructed and tested. The length $h$ was chosen to 10 mm, thus the total antenna height (10.8 mm) was only about 8.8% of the wavelength at 2442 MHz. The length $h + 0.5T$ (=28 mm) is close to a quarter-wavelength at 2442 MHz. The length $d$ of the four printed dipoles, however, is 44.8 mm only (about 36.5% of the free-space wavelength at 2442 MHz, much less than a half-wavelength). This behavior is mainly due to the presence of the FR4 substrate, which decreases the resonant length of the printed dipoles. So that the optimal diameter $S$ of the cylindrical monopole would achieve good impedance matching, it was found to be 7 mm in this design. In order to achieve equal amplitude and 90° phase difference for the two orthogonal polarizations, the spacing $t$ and the ground-plane diameter $D$ were determined to be 4 and 50 mm, respectively.

The measured and simulated return loss is presented in Figure 2. The simulated results were obtained using Ansoft HFSS. The measured data agree with the simulated results. The measured 10-dB return-loss impedance bandwidth was 110 MHz (2395–2505 MHz). Figure 3 shows the measured axial ratio at $\theta = 90^\circ$ (the azimuthal plane), 75°, and 60°. With regard to the CP radiation in the azimuthal plane, the CP bandwidth, defined by a 3-dB axial ratio, reaches 145 MHz (2370–2515 MHz), which covers the 2.4-GHz WLAN band. It is also seen that at $\theta = 75^\circ$, the axial ratio of the CP radiation over the 2.4-GHz band is still less than 3 dB. Figure 4 plots the measured spinning linear radiation patterns in the $x$–$y$ plane (azimuthal plane) and $x$–$z$ plane (elevation plane) at 2442 MHz. Good omnidirectional CP radiation in the azimuthal plane with ripples of less than 1 dB is obtained (the ripple corresponds to the measured axial ratio of the CP radiation). Figure 5 presents the simulated vertical polarization ($E_\theta$) and horizontal polarization ($E_\phi$) radiation patterns. It is seen that the $E_\theta$ and $E_\phi$ radiation are almost of the same amplitude, except in the directions

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The page contains figures and tables that are not included in the text. The figures are labeled as follows:

- Figure 2: Measured and simulated return loss with $h = 10$ mm, $T = 36$ mm, $d = 44.8$ mm, $S = 7$ mm, $t = 4$ mm, and $D = 50$ mm
- Figure 3: Measured axial ratio vs. frequency
- Figure 4: Measured spinning linear radiation patterns at 2442 MHz
- Figure 5: Simulated vertical polarization ($E_\theta$) and horizontal polarization ($E_\phi$) radiation patterns.
near $\theta = 0^\circ$ and $180^\circ$ in the $x$–$z$ plane pattern, which agrees with the measured spinning linear radiation patterns shown in Figure 4. The measured antenna gain is shown in Figure 6, and a stable antenna gain of about 0.4–0.5 dBi is obtained.

To analyze the effect of the spacing $t$ between the printed dipoles and the circular ground plane on the amplitude of the horizontal polarization ($E_y$), Figure 7 shows the simulated radiation patterns for the cases with $t = 1$, $3$, and $5$ mm. For $t = 1$ mm, the $E_y$ radiation is seen to have a smaller amplitude than the $E_z$ radiation. This behavior is largely due to the large capacitive coupling contributed from the small spacing between the printed dipoles and the circular ground plane. With an increase in $t$, the amplitude of the $E_y$ radiation increases and can be about the same as that of the $E_z$ radiation (see $t = 5$ mm in Fig. 7 and the antenna studied in Fig. 5 with $t = 4$ mm), thus making it possible to achieve good CP radiation. In this study, the case of $t = 4$ mm, which leads to a smaller substrate, is selected for the constructed prototype.

