修士論文要旨 (2014 年度)

WLAN 用小形メタマテリアルアンテナについての研究

Compact metamaterial antennas for WLAN application

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1. Introduction

In recent years, wireless communication technologies have been employed in large range of applications, which brings many economical profits as well as the convenience for life. This rapid progress requires the development of lightweight, compact, cheap and efficient wireless devices. Therefore, the necessary components in wireless devices become smaller. When miniaturizing the RF module, antenna is one of most difficult components to reduce the size. Because the conventional antenna size is depend on the operational frequency. Metamaterial technologies provide an opportunity to reduce the size of antennas while maintaining good performance at low cost because they exploit the unconventional electromagnetic properties. Metamaterial was first introduced by Veselago in 1968 [1]. With non-existence in nature and lack of experimental verification, metamaterial was not attractive over 30 years in science community. In 1996, Pendry demonstrated metamaterial [2]. Then metamaterials have attracted a lot of attentions in both theoretical exploration and experimental study. With metamaterial transmission line approach, composite right/left-handed transmission line (CRLH TL) is a conceptual way to design metamaterial antennas. CRLH TL exhibits an unusual characteristic that is called Zeroth Order Resonance (ZOR). ZOR of metamaterial structures can be applied to realize small antennas because the resonance frequency becomes independent of the antenna size [3].

In this thesis, the concept of CRLH TL is established and used to develop a compact metamaterial for WLAN application.

2. Antenna analysis and design

The configuration of a compact metamaterial antenna is illustrated in Fig. 1. In top layer of the antenna, a microstrip line is used to feed a circular patch on the same substrate. The circular patch has a circular gap g_1, which gives us series capacitance C_1. The mushroom structure is established by a circular patch and a metallic via of radius R_{via} at the center. This mushroom structure gives us a loading of distributed capacitances and shunt inductances, which are important elements for left handed metamaterial properties. In bottom layer, two metallic meander strips connect the circular patch and ground plane. According to Ref. [4], one solution to increase the bandwidth without increase the antenna's size, is to truncate the ground plane underneath the unit cell. The shape of the metallic meander strip will be designed in accordance with the antenna characteristics. Applying the transmission line theory on this structure, it can be seen that this structure is CRLH TL whose equivalent circuit model may be expressed by lumped elements of inductors and capacitors, as shown in Fig. 2. The inductors are formed by circular patch (L_1, L_2), the metallic via (L_3), and the metallic meander strips (L_4). The capacitors are formed by the gap between the circular patches (C_1), and the thickness of the substrate (C_2). Then the resonance frequency of the antenna will be calculated from this equivalent circuit. By using the Bloch-Floquet theorem, the dispersion relation of the CRLH TL unit cell for propagation β can be obtained as [5]:

$$\cos\beta p = 1 - \frac{1}{2}\left(\frac{\omega_L^2}{\omega^2} + \frac{\omega^2}{\omega_R^2} - \frac{\omega_L^2}{\omega_{se}^2} - \frac{\omega_L^2}{\omega_{sh}^2}\right), \tag{1}$$

where

$$\omega_L = \frac{1}{\sqrt{L_L C_L}} = \frac{1}{\sqrt{(L_3 + L_4/2)(C_0 + C_1)}}, \tag{2}$$

$$\omega_R = \frac{1}{\sqrt{L_R C_R}} = \frac{1}{\sqrt{(L_1 + L_2)C_2}}, \tag{3}$$

$$\omega_{se} = \frac{1}{\sqrt{L_R C_L}} = \frac{1}{\sqrt{(L_1 + L_2)(C_0 + C_1)}}, \tag{4}$$

$$\omega_{sh} = \frac{1}{\sqrt{L_L C_R}} = \frac{1}{\sqrt{(L_3 + L_4/2)C_2}}, \tag{5}$$

and p is the length of the unit cell, respectively. To realize an antenna with CRLH TL, a resonance condition $\beta_n = \frac{n\pi}{p}$ be applied [6]. Then the resonance frequency can be obtained from

$$\frac{\omega_L^2}{\omega_n^2} + \frac{\omega_n^2}{\omega_R^2} - \frac{\omega_L^2}{\omega_{se}^2} - \frac{\omega_L^2}{\omega_{sh}^2} = 2(1 - \cos\frac{n\pi}{N}), \quad n = 0, \pm 1, \pm 2, \ldots, \pm(N-1). \tag{6}$$

