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Electrically small dual-band substrate-integrated-waveguide antenna with fixed low-frequency and tunable high-frequency bands

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Abstract: In this paper, we propose a novel electrically small dual-band reconfigurable antenna. Low- and high-frequency resonances are generated by a substrate-integrated-waveguide (SIW) cavity resonator and a complementary split-ring resonator (CSRR). The overall size is reduced because of the use of an eighth-mode SIW (EMSIW) and an evanescent-mode CSRR. The varactor-loaded CSRR enables only the high frequencies to continuously vary from 2.38 GHz to 2.5 GHz for channel selection of wireless local area network (WLAN), while the low frequencies are fixed at 1.575 GHz for global position system (GPS), because the two resonant frequencies can be independently controlled. The proposed concept is demonstrated by the simulated and measured s-parameters and radiation patterns.

Keywords: substrate-integrated-waveguide (SIW), eighth-mode substrate-integrated-waveguide (EMSIW), complementary split-ring resonator (CSRR), dual-band antenna, reconfigurable antenna

Classification: Electromagnetic theory

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1 Introduction

Given that wireless communication is widely used at present, a wide range of frequency spectra is available for use. Therefore, dual-band or frequency-reconfigurable components that satisfy various frequency standards have been developed [1, 2, 3, 4, 5, 6, 7]. Unlike a dual-band antenna that always satisfies the condition of two different bandwidths, a frequency-reconfigurable antenna can vary its frequency, and this antenna exhibits reduced interference between the different frequency bandwidths, even though simultaneous multifunctionality is not allowed [5, 6, 7].

To realize an electrically small antenna, an eighth-mode substrate-integrated-waveguide (EMSIW) cavity is introduced. The size of the EMSIW is only one-eighth the size of a full-mode SIW with the same resonant frequency and electric/magnetic field distributions. The EMSIW cavity antenna is employed for global positioning system (GPS) at 1.575 GHz. A circular polarization (CP) signal is used for GPS service. However, a linear polarization (LP) antenna can be a substitute for the GPS antenna on a space-limited device [8, 9], and existing studies have revealed that both CP and LP exhibit similar performance in an urban environment with heavy reflections [10].

An electrically small complementary split-ring resonator (CSRR) is used to cover the higher frequencies of 2.4-GHz wireless local area network (WLAN) band. WLAN is divided into 14 channels at the 22-MHz bandwidth as per IEEE 802.11 standard because the dividing channels can reduce the interference caused among many users of the same frequency bandwidth. These standards regulate not only the frequencies but also the attenuation rate of the spectrum power in the neighboring channels. Thus, channel selection filters are required in order to satisfy such standards, and in turn, more space is required in a mobile device. The CSRR with a varactor diode has narrow bandwidth as well as continuous variation of the resonant frequency from 2.4–2.5 GHz. Therefore, it is suitable to satisfy the channel selection function of the WLAN standard. The two slot rings of the CSRR possess a high Q factor, and many of the existing CSRRs are combined with the SIW [11]. Further, these rings are coupled, thereby allowing negative permittivity and permeability [12].

In this study, we loaded a CSRR structure on the EMSIW and developed a dual-band antenna. The EMSIW cavity resonates at 1.575 GHz

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in a small footprint. The CSRR structure also resonates at the 2.4-GHz band, and its overall area is small enough to allow it to be loaded on the EMSIW. In addition, a CSRR with a varactor diode facilitates variation in the resonant frequency at the 2.4-GHz band.

2 Design

The electric and magnetic field components of the conventional full-mode SIW structure are given by [13, 14]

$$E_z = E_0 \sin\frac{\pi x}{W} \sin\frac{\pi y}{L} \tag{1}$$

$$H_y = \frac{j\pi E_0}{k\eta W} \cos\frac{\pi x}{W} \sin\frac{\pi y}{L} \tag{2}$$

$$H_x = \frac{-j\beta E_0}{k\eta} \sin\frac{\pi x}{W} \sin\frac{\pi y}{L}, \tag{3}$$

where L and W indicate the height and width, respectively, of the rectangular waveguide; η, k, and β are the intrinsic impedance, wavenumber, and the phase constant, respectively.

