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A Compact Metamaterial Quad-band Antenna Based on Asymmetric E-CRLH Unit Cell

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Abstract—In this paper, a compact metamaterial quad-band antenna is presented. The antenna is designed from a unit cell of asymmetric extended-composite right/left handed transmission line (E-CRLH TL) as the main resonator part and a 50 Ω coplanar waveguide (CPW) as the feeding part. The design concept and resonant frequencies are analyzed and discussed. The results show that the proposed antenna exhibits four frequency bands covering GSM810, WLAN 2.45/5.5 GHz and WiMAX 3.5 GHz bands. The overall size of the fabricated antenna is only 57.2 mm × 31.2 mm × 1.6 mm and is very small compared with other proposed quad-band antennas. In addition, a good agreement can be seen among the estimated resonant frequencies, HFSS simulated and measured results.

1. INTRODUCTION

Nowadays, the development of modern wireless communication systems requires compact devices to work at different standards. It leads to the necessity of designing small antennas with multiband operation [1]. Many studies on dual- and tri-band antennas have been reported while quad-band antennas are seldom proposed. Compared with dual- and tri-band antennas, quad-band antennas are more desirable for reducing the fabrication cost. Some types of quad-band antennas have been designed by using conventional methods such as a monopole antenna [2], slot antennas [3–6], a fractal antenna [7], a pentangle-loop antenna [8], and a Hilbert shaped antenna [9]. However, they still have large size corresponding to the wavelength at their operating frequencies. Other more compact techniques have also been presented in the literature based on parasitic elements [10], loops [11], and matching networks [12].

Metamaterials provide a conceptual way to reduce the size of antennas for satisfying the requirements of modern wireless communication systems. With metamaterial transmission line approach, composite right/left handed transmission line (CRLH TL) and dual-composite right/left handed transmission line (D-CRLH TL) were introduced in 2006 [13, 14]. These transmission lines have been employed to design many compact dual- and tri-band antennas for wireless communications [15–19]. In order to get more interesting properties, extended-composite right/left handed transmission line (E-CRLH TL) has been developed [20]. This TL is also known as a generalized negative refraction index transmission line (NRI-TL) [21].

Few antennas based on E-CRLH TL and NRI-TL were proposed up to now. A dual-band leaky wave antenna comprising 10 NRI-TL unit cells is simulated in [22]. Another dual-band leaky wave antenna is designed from 10 E-CRLH unit cells in [23]. These leaky wave antennas are used in specific radar applications for their capability of beam scanning and high directivity. Ref. [24] presents a dual-band antenna based on a modified asymmetric NRI-TL unit cell. This antenna has a compact size but exhibits a very low gain at the operating frequencies. More recently, a multiband antenna based on one

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asymmetric E-CRLH unit cell is designed for quad-band [25]. However, the resonant frequencies have not been analyzed yet.

The E-CRLH TL exhibits two RH and two LH bands. In our previous investigation [26], symmetric E-CRLH TL can be applied to quad-, dual-, and tri-band applications by selecting cutoff frequencies of the dispersion diagram in unbalanced, balanced, and mixed cases, respectively. In this paper, the design scheme is now extended to be applied to asymmetric cases. By carefully choosing L-C lumped circuit elements, one can analyze resonant frequencies of E-CRLH antennas. In order to show the realizability of this design, a quad-band antenna has been made from a unit cell of asymmetric E-CRLH TL on a co-planar printed circuit board. The proposed antenna in this paper has a compact size compared with the previously proposed quad-band antennas. In the following Section 2, basic antenna design concept is explained analytically first, and S_{11} simulated result is discussed. The antenna is fabricated, and its performance is presented in Section 3. Finally, conclusions are made in Section 4.