The mode number of the resonance (n) can be positive or negative integer or zero. Because the number of unit cell N is 1, the possible resonance mode number is 0. Accordingly zeroth order resonance (ZOR) whose resonant frequency is independent of physical length, be excited

$$\frac{\omega_L^2}{\omega_0^2} + \frac{\omega_0^2}{\omega_R^2} - \frac{\omega_L^2}{\omega_{se}^2} - \frac{\omega_L^2}{\omega_{sh}^2} = 0. \tag{7}$$

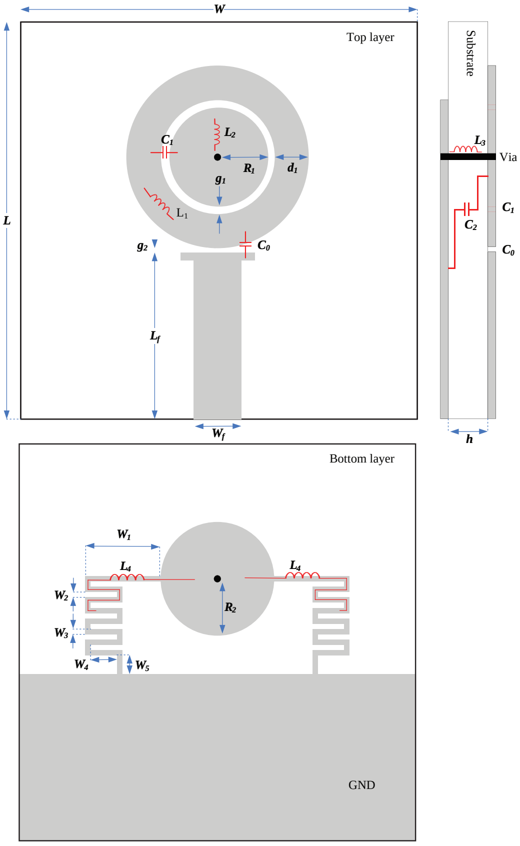

Figure 1: The geometry of the proposed antenna

Solving Eq. (7), we get two solutions for resonant frequency of proposed antenna

$$\omega_{01} = \frac{1}{\sqrt{L_R C_L}} = \omega_{se}, \tag{8}$$

$$\omega_{02} = \frac{1}{\sqrt{L_L C_R}} = \omega_{sh}. \tag{9}$$

See on the antenna structure and equivalent circuit, we determined that this is opend-ended case of ZOR. Input impedance of ZOR can be expressed as

$$\begin{aligned} Z_{in}^{open} &= -jZ_c \cot(\beta l) \overset{\beta\to 0}{\approx} -jZ_c \frac{1}{\beta l} \\ &= -j\sqrt{\frac{Z'}{Y'}}\left(\frac{1}{-j\sqrt{Z'Y'}}\right)\frac{1}{l} = \frac{1}{Y'l} \\ &= \frac{1}{Y'(Np)} = \frac{1}{NY} = \frac{1}{Y}. \end{aligned} \tag{10}$$

where

$$Y = j\left[\omega C_2 - \frac{1}{\omega(L_3 + L_4/2)}\right]. \tag{11}$$

From Eqs. (8) to (11) the resonant frequency of the antenna is

$$f = f_{sh} = \frac{\omega_{sh}}{2\pi} = \frac{1}{2\pi\sqrt{C_2(L_3 + L_4/2)}}. \tag{12}$$

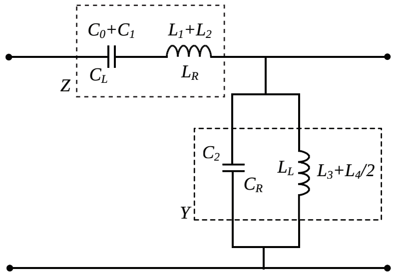

Figure 2: The equivalent circuit of the proposed antenna

As can be seen from Eq. (12), the shunt resonance frequency of the circuit depends on the values of C_2, L_3 and L_4. However, with substrate's parameters and the metallic via, the value of L_3 is approximate constant. In addition, C_2 also are not changed much in this design. Thus the only possible variable parameter for choosing resonance frequency is metallic meander strip L_4.