Fig. 1 shows the magnitudes of the electric fields for the full-mode square SIW, the half-mode SIW (HMSIW), the quarter-mode SIW (QMSIW), and the EMSIW structure. The corresponding field distributions are plotted in the TE_{10} mode, which is a dominant mode. The E-field has only z-components with its maximum magnitude at the point O in Fig. 1 (a). As shown in Fig. 2, the full-mode SIW has the H-field revolving

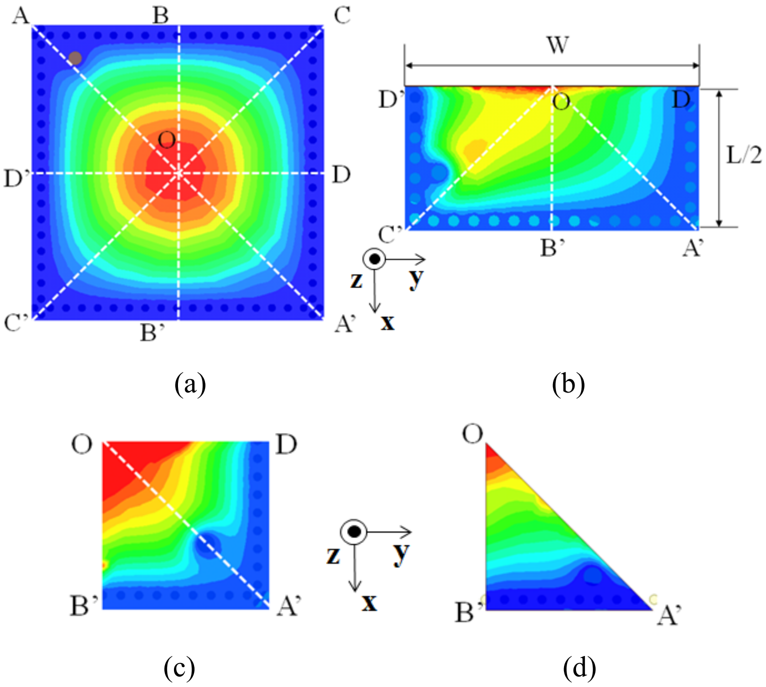

Fig. 1. Magnitude of the electric field distribution: (a) Full-mode substrate-integrated waveguide (SIW), (b) Half-mode SIW (HMSIW), (c) Quarter-mode SIW (QMSIW), and (d) Eighth-mode SIW (EMSIW)

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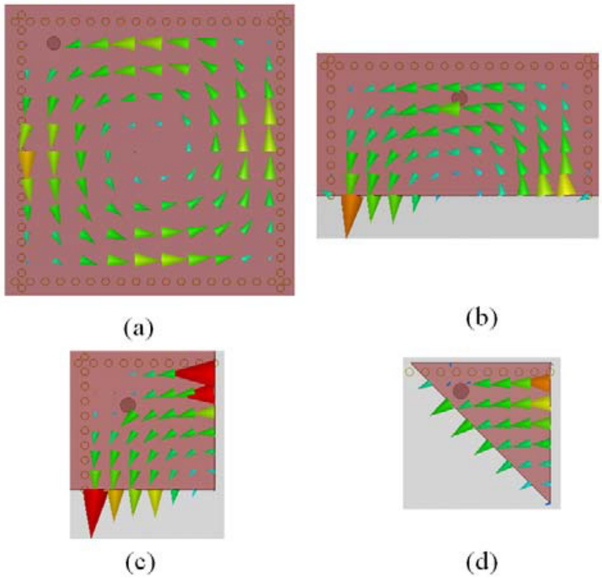

Fig. 2. Magnetic field vector distribution: (a) SIW, (b) HMSIW, (c) QMSIW, and (d) EMSIW

around the center point O in Fig. 1 (a) where the H-field has only x- and y-components. The fictitious magnetic wall is generated on the AA′, BB′, CC′, and DD′ planes because the H-field is vertically located on all the planes. The HMSIW is realized by bisecting the full-mode SIW structure along the D-D′ plane owing to a fictitious magnetic wall as shown in Fig. 1 (b). Similarly, the QMSIW can be realized by cutting the HMSIW in half. Finally, an EMSIW is realized by bisecting the QMSIW [11]. The EMSIW has a reduced one-eighth of a full-mode SIW size while the electric field distribution and the resonant frequency are maintained constant. The EMSIW provides an additional advantage for antenna applications. Because it has two open sides (O-A′ and O-B′ planes), the radiating slot is not required.

