修士論文要旨 (2015 年度)

WLANおよびWiMAX用コンパクトメタマテリアルアンテナの研究

A Compact Tri-band Metamaterial Antenna for WLAN and WiMAX Applications

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1. Introduction

Nowadays, wireless communication technologies have been becoming more and more popular and wireless antennas are integrated in many electronic devices. Recently, social demand on multimedia communication has been rapidly increasing due to the development of modern wireless communication systems such as Wi-Fi, WiMAX, 3G/4G. Along with these applications, modern antennas are required to be a compact size, multiband, light weight and easy fabrication. However, the typical difficulties, encountered when designing compact antennas, include: narrow bandwidth, impedance matching to a low radiation resistance, and low radiation efficiency, so that they are difficult to meet the requirements of modern antennas. There are several techniques to reduce the size of antennas, such as incorporating a shorting pin in a microstrip patch, using short circuit, and cutting slots in the radiating patch [1] – [3].

In recent years, the Composite Right/Left Handed Transmission Line (CRLH TL) metamaterials are introduced as one of the methods for implementing compact antennas [4]. The CRLH TL metamaterials using zero-th order resonance (ZOR) are popular for their inherent multiband property which can be extensively exploited to meet various demands. Another advantage of the CRLH TL is that the resonant frequency is independent of the physical dimensions of the antenna at the zero-th order mode. This property is exploited in the miniaturization of antenna size. The CRLH TL has been recommended in a large number of antenna designs as well as microwave components [5] – [7]. In addition, Caloz proposed the equivalent circuit models of dual CRLH TL (D-CRLH TL) in 2006, exhibiting a left-handed band at the high frequencies and a right-handed band at the low frequencies [8]. This provides new way to design multiband antenna with good electromagnetic performances [9]–[10]. In this paper, a tri-band metamaterial antenna for WLAN and WiMAX applications has been designed using D-CRLH TL.

2. Antenna analysis and design

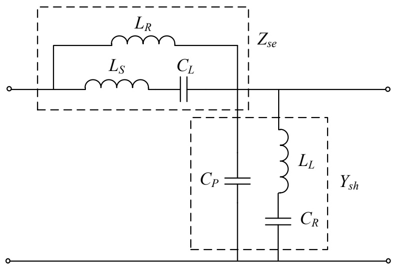

Figure 1: Equivalent circuit of the D-CRLH TL

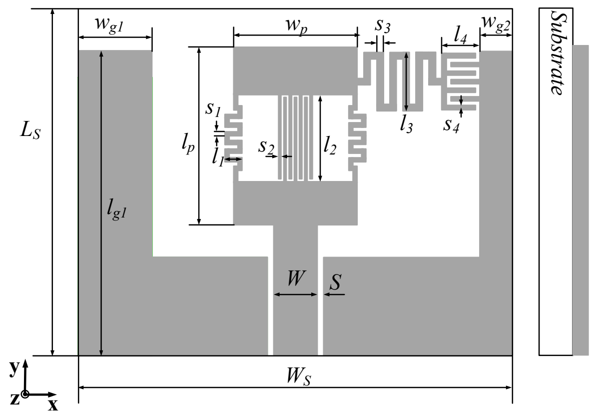

Figure 2: The configuration of the proposed antenna

2.1 Theoretical Analysis

The equivalent circuit model for a D-CRLH transmission line is shown in Fig. 1. The transmission line consists of a series circuit of left-handed capacitor C_L, right-handed inductor L_R, parasitic inductance L_S, and a shunt circuit of right-handed capacitor C_R, left-handed inductor L_L and parasitic capacitance C_P. The resonant frequency of the antenna can be calculated from the equivalent circuit. By using the Bloch-Floquet theorem, the dispersion relation of the CRLH TL unit cell for the propagation constant β can be obtained as

$$\beta d = \cos^{-1}\left(1 + \frac{Z_{se}Y_{sh}}{2}\right) \tag{1}$$

In the above equation, d represents the periodic length of the unit cell, and Z_{se}, Y_{sh} represent the series impedance and shunt admittance, respectively. The dispersion relation can therefore be determined using the following formulas:

$$\beta d = \cos^{-1}\left[1+\frac{\omega^2}{2\omega_S^2}\frac{1-\dfrac{\omega^2}{\omega_0^{se2}}}{1-\dfrac{\omega^2}{\omega_\infty^{se2}}}\frac{1-\dfrac{\omega^2}{\omega_0^{sh2}}}{1-\dfrac{\omega^2}{\omega_\infty^{sh2}}}\right], \quad (2)$$

with

$$\omega_S = \frac{1}{\sqrt{L_R(C_R+C_P)}}, \quad \omega_0^{se} = \frac{1}{\sqrt{L_S C_L}}, \quad (3)$$

$$\omega_\infty^{se} = \frac{1}{\sqrt{C_L(L_R+L_S)}}, \quad \omega_\infty^{sh} = \frac{1}{\sqrt{C_R L_L}}, \quad (4)$$

$$\omega_0^{sh} = \frac{1}{\sqrt{\dfrac{L_L C_R C_P}{C_R+C_P}}}. \quad (5)$$

Then, the resonant mode of the D-CRLH TL can be obtained by the following condition:

$$\beta d = n\pi \ \ (n = 0, \pm 1, \pm 2, ...). \quad (6)$$

From Eq. (6), the zeroth oder mode propagates in the condition $n = 0$. Therefore, the equation (6) can be expanded as

$$\cos^{-1}\left[1+\frac{\omega^2}{2\omega_S^2}\frac{1-\dfrac{\omega^2}{\omega_0^{se2}}}{1-\dfrac{\omega^2}{\omega_\infty^{se2}}}\frac{1-\dfrac{\omega^2}{\omega_0^{sh2}}}{1-\dfrac{\omega^2}{\omega_\infty^{sh2}}}\right] = 0. \quad (7)$$

By solving Eq. (7), we can get the zeroth order mode of the D-CRLH TL unit cell at the frequencies:

$$\omega = \omega_0^{se}, \quad (8)$$

$$\omega = \omega_0^{sh}. \quad (9)$$

The first oder mode of the D-CRLH TL propagates in the condition $n = 1$, then the Eq. (6) becomes

$$\frac{\omega^6}{\omega_0^{se2}\omega_0^{sh2}} - \omega^4\left(\frac{1}{\omega_0^{se2}}+\frac{1}{\omega_0^{sh2}}-\frac{4\omega_S^2}{\omega_\infty^{se2}\omega_\infty^{sh2}}\right) + \omega^2\left(1-\frac{4\omega_S^2}{\omega_\infty^{se2}}-\frac{4\omega_S^2}{\omega_\infty^{sh2}}\right)+4\omega_S^2 = 0. \quad (10)$$

Then, solving Eq. (10), we can get the first order resonant frequencies .Therefore, by adjusting the values of the lumped elements, the resonant frequencies of the D-CRLH TL can be controlled.

In order to design a tri-band antenna for WLANs and WiMAX applications, the resonant frequencies of the proposed antenna are chosen as f_1 = 2.45 GHz, f_2 = 3.60 GHz and f_3 = 5.60 GHz. Here, frequency f_1 can be determined by the first order mode defined from Eq. (10)

$$\frac{\omega_1^6}{\omega_0^{se2}\omega_0^{sh2}} - \omega_1^4\left(\frac{1}{\omega_0^{se2}}+\frac{1}{\omega_0^{sh2}}-\frac{4\omega_S^2}{\omega_\infty^{se2}\omega_\infty^{sh2}}\right) + \omega_1^2\left(1-\frac{4\omega_S^2}{\omega_\infty^{se2}}-\frac{4\omega_S^2}{\omega_\infty^{sh2}}\right)+4\omega_S^2 = 0, \quad (11)$$

with $\omega_1 = 2\pi f_1$. The resonant frequencies f_2, f_3 may be defined from the zeroth order mode, can be expressed as

$$f_2 = \frac{\omega_0^{se}}{2\pi} = \frac{1}{2\pi\sqrt{L_S C_L}} = 3.60\times 10^9, \quad (12)$$

$$f_3 = \frac{\omega_0^{sh}}{2\pi} = \frac{1}{2\pi\sqrt{\dfrac{L_L C_R C_P}{C_R+C_P}}} = 5.60\times 10^9. \quad (13)$$