In oder to have lower resonance frequency, one has to increase C_2, L_3, L_4 from Eq. (12). Simplest way to increase these values, L_4 can be easily enlarged by using metallic meander strip lines without increasing the antenna's physical size. The values of C_2, L_3 and L_4 can be calculated from following formulas:

$$C_2 = 8.85 \times 10^{-3}\varepsilon_r \frac{\pi R_1^2 + \pi(R_1+g_1+d_1)^2 - \pi(R_1+g_1)^2}{h},$$

$$L_3 = 0.2\left[h \ln\left(\frac{h + \sqrt{R_{via}^2 + h^2}}{R_{via}}\right) + \frac{3}{2}\left(R_{via} - \sqrt{R_{via}^2 + h^2}\right)\right],$$

$$L_4 = 2 \times 10^{-5} l_0 \left[\ln \frac{l_0}{W_3 + t} + 1.193 + \frac{W_3 + t}{3l_0}\right] K_g,$$

where

$$K_g = \left(0.57 - 0.145 \ln \frac{W_3}{h}\right),$$

$$l_0 = W_1 + 7(W_2 + 2W_3 + W_4) + W_5.$$

Although using even more unit cells $N \geq 2$ can create lower resonance frequencies, left handed mode with index more than $(n = -1)$ can result in impedance matching issues and low radiation efficiency problems as discussed in Refs. [7, 8]. A minimum number of unit cells should be used for the antenna's design. So the proposed antenna used a single CRLH unit cell in order to obtain the best radiation performance while maintaining a small physical size for the practical application.

The antenna is simulated on a low-cost FR4 substrate with dielectric constant $\varepsilon = 4.4$, thickness of substrate $h = 1.6$ mm, thickness of copper layer $t = 18$ μm and dielectric loss tangent $\delta = 0.02$ [9]. The length of the metallic meander strips L_4 is critically related to the resonance frequency. One can observe this effect in Fig. 3, by changing the length W_1. When W_1 increases, L_4 increases and the corresponding resonance frequency decreases. The proposed antenna is designed to have zeroth order resonance at 2.44 GHz and enough bandwidth for WLAN application. After optimizing the antenna parameters, final dimensions of the antenna are shown in Tab. 1. The simulated return loss S11 is presented in Fig. 4, which shows that S11 ≤ -10 dB for entire WLAN band (2.40−2.48 GHz). The gain of the antenna is calculated from 2 to 3 GHz and illustrated in Fig. 5. The proposed antenna has peak gain of 1.5 dB in operating bandwidth. In addition, the simulated radiation patterns are plotted in Fig. 6. One can see that this antenna has omni-directional characteristic which is suitable for wireless applications.

Table 1: Dimensions of the proposed antenna

Parameter	Value	Parameter	Value
R_1	3.00 mm	L	22.00 mm
R_2	4.00 mm	W	20.00 mm
g_1	0.20 mm	L_f	10.00 mm
g_2	0.20 mm	W_1	4.85 mm
d_1	2.00 mm	W_2	0.30 mm
h	1.60 mm	W_3	0.30 mm
R_{via}	0.20 mm	W_4	2.30 mm
W_f	3.20 mm	W_5	1.05 mm

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. 7. You can see the fabricated antenna has very small size (20mm×22mm).

The return loss, radiation patterns of this antenna are measured with E8361A network analyzer in an anechoic chamber. Figure 8 presents the simulated and measured return loss of the proposed antenna. The measured line closely resembles the simulated line. The difference between simulated and measured S11 may come from substrate loss, the connection loss between the printed circuit board and SMA connector. The measured radiation pattern results have good agreement with simulated ones in Figs. 9 and 10. The antenna exhibits omni-directional radiation pattern in the E and H planes.

By observing the simulated and measured results, it is very clear that the proposed metamaterial antenna has more advantages than the conventional microstrip antennas and many previous metamaterial antennas.