Fig. 3 shows the layout of the proposed antenna with its geometrical dimensions. The triangular shape of the EMSIW resonates at 1.575 GHz. It comprises a series of silver vias on only one side and is built on a Rogers/Duroid 5880 substrate with a thickness of 1.575 mm and a permittivity of 2.2. Using the additional stub and gap at the end of the feeding line, the

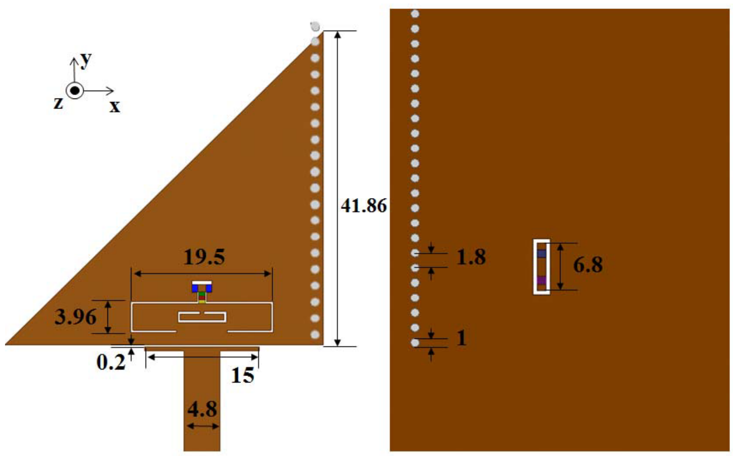

Fig. 3. Overall configuration of the proposed antenna. (a) Top view and (b) bottom view (unit: mm)

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impedance is matched to 50 Ω.

For generating an additional resonance at the 2.4-GHz band, the CSRR is etched on the top surface of the EMSIW structure, which does not require additional space. The input impedance at 2.4 GHz is dependent on the position of the CSRR, whereas the input impedance at 1.575 GHz is dependent on the stub and the gap at the feeding line. The varactor-loaded CSRR generates an additional resonance and continuously varies the resonant frequency from 2.38 GHz to 2.5 GHz by varying the bias voltage. A bias network is carefully designed so that the bias voltage of the varactor diode can be applied without disturbing the radiation performance.

Fig. 4 shows the bias network design. The slot around a copper pad enables the DC signal to be isolated from the RF signal. However, because the slot considerably affects the CSRR, a 51-pF chip capacitor is employed as a connection for the RF signal from the antenna to the copper pad and a DC block at each end of the slot on the top plane. A 100-nH chip inductor and a 10-kΩ chip resistor are employed for the RFC and the CLR, respectively, and are connected in series. The RFC and CLR are placed on the bottom, isolated from the ground plane by a narrow slot, and connected to the copper pad on the top plane through a silver via to minimize the impact on the radiation performance.