2. ANTENNA ANALYSIS AND DESIGN

The equivalent circuit of an asymmetric E-CRLH unit cell is shown in Fig. 1. In the horizontal branch, a series L_1-C_1 resonator connects in series with a parallel L_2-C_2 resonator. The vertical branch contains a parallel L_3-C_3 resonator in shunt with a series L_4-C_4 resonator. Impedance Z_h of the horizontal branch and admittance Y_v of the vertical branch are given respectively by

$$Z_h = \frac{jL_1\left(\omega^2-\omega_{Z01}^2\right)\left(\omega^2-\omega_{Z02}^2\right)}{\omega\left(\omega^2-\omega_{Z\infty}^2\right)}, \tag{1}$$

$$Y_v = \frac{jC_3\left(\omega^2-\omega_{Y01}^2\right)\left(\omega^2-\omega_{Y02}^2\right)}{\omega\left(\omega^2-\omega_{Y\infty}^2\right)}, \tag{2}$$

where

$$\omega_{Z\infty}^2 = \frac{1}{L_2C_2}, \quad \omega_{Y\infty}^2 = \frac{1}{L_4C_4}, \tag{3}$$

$$\omega_{Z01}^2 = \frac{B_1+\sqrt{B_1^2-4A_1}}{2}, \quad \omega_{Z02}^2 = \frac{B_1-\sqrt{B_1^2-4A_1}}{2}, \tag{4}$$

$$\omega_{Y01}^2 = \frac{B_2+\sqrt{B_2^2-4A_2}}{2}, \quad \omega_{Y02}^2 = \frac{B_2-\sqrt{B_2^2-4A_2}}{2}, \tag{5}$$

$$A_1 = \frac{1}{L_1C_1L_2C_2}, \quad B_1 = \frac{1}{L_1C_1}+\frac{1}{L_2C_2}+\frac{1}{L_1C_2}, \tag{6}$$

$$A_2 = \frac{1}{L_3C_3L_4C_4}, \quad B_2 = \frac{1}{L_3C_3}+\frac{1}{L_4C_4}+\frac{1}{L_4C_3}. \tag{7}$$

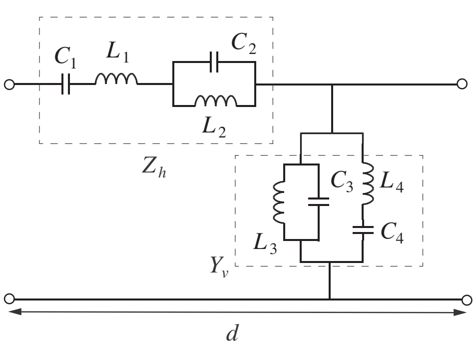


Figure 1. Equivalent circuit of an asymmetric E-CRLH unit cell.

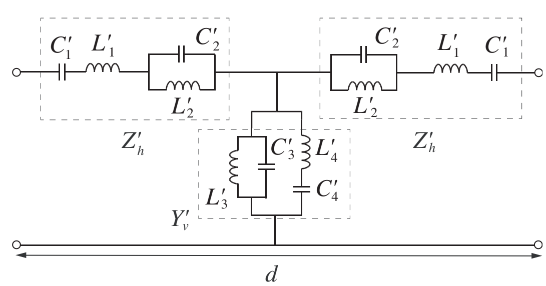


Figure 2. Equivalent circuit of a symmetric E-CRLH unit cell.

By applying the periodic boundary conditions related with Bloch-Floquet theorem [27] to the asymmetric unit cell, the dispersion relation is obtained as

$$\cos(\beta d) = 1 + \frac{Z_h Y_v}{2}, \tag{8}$$

where β is the propagation constant for the Bloch waves, and d is the length of the unit cell. To realize an antenna based on E-CRLH resonators in general N unit cells, the following resonant condition should be applied:

$$\beta d N = n\pi, \tag{9}$$

where the resonant modal index n can be positive integers for RH bands, zero and negative integers for LH bands. Therefore, one may calculate the resonant frequencies of an antenna based on N-asymmetric E-CRLH unit cell as:

$$1 + \frac{Z_h Y_v}{2} - \cos\left(\frac{n\pi}{N}\right) = 0, \quad n = 0, \pm 1, \pm 2, \ldots, \pm N, \tag{10}$$

which is a fourth order equation with respect to ω^2. Since the proposed antenna has been built from one unit cell ($N = 1$), possible resonant frequencies are calculated for even integer n from

$$\left(\omega^2 - \omega_{Z01}^2\right)\left(\omega^2 - \omega_{Z02}^2\right)\left(\omega^2 - \omega_{Y01}^2\right)\left(\omega^2 - \omega_{Y02}^2\right) = 0, \tag{11}$$

and for odd integer n from

$$\omega^8 - \left(B_1 + B_2 + \frac{4}{L_1 C_3}\right)\omega^6 + \left(A_1 + A_2 + B_1 B_2 + \frac{4\omega_{Z\infty}^2 + 4\omega_{Y\infty}^2}{L_1 C_3}\right)\omega^4$$
$$- \left(B_1 A_2 + B_2 A_1 + \frac{4\omega_{Z\infty}^2 \omega_{Y\infty}^2}{L_1 C_3}\right)\omega^2 + A_1 A_2 = 0. \tag{12}$$