4. Conclusion

In summary, CRLH TL and its ZOR have been investigated and explained. The compact metamaterial antenna was fabricated and verified its properties through

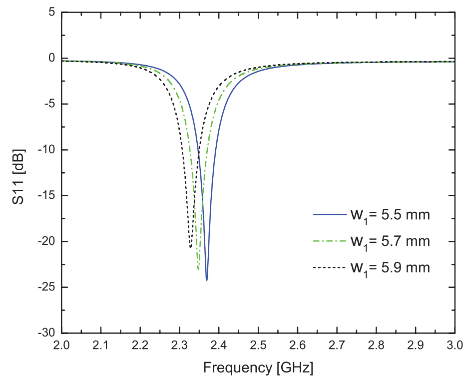

Figure 3: Simulated return loss S11 with various W_1

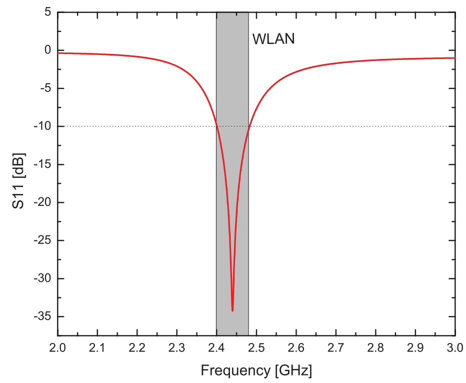

Figure 4: Simulated return loss S11 of the final design

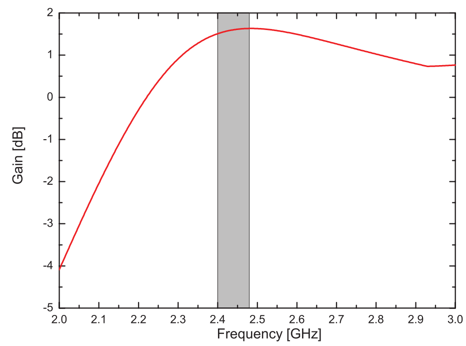

Figure 5: Gain of the proposed antenna

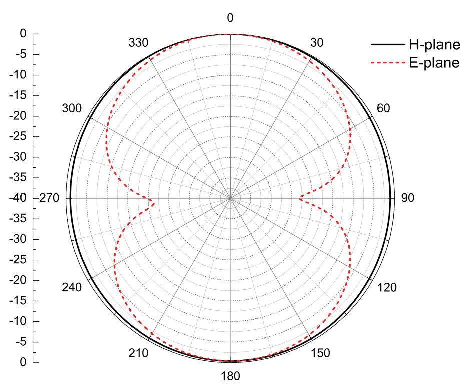

Figure 6: Normalized radiation pattern at 2.44 GHz

Figure 7: The photograph of fabricated antenna

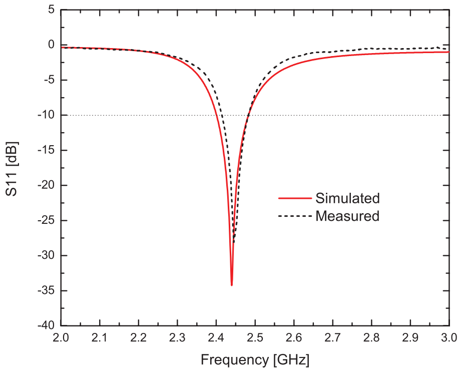

Figure 8: The measured S11

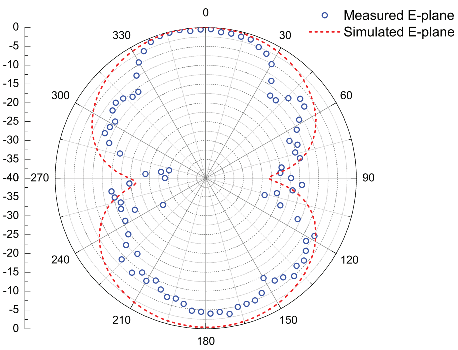

Figure 9: The measured E-plane at 2.44 GHz

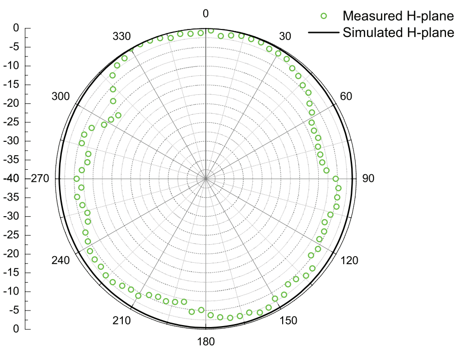

Figure 10: The measured H-plane at 2.44 GHz

measurements. Therefore, the proposed antenna can be suggested as a good candidate for wireless devices.

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