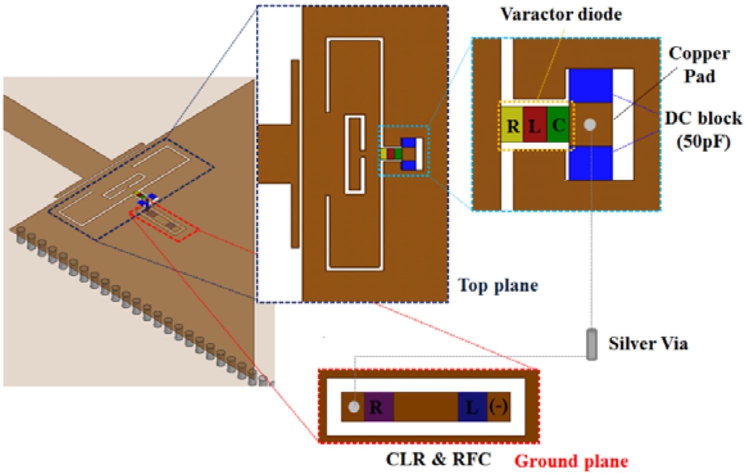

Fig. 4. Layout of the proposed antenna, varactor diode, and bias network

3 Simulation and measurement results

Fig. 5 shows the fabricated prototype. The positive DC signal is applied from the connector to the ground plane through the bias tee, and the negative DC signal is applied to the isolated part on the ground plane. Fig. 6 shows the simulated and measured return losses of the proposed antenna at different varactor diode capacitances of 0.31 pF, 0.36 pF, 0.41 pF, and 0.46 pF, which correspond to bias voltages of 30 V, 24.5 V, 17.2 V, and 11.4 V, respectively, through EM simulation. It is observed from Fig. 6 (a) that the resonant frequency at 1.575 GHz remains constant, although the bias voltage is varied. The frequency is shifted by about 2 MHz in the measurement result compared with the simulation result. This is due to fabrication errors, such as a via array. A via is built by

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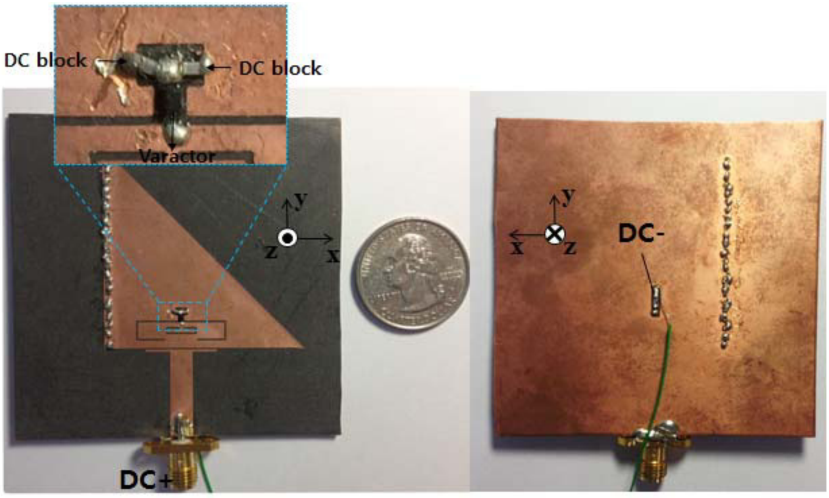
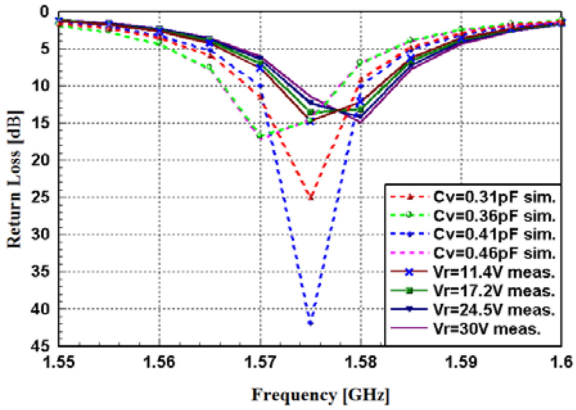
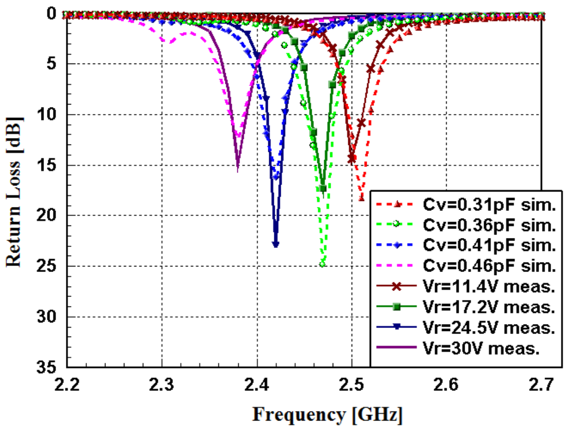


(a) (b)

Fig. 5. Picture of the proposed antenna prototype; (a) top view and (b) bottom view



(a)



(b)

Fig. 6. Measured and simulated gain patterns of the proposed antenna (a) at $\phi = -30^{\circ}$ (b) at $\phi = 0^{\circ}$ and (c) at $\phi = 90^{\circ}$

drilling a hole and then filling the silver.