In our previous paper [26], new closed-form solutions have been presented to determine the L-C lumped circuit elements from cutoff frequencies of the desired dispersion diagram for a symmetric E-CRLH unit cell in Fig. 2. In this paper, an asymmetric E-CRLH unit cell in Fig. 1 is chosen to design the antenna because of smaller configuration than a symmetric E-CRLH unit cell. Closed-form solutions can still be derived for the asymmetric E-CRLH unit cell by setting

$$L_1 = 2L_1',\ C_1 = C_1'/2,\ L_2 = 2L_2',\ C_2 = C_2'/2,\ L_3 = L_3',\ C_3 = C_3',\ L_4 = L_4',\ C_4 = C_4', \tag{13}$$

or

$$L_1 = L_1',\ C_1 = C_1',\ L_2 = L_2',\ C_2 = C_2',\ L_3 = L_3'/2,\ C_3 = 2C_3',\ L_4 = L_4'/2,\ C_4 = 2C_4'. \tag{14}$$

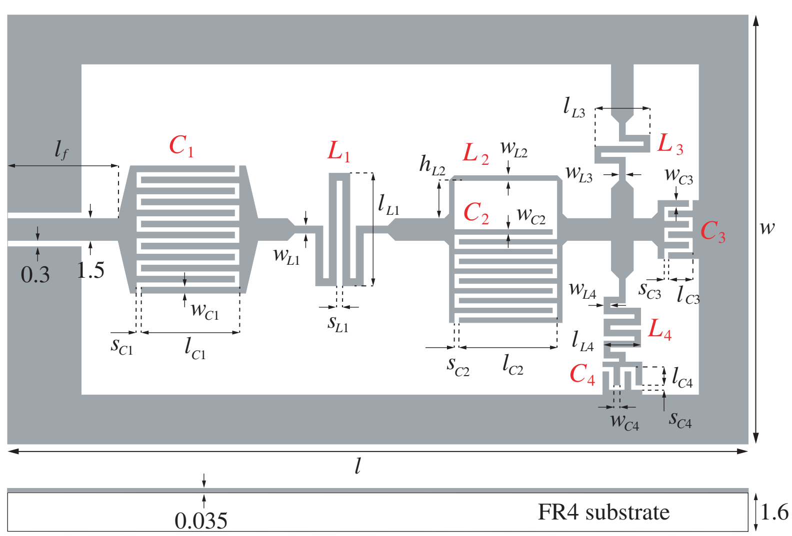


Figure 3. The configuration of the proposed antenna.

These settings lead to $Z_h Y_v/2 = Z'_h Y'_v$, then one gets the same dispersion diagram between asymmetric and symmetric E-CRLH unit cells.

Configuration of the proposed antenna has been built from one asymmetric E-CRLH unit cell in Fig. 3. A CPW configuration with an FR4 substrate with dielectric constant $\varepsilon_r = 4.4$, $\tan\delta = 0.02$, 1.6 mm substrate thickness and 35 µm copper layer thickness is used in this study. As can be seen from this figure, the antenna structure can be expressed by lumped elements. Capacitors C_1, C_2, C_3 and C_4 are formed by interdigital capacitors while inductors L_1, L_2, L_3 and L_4 are realized by meander strip lines. Dimensions of the proposed antenna are shown in Table 1. Design equations to calculate the capacitance value of interdigital capacitors and inductance value of the meander strip lines can be found in [28]. Then we get $L_1 = 25.9$ nH, $C_1 = 2.62$ pF, $L_2 = 22.3$ nH, $C_2 = 1.93$ pF, $L_3 = 8.36$ nH, $C_3 = 0.36$ pF, $L_4 = 12.0$ nH, $C_4 = 0.12$ pF. Using Eq. (8), a dispersion diagram of the proposed asymmetric E-CRLH unit cell is plotted in Fig. 4. The dispersion diagram is in an unbalance case with two RH and two LH bands.