Then, we can assume two cut-off frequencies f_{c1}, f_{c2} defined as ω_∞^{se}, ω_∞^{sh} are equal to 2.3 GHz and 3.0 GHz, respectively. These frequencies f_{c1}, f_{c2} can be expressed as:

$$f_{c1} = \frac{\omega_\infty^{se}}{2\pi} = \frac{1}{2\pi\sqrt{(L_S+L_R)C_L}} = 2.30\times 10^9, \quad (14)$$

$$f_{c2} = \frac{\omega_\infty^{sh}}{2\pi} = \frac{1}{2\pi\sqrt{L_L C_R}} = 5.60\times 10^9. \quad (15)$$

Therefore, from Eqs. (11) to (15), the lumped parameters can be calculated by setting L_R= 3.2 nH as L_R = 3.2 nH, C_P = 0.11 pF, C_R = 0.28 pF, L_S = 2.5 nH, L_L = 10.2 nH, C_L = 0.79 pF, respectively.

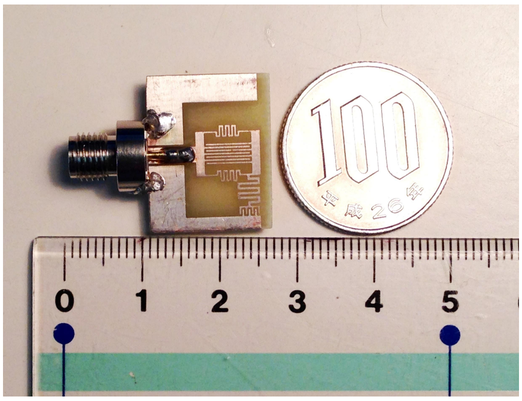

Figure 3: The prototype of the proposed antenna

Table 1: Dimensions of the proposed antenna

Parameter	Value	Parameter	Value
L_s	16 mm	l_1	1.50 mm
W_s	20 mm	s_1	0.20 mm
W	2.0 mm	l_2	6.0 mm
s	0.30 mm	s_2	0.20 mm
l_{g1}	14 mm	l_3	2.90 mm
w_{g1}	2.0 mm	s_3	0.30 mm
w_{g2}	3.40 mm	l_4	1.25 mm
l_p	8.20 mm	s_4	0.20 mm
w_p	5.70 mm		

2.2 Antenna Design

The antenna considered in this study is designed on a FR4 substrate with following parameters: relative permittivity $\epsilon_r = 4.4$, dielectric tangential loss $\delta = 0.02$, and thickness $h = 1.6$ mm. The antenna is a printed monopole type with a loaded D-CRLH metamaterial unit cell. The idea of loading metamaterial unit cell onto a conventional monopole is to impose a left handed property onto the antenna, so that the overall design can be downsized. The antenna is fed by a 50 Ω coplanar waveguide to integrate with the communication circuits. The geometrical configuration of the proposed antenna is shown in Fig. 2.

Applying the TL theory on this structure, it can be seen that this structure is a D-CRLH TL whose equivalent circuit model may be expressed by lumped elements of inductors and capacitors, as shown in Fig. 1. The series and shunt inductances L_R, L_L are realized by meandered line inductors. The inter-digital capacitors are responsible for the capacitances C_L, C_R and parasitic inductance L_S. On the other hand, the parasitic capacitance C_P is formed between the monopole and the ground plane. Hence, the electrical size of the D-CRLH TL unit cell may be reduced by increasing the lumped elements L_R, L_L, C_L, C_R, L_S, C_P, which can easily be obtained by adjusting the dimensions of the meander line inductor and inter-digital capacitors. In this structure, the physical dimensions of the elements are roughly determined by the theory, then they are analyzed by the HFSS simulator.

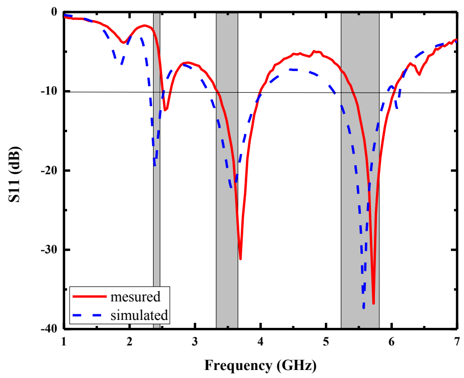

Figure 4: The simulated and measured S11

3. Experimental results

Firstly, AutoCAD DXF files are exported from HFSS simulator. These files are used to layout antenna structure. Then gerber files are the final data to make antenna. The proposed antenna was made carefully by using Quick Circuit QC5000S-E machine. The photograph of the fabricated antenna is shown in Fig. 3. You can see the fabricated antenna has very small size (16mm×20mm). The return loss, radiation patterns of this antenna are measured with E8361A network analyzer in an anechoic chamber.