Another important factor for achieving good CP radiation is that the phase difference of the $E_y$ and $E_z$ components should be about $90^\circ$, which can be achieved by adjusting the diameter $D$ of the circular ground plane. When $D$ is varied, the length $[(D - S)/2]$ of the four microstrip feedlines is varied; thus, the phase difference of the signals fed into the printed dipoles referenced to that into the top-loaded cylindrical monopole is changed. The calculated results of the phase difference due to the microstrip feedlines are listed in Table 1 for comparison. It is seen that with an increase in $D$, the phase difference increases. Thus, through selecting a proper value of $D$, the desired phase difference can be obtained. However, it is noted that the calculated phase difference is not equal to the measured phase difference of the $E_y$ and $E_z$

Finally, a simulation study on further decreasing the antenna height $h$ was also conducted. The results indicate that when $h$ is decreased from 10 to 7 mm, the cylindrical monopole requires a larger diameter ($S = 9$ mm for $h = 7$ mm versus $S = 7$ mm for $h = 10$ mm) for achieving good impedance matching over the 2.4-GHz band. In this case, in order to achieve a $90^\circ$ phase difference between the $E_y$ and $E_z$ components, an FR4 substrate with a larger diameter (>62 mm) is required. Moreover, with a decreased height of the cylindrical monopole, a larger size of the top-loading circular patch is required for resonating at 2.4 GHz. In this case, the coupling between the circular patch and the ground plane will increase, which decreases the achievable CP bandwidth. For $h = 7$ mm, the 3-dB axial-ratio CP bandwidth is found to be about 98 MHz, about 67% that of the case with $h = 10$ mm studied here.

4. CONCLUSION

A novel low-profile omnidirectional CP antenna has been proposed. The antenna can be implemented at low cost, and a proto-
type suitable for WLAN access-point application at 2.4 GHz has been constructed and tested. The constructed prototype has a low profile of about 8.8% of the wavelength at 2442 MHz and shows very good omnidirectional CP radiation in the azimuthal plane (with an axial-ratio variation of less than 1 dB). In addition, the prototype shows a stable antenna-gain level of about 0.5 dBic for frequencies across the 2.4-GHz band.

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SUPERPOSITION OF MODE COUPLING INDUCED BY DUAL ACOUSTO-OPTIC MODULATIONS
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ABSTRACT: The high-order-mode coupling of acoustic-optic modulations in a tilted fiber grating is experimentally demonstrated to be greatly enhanced due to the superposition of dual acoustic-optic modulations. The performance is based on the phase-matching condition for the acoustic-wave vector equivalent to the difference between the core-mode and cladding-mode wave-vectors. The wave vector is considered to be a tunable long-period fiber grating. Sun et al. used phase-matching theory to identify the coupling relationship between the cladding radius and flexural acoustic-wave periods with a chosen optical wavelength [5]. However, based on the previous experimental results [3], the coupling phenomenon of the 3rd-order diffraction by acousto-optic interactions in the fiber grating is not obvious, due to the insufficient acoustic power. This work proposes a new configuration to improve the coupling efficiency of the high-order diffractions of a fiber grating via dual acoustic-optic modulations in different acoustic frequencies. The superposition phenomenon of acoustic-optic modulations via two different-frequency acoustic waves in a tilted fiber grating is experimentally demonstrated by high-efficiency high-order-mode coupling. This method can induce several optical channels for applications in switchable multichannel fiber lasers and wavelength-division multiplexing (WDM) filters.

2. PRINCIPLE
A tilted fiber Bragg grating is fabricated by means of rotating the phase mask and UV-laser exposure to cause the Bragg grating plane to be blazed with respect to the fiber axis. The guided light in the fiber core thus can be partly coupled into the guided-cladding or radiation modes, in which the coupling intensity is proportional to the slanted angle. Because the axis-direction component of the wave-vector of a slanted grating can cause coupling between the forward- and backward-propagating core modes at the Bragg wavelength \( \lambda_B \), for the wavelength \( \lambda_{i,1,2,\ldots} \), which is shorter than Bragg wavelength \( \lambda_B \), the grating-wave vector can cause the coupling from the forward-core mode \( \beta_{0i} \) to the backward-cladding mode \( -\beta_{1i} \), as illustrated in Figure 1. The phase-matching condition can be satisfied by

\[ \beta_{0i}(\lambda_B) - [-\beta_{1i}(\lambda_B)] = K \cos \theta \]

\[ \beta_{0i}(\lambda) - [-\beta_{1i}(\lambda)] = K \cos \theta \quad \lambda < \lambda_B, \quad (1) \]

where \( \beta_{0i} \) and \( \beta_{1i} \) denote the propagation constant of the core and cladding modes, respectively, and \( \theta \) is the slanted angle of a fiber grating. Moreover, \( K \) represents the wave vector of a fiber grating. When the acoustical flexural waves are launched into the fiber

Figure 1 Wave-vector diagram for the phase matching of acoustic-optic modulations in a tilted Bragg fiber grating (the induced wavelength \( \lambda_i \) is determined by the acoustic frequency)