One may use the dispersion diagram to determine which resonant modal index n corresponds to resonant frequencies of the proposed antenna. Possible resonant frequencies of the proposed antenna are calculated from Eq. (11) for $n = 0$ as 0.41 GHz, 1.14 GHz, 2.38 GHz and 5.12 GHz. This corresponds to $\beta d = 0$ in the dispersion diagram. Other possible resonant frequencies of the proposed antennas can be obtained from Eq. (12) and $\beta d = \pi$ in the dispersion diagram as (3.29 GHz, 5.69 GHz) for $n = +1$, and (0.32 GHz, 0.92 GHz) for $n = -1$.

Table 1. The dimensions of the proposed antenna.

Parameter	Dimension	Parameter	Dimension	Parameter	Dimension	Parameter	Dimension
l	57.2 mm	s_{C2}	0.2 mm	w_{L2}	0.2 mm	l_{C4}	1.5 mm
w	31.2 mm	l_{L1}	8.8 mm	h_{L2}	4.0 mm	w_{C4}	0.4 mm
l_f	9.0 mm	w_{L1}	0.4 mm	l_{C3}	2.0 mm	s_{C4}	0.2 mm
l_{C1}	8.0 mm	s_{L1}	0.4 mm	w_{C3}	0.4 mm	l_{L4}	2.6 mm
s_{C1}	0.3 mm	l_{C2}	8.0 mm	s_{C3}	0.2 mm	w_{L4}	0.3 mm
w_{C1}	0.3 mm	w_{C2}	0.3 mm	l_{L3}	4.3 mm	w_{L3}	0.4 mm

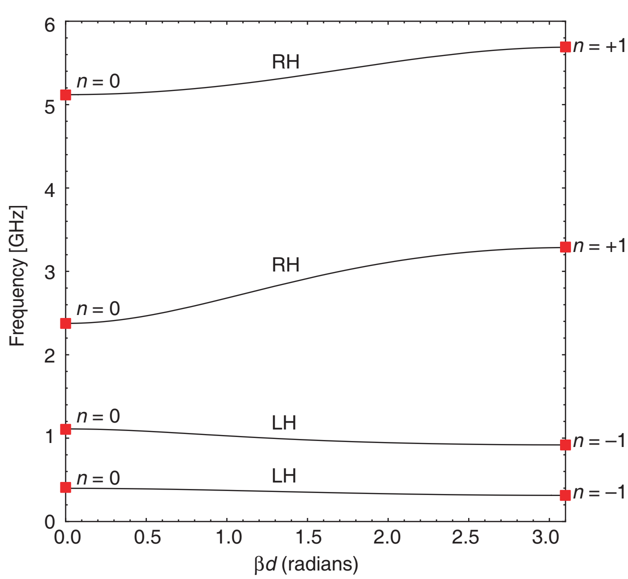


Figure 4. Dispersion diagram of the proposed asymmetric E-CRLH unit cell.

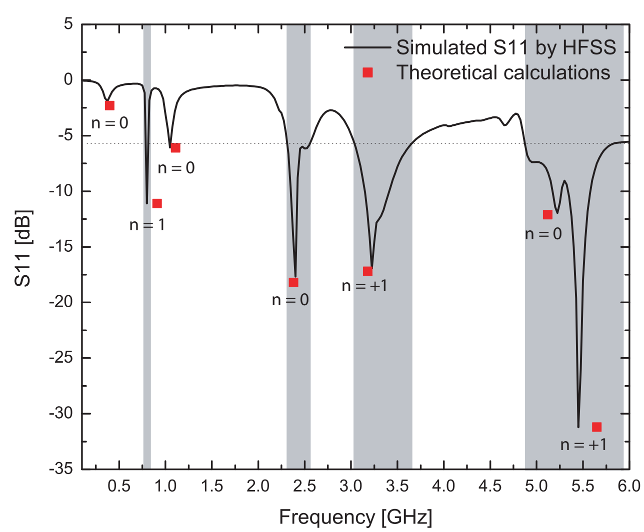


Figure 5. S_{11} characteristics of the proposed antenna.