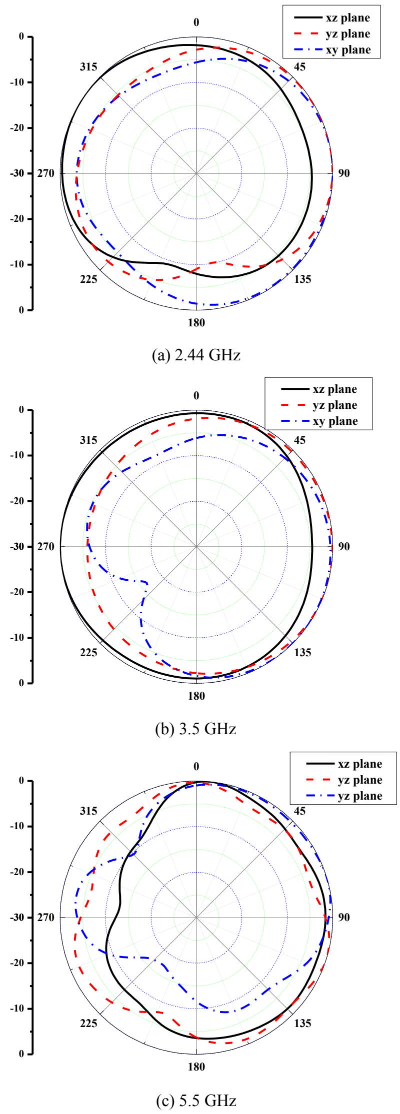

Figure 5: The simulated radiation pattern of the proposed antenna at different frequencies

The simulated and measured reflection coefficients of the proposed antenna are plotted in Fig. 4. The simulated result exhibits the S11 performance below −10 dB in three frequency bands: 2.31–2.52 GHz, 3.2–4.03 GHz and 5.11–5.93 GHz, covering the WLAN and WiMAX bands. The measured result exhibits the S11 performance below −10 dB in three frequency bands: 2.51–2.64 GHz, 3.34–4.03 GHz and 5.40–6.03 GHz. High relevance with the simulated result has been found. The minor

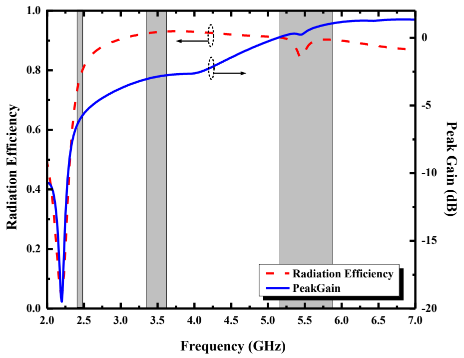

Figure 6: The simulated gain and radiation efficiency of the proposed antenna

difference observed is due to the effect of the connector of the feeding coaxial line and the substrate uniformity, which can cause calibration errors.

The radiation patterns of the proposed antenna for different frequencies are shown in Fig. 5. Due to the effect of the asymmetric configuration, the radiation pattern of the proposed antenna is not isotropic. The simulated gain and radiation efficiency at the operation band is presented in Fig. 6. The radiation efficiency of the antenna is above 0.72 for all operation frequencies and reaches the maximum value 0.93 at 3.60 GHz. The antenna gain varies from –6.50 to 1.17 dB within the frequency band of 2.4–6.0 GHz. The gains of –6.0, –2.78 and 0.68 dB are obtained at 2.45, 3.60 and 5.60 GHz, respectively. The low gain is as a result of out of phase current along the interdigital capacitor which cancels out the radiation in the far field.

4. Conclusion

In summary, D-CRLH TL and its ZOR have been investigated and explained. The compact multiband metamaterial antenna is presented in this study. The proposed antenna is based on D-CRLH TL which employs metamaterial loading on a conventional monopole to attain a certain degree of miniaturization. The antenna is fabricated, and the measurement of S11 parameter shows good agreement with the simulation results. The antenna has three operating bands at 2.45, 3.60, and 5.60 GHz for the WLANs and WiMAX applications.

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