Fig. 6 (b) shows the simulated and measured return losses at a higher-frequency band at different bias voltages. When the voltage is varied from

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11.4 V to 30 V, that is, the capacitance value of the varactor diode is varied from 0.46 pF to 0.31 pF, the resonant frequency varies from 2.38 GHz to 2.5 GHz while the10-dB bandwidth is maintained at approximately 17 MHz.

The radiation patterns are measured in an anechoic chamber. Fig. 7 shows the simulated and measured gain patterns on the z-x plane. Fig. 7 (a) shows the gain pattern of the EMSIW at 1.575 GHz. The measured peak gain is 0.11 dBi at boresight. Fig. 7 (b) shows the gain patterns at 2.42 GHz with a 24.5 V bias voltage. The peak gain is −0.6 dBi along the z-axis. Fig. 7 (c) shows the gain patterns at 2.47 GHz with a 17.2 V bias voltage. The peak gain is −1.2 dBi along the z-axis. When an additional gain is required depending on the applications, the gain can be easily compensated by a low-noise amplifier, which is implemented in a

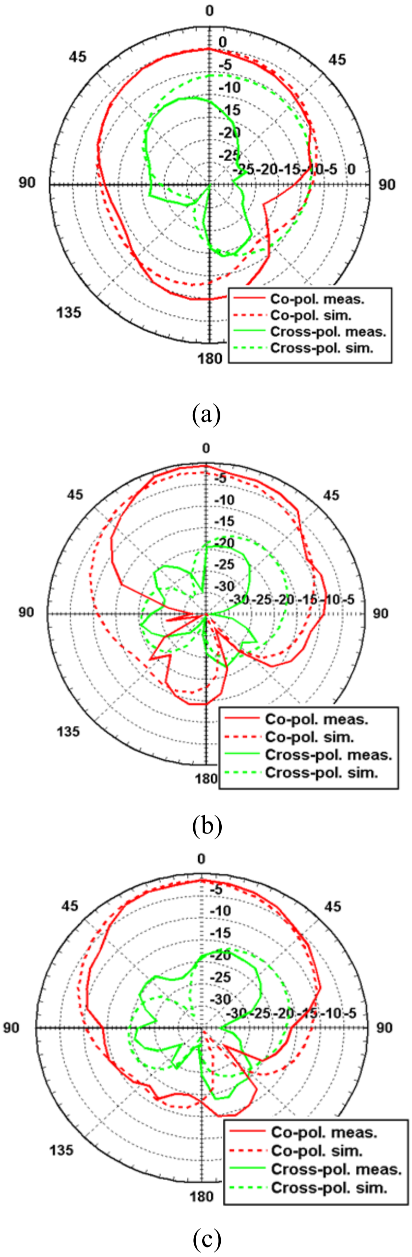

Fig. 7. Measured gain patterns of the proposed antenna. (a) 1.575 GHz, (b) 2.42 GHz at a bias voltage of 17.2 V, (c) 2.47 GHz at a bias voltage of 24.5 V

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radio-frequency integrated circuit (RFIC) chipset.

4 Conclusion

In this paper, we proposed a dual-band reconfigurable antenna by using an EMSIW resonator and a tunable CSRR. The proposed antenna satisfactorily operates at the 1.575-GHz and 2.38–2.5-GHz bandwidths. The antenna resonates at a lower frequency, when an electronically small EMSIW structure is used, and the antenna additionally resonates at a higher-frequency band, when a CSRR structure is built in the EMSIW without additional space. The reverse voltage of the varactor diode is varied from 11.4 V to 30 V, and the resonant frequency is varied from 2.38 GHz to 2.5 GHz, while the lower resonant frequency is maintained constant at 1.575 GHz. The peak gain is 0.11 dBi at 1.575 GHz and −0.6 dBi at 2.47 GHz. Therefore, the proposed antenna is suitable for mobile devices requiring GPS and WLAN services with channel selection capability.

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