Simulated return loss of the proposed antenna from 0.1 GHz to 6 GHz is presented in Fig. 5. Resonant frequencies of the antenna are simulated to be 0.38 GHz, 0.81 GHz, 1.05 GHz, 2.43 GHz, 3.32 GHz, 5.20 GHz and 5.45 GHz. These simulated results resemble theoretical calculations. It is noticed that 7 resonance frequencies are appeared while one theoretically predicted resonant frequency 0.32 GHz of $n = -1$ mode is not observed. The difference between simulation and theoretical calculation may come from parasitic effects by the mutual coupling between elements in the circuit. With S_{11} smaller than -6 dB, the proposed antenna exhibits four frequency bands: 0.773–0.836 GHz, 2.35–2.55 GHz, 3.05–3.71 GHz and 4.88–5.81 GHz. These frequency bands cover four standard bands including GSM810, WLAN 2.45/5.5 GHz and WiMAX 3.5 GHz bands. Fractional bandwidth of the antenna is found to be 7.78% for the first (0.81 GHz) band, 8.16% for the second (2.45 GHz) band, 18.86% for the third (3.5 GHz) band and 16.91% for the fourth (5.5 GHz) band.

3. FABRICATION, MEASUREMENT RESULTS AND DISCUSSIONS

A photograph of the fabricated antenna is shown in Fig. 6. Fig. 7 presents the measured return loss S_{11} of the antenna from 0.5 to 6 GHz. Measurements are executed by Agilent E8361A network analyzer in an anechoic chamber and compared with the corresponding HFSS simulated results.

The first band (GSM810) and second band (WLAN 2.45 GHz) have been shifted by 50 MHz and 120 MHz, respectively. At the third band (WiMAX 3.5 GHz) and fourth band (WLAN 5.5 GHz), the measured bandwidth is slightly wider than the simulated bandwidth. The normalized radiation patterns of the antenna at different frequencies are shown in Fig. 8. As can be seen from Figs. 8(a) and (b), the measured radiation patterns at 0.81 GHz and 2.45 GHz are not so good to compared with simulated ones as a result of shifting of the resonant frequency of GSM810 and WLAN 2.45 GHz bands. Nevertheless, the measured radiation patterns at these frequencies show omnidirectional radiation patterns which are suitable for wireless communications. At WiMAX 3.5 GHz and WLAN 5.5 GHz bands, a better agreement can be observed between measured radiation patterns and the simulated ones at 3.5 GHz and 5.5 GHz in Figs. 8(c) and (d). The shifting frequency in S_{11} and the difference between simulated and measured radiation patterns may come from unstable FR4 substrate parameters and the manufacturing tolerance of antenna dimensions. In addition, the SMA connector also affects the measured results because of connection loss between the board and SMA connector. Gains of the proposed antenna are estimated as 3.66 dBi at 5.5 GHz, 1.46 dBi at 3.5 GHz, -1.31 dBi at 2.45 GHz and -8.12 dBi at 0.81 GHz. Due to the compact size, gains of the antenna are quite low at low frequencies.

Table 2 summarizes recently reported works about quad-band antennas including our design. Our quad-band antenna has an electrical size $0.15\lambda_0 \times 0.08\lambda_0$ at the center frequency (0.805 GHz) of the

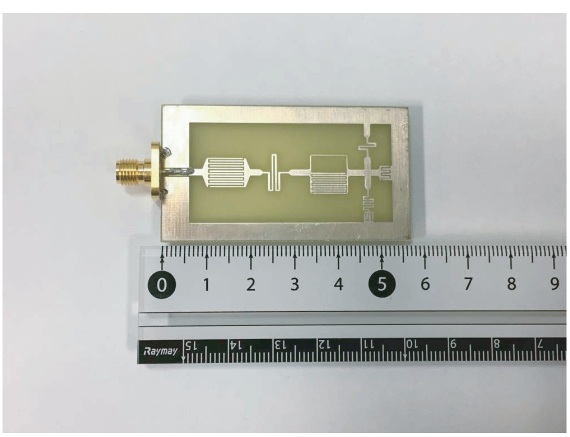


Figure 6. Fabricated antenna.

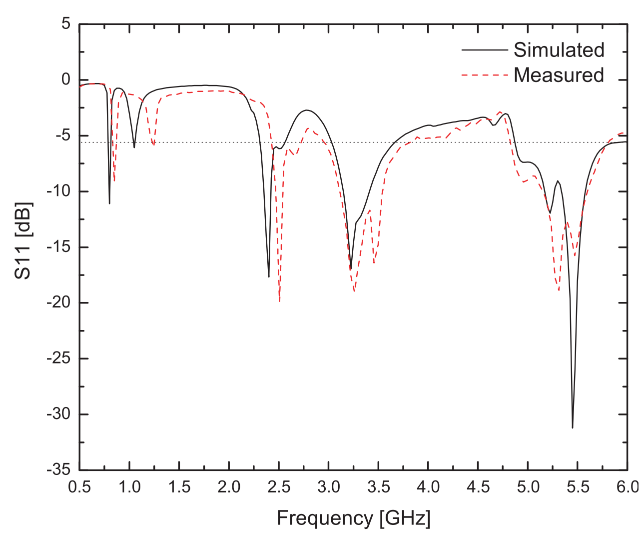
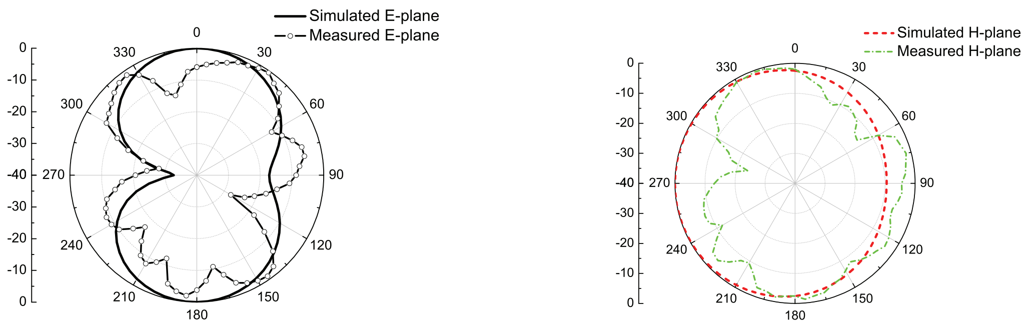
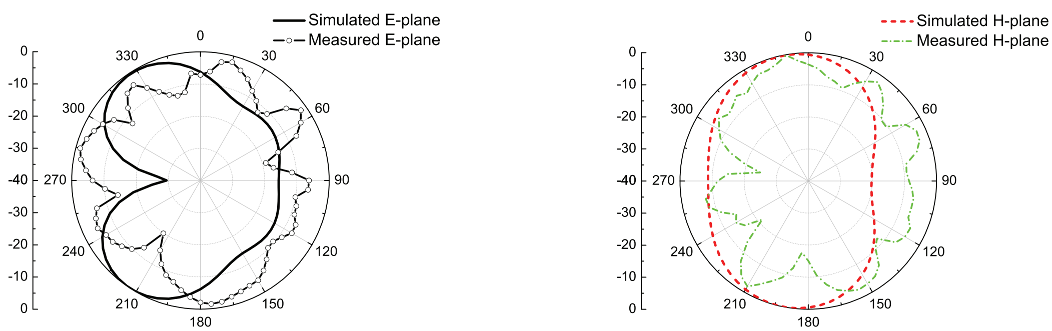


Figure 7. S_{11} characteristics of the fabricated antenna.

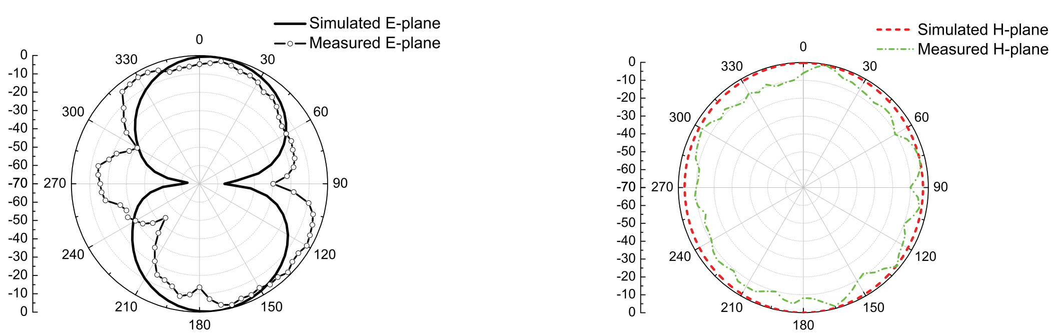
lowest band (GSM810), and its size is very small compared with other antennas [2–9] designed by conventional methods. Although that in [25] has been made from an E-CRLH unit cell, our antenna can be designed by roughly one half size. The antenna in [29] has a similar electrical size to the proposed



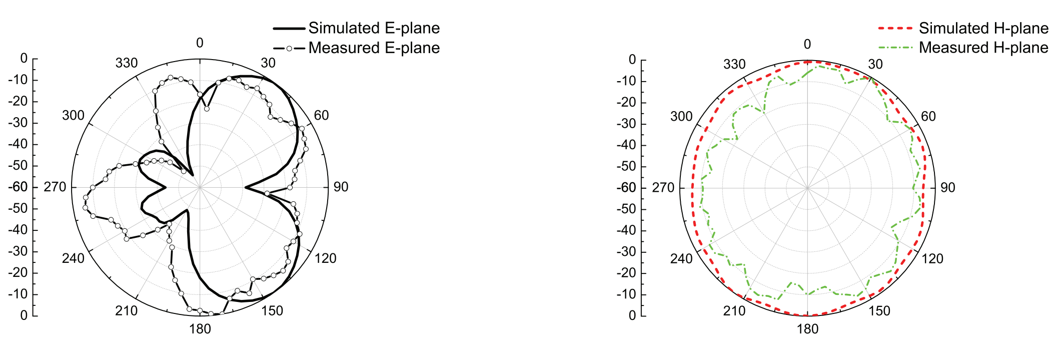
(a)



(b)



(c)



(d)

Figure 8. Normalized radiation patterns at different frequencies. (a) 0.81 GHz, (b) 2.45 GHz, (c) 3.5 GHz, (d) 5.5 GHz.

Table 2. The comparison with the previously proposed quad-band antennas.

Ref. Year	Size [mm × mm]	Electrical size by lowest band	Operating bands [GHz]	Gain [dBi]	Substrate, parameters	Design methods
Ours 2017	57.2×31.2	$0.15\lambda_0 \times 0.08\lambda_0$, $\lambda_0 = 373\,\text{mm}$	0.773–0.836, 2.35–2.55, 3.05–3.71, 4.88–5.81	−8.12, −1.31, 1.46, 3.66	FR4 substrate, $\varepsilon_r = 4.4$, $\tan\delta = 0.02$, $h = 1.6\,\text{mm}$.	E-CRLH unit cell, CPW.
[25] 2017	64×56	$0.19\lambda_0 \times 0.17\lambda_0$, $\lambda_0 = 338\,\text{mm}$	0.862–0.912, 1.75–2.69, 3.40–3.69, 4.4–6.0	−3.38, 1.5, 1.6, 3.8	Rogers 5880, $\varepsilon_r = 2.2$, $\tan\delta = 0.0009$, $h = 1.575\,\text{mm}$.	E-CRLH unit cell, Microstrip line.
[4] 2015	56×44	$0.30\lambda_0 \times 0.24\lambda_0$, $\lambda_0 = 185\,\text{mm}$	1.575–1.665, 2.4–2.545, 3.27–3.97, 5.17–5.93	3.55, 3.93, 5.02, 4.86	$\varepsilon_r = 3.5$, $\tan\delta = 0.004$, $h = 0.8\,\text{mm}$.	T-shaped stubs, E-shaped stubs, Microstrip line.
[8] 2015	90×60	$0.32\lambda_0 \times 0.21\lambda_0$, $\lambda_0 = 280\,\text{mm}$	0.94–1.20, 2.23–2.43, 3.58–3.74, 4.93–5.29	5.47, 5.88, 1.97, 3.56	Flexible Rogers 5880, $\varepsilon_r = 2.2$, $\tan\delta = 0.0009$, $h = 0.127\,\text{mm}$.	Pentangle-loop radiator, CPW.
[29] 2013	31×21	$0.13\lambda_0 \times 0.09\lambda_0$, $\lambda_0 = 236\,\text{mm}$	1.25–1.28, 2.44–2.73, 3.17–3.82, 5.03–6.16	−4.88, 2.82, 1.84, 1.78	FR4 substrate, $\varepsilon_r = 4.4$, $\tan\delta = 0.02$, $h = 2.0\,\text{mm}$.	CRLH unit cell, Meander monopole, CPW.
[5] 2012	42×36	$0.23\lambda_0 \times 0.19\lambda_0$, $\lambda_0 = 185\,\text{mm}$	1.54–1.70, 2.38–2.76, 3.20–3.77, 5.12–6.25	1.84, 1.81, 2.03, 2.80	FR4 substrate, $\varepsilon_r = 4.4$, $h = 1.0\,\text{mm}$.	Wide slots, Microstrip line.
[6] 2012	25×20	$0.20\lambda_0 \times 0.16\lambda_0$, $\lambda_0 = 124\,\text{mm}$	2.07–2.77, 3.30–3.80, 5.15–5.35, 5.70–5.89	2.6, 2.0, 3.2, 2.9	FR4 substrate, ε_r=4.4, $h = 1.6\,\text{mm}$.	L-shaped slot, Rectangular slot, CPW.
[3] 2011	60×53	$0.32\lambda_0 \times 0.28\lambda_0$, $\lambda_0 = 190\,\text{mm}$	1.525–1.625, 2.332–2.495, 3.42–3.7, 5.05–5.94	−1.0, 3.25, 2.0, 4.5	FR4 substrate, $\varepsilon_r = 2.7$, $\tan\delta = 0.02$, $h = 0.8\,\text{mm}$.	Circular slots, Concave slot, Microstrip line.
[2] 2011	30×30	$0.14\lambda_0 \times 0.14\lambda_0$, $\lambda_0 = 218\,\text{mm}$	1.17–1.58, 2.4–2.70, 3.40–3.69, 4.70–5.50	0.51, 1.41, 2.68, 2.98	FR4 substrate, $\varepsilon_r = 4.4$, $\tan\delta = 0.02$, $h = 1.6\,\text{mm}$.	Invert-C slots, Invert-F strips, Microstrip line.
[7] 2011	86×62	$0.26\lambda_0 \times 0.19\lambda_0$, $\lambda_0 = 335\,\text{mm}$	0.86–0.93, 1.69–1.73, 2.18–2.23, 2.67–2.73	$1.0 \sim 4.0$	FR4 substrate, $\varepsilon_r = 4.7$, $h = 0.78\,\text{mm}$.	Rectangular fractal, Microstrip line.
[9] 2009	52×49	$0.15\lambda_0 \times 0.14\lambda_0$, $\lambda_0 = 345\,\text{mm}$	0.868–0.870, 1.225–1.227, 1.71–1.99, 2.40–2.48	N/A	Arlon substrate, $\varepsilon_r = 3.38$, $\tan\delta = 0.002$, $h = 0.8\,\text{mm}$.	Hilbert shapes, Microstrip line.

antenna, but that design used a thicker FR4 substrate.

It is found that the proposed antenna has low gain for low frequency bands. This characteristic is typical for electrically small antennas. At higher frequencies, gains of the antenna are enhanced and comparable with previous quad-band antennas. In addition, the proposed antenna has a better gain at low frequencies than the reported dual-band NRI-TL antenna (-17 dBi at 0.9 GHz and -8 dBi at 2.4 GHz) in [24].

4. CONCLUSION

A compact quad-antenna has been designed from one E-CRLH TL unit cell. Resonant frequencies of the proposed antennas have been predicted from theoretical L-C lumped elements and compared with simulated values as well as measured ones. Furthermore, the proposed antenna shows advantages of small size, omnidirectional radiation characteristics, and easy fabrication of single copper layer on a low cost FR4 substrate. The antenna has acceptable gain except at 0.773–0.836 GHz which should be improved and is under investigation. Basically, the antenna is a candidate for wireless communications.

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