

# A Highly Miniaturized Patch Antenna Based on Zeroth-Order Resonance

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**Abstract**—A highly miniaturized antenna with electrical length  $0.036\lambda_0 \times 0.044\lambda_0 \times 0.005\lambda_0$  was implemented based on the zeroth-order resonance metamaterial approach, in which the resonance frequency becomes independent of the physical length of the antenna. The simulated gain is -12.75dBi in the 960MHz frequency, and the experimental return loss is also shown. The design methodology is carefully presented.

**Keywords**—Antenna miniaturization, zeroth-order resonator, CRLH TL.

## I. INTRODUCTION

Antenna miniaturization is a critical issue in today's wireless and communication systems. Several techniques to reduce antenna size have been proposed in the recent years [1]. One approach that has been receiving special attention is the metamaterial (MTM) based zeroth-order resonator (ZOR), due to the fact that ZOR antennas with the same physical size present better performance if compared to other techniques like high permittivity substrates, reactive elements, or shorting posts [2].

MTM's are artificial periodic structures that present unusual electromagnetic properties such as simultaneously negative electric permittivity ( $\epsilon_r$ ) and negative magnetic permeability ( $\mu_r$ ), the so called left-handed (LH) materials. Because of the singular nature of the zeroth-order mode, the resonance condition does not rely on its physical length anymore, but only on the reactance of its unit cells [3].

Reducing antenna size necessarily results in decrease of directivity, no matter what techniques are employed, and thereby of the gain. An advantage of the ZOR antenna lies in the uniformity on the distribution of the fields in the patch surface. The uniform repartition of energy along the structure leads to smaller ohmic losses than in other small antennas where high current concentrations near discontinuities produce high losses [4].

The composite right-left handed (CRLH) transmission line (TL) is a general approach that have made the design of MTM based systems much more attractive to microwave designers. It has many unique properties such as supporting a fundamental backward wave (anti-parallel group and phase velocities) and zero propagation constant at a discrete frequency ( $\beta = 0, \omega \neq 0$ ), a property that is being largely explored to the realization of novel, small *half-wavelength* resonant antennas [3].

This work presents a highly miniaturized antenna based on the techniques previously summarized, operating at the frequency of 960MHz, and implemented on the FR4 substrate.

Section II summarizes the necessary theory for the comprehension of the antenna operation, section III details the steps utilized in the design of the antenna (other works skip from theory to experimental results, not addressing details about the design methodology), section IV presents and discusses the experimental data obtained on the measurements, and a sensitivity analysis is done in order to help explain the discrepancies. A brief summary of this work and the conclusions are given in section V.

## II. THEORY

To implement a planar antenna whose resonance frequency is independent of its physical size, a CRLH TL unit-cell can be utilized, whose equivalent circuit model is shown on Fig. 1. By applying periodic boundary conditions (PBCs), the CRLH TL dispersion diagram can be determined, revealing that the structure yields LH wave propagation in low frequencies and RH wave propagation in high frequencies. The dispersion relationship is given by [4]:

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

where,

$$\begin{aligned} \omega_L &= \frac{1}{\sqrt{C_L L_L}}, & \omega_R &= \frac{1}{\sqrt{C_R L_R}} \\ \omega_{se} &= \frac{1}{\sqrt{C_L L_R}}, & \omega_{sh} &= \frac{1}{\sqrt{C_R L_L}} \end{aligned} \quad (2)$$

When  $\omega = \omega_{se}$  or  $\omega = \omega_{sh}$ ,  $\beta = 0$ . In general  $\omega_{se} \neq \omega_{sh}$  and therefore we have two frequencies with infinite wavelength propagation ( $\beta = 2\pi/\lambda$ ).

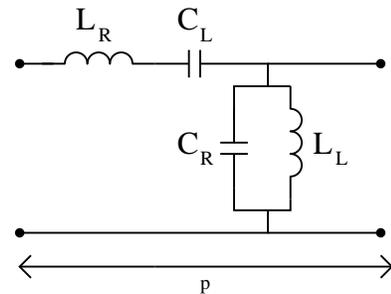


Fig. 1. Equivalent circuit model of the CRLH TL unit-cell.

Cascading an unit-cell of length  $p$ ,  $N$  times, a CRLH TL of length  $L = N * p$  can be implemented. The CRLH TL can be used as a resonator under the resonance condition [3]:

$$\beta_n = \frac{n\pi}{L} \quad (3)$$

Due to the fact that the structure supports LH propagation,  $n$  can be a positive or negative integer, or even zero. When  $n = 0$ , the structure supports a zero phase constant ( $\beta = 0$ ) and hence infinite wavelength, and the consequence is that the length of the unit-cell becomes independent of the resonant condition. If the CRLH TL is open-ended, the resonance frequency is given by  $\omega_{sh}$ , and if it's short-ended, given by  $\omega_{se}$ . Usually  $\omega_{se} > \omega_{sh}$ , so as we want miniaturization, it was chosen an open-ended configuration. Being that the case, we don't necessarily need the left-handed capacitance on the CRLH TL, and hence we will use a variation of this model, known as inductor-loaded TL. The equivalent circuit model is shown in Fig. 2, and the new dispersion relation is given by [3]:

$$\beta(\omega) = \frac{1}{p} \cos^{-1} \left( 1 + \frac{1}{2} \left( \frac{L_R}{L_L} - \frac{\omega^2}{\omega_R^2} \right) \right) \quad (4)$$

### III. ANTENNA DESIGN

To implement a radiator that emulates the equivalent circuits of Figs. 1 and 2, the well-known *mushroom structure* was the chosen topology [3]. In this structure, the left-hand capacitance  $C_L$  is due to the gaps between the microstrip patches, and the left-hand inductance  $L_L$  by the short-circuit vias. We also have the unavoidable RH effects due to the coupling between patch and ground plane and the patch distributed inductance, represented by  $C_R$  and  $L_R$ , respectively.

We want a high miniaturization factor, and, for this reason, it was convenient to use only one unit-cell. Due to the fact that the equations shown on the previous section are based on PBC's, the more the number of cells (ideally infinite), more accurate are the equations, so we can expect some variation of resonant frequency in our simulations. The inductor-loaded unit-cell resonates at a frequency given by  $\omega_{sh}$  in equation (2), and therefore our resonant frequency is approximately given by:

$$f_0 \approx \frac{1}{2\pi\sqrt{C_R L_L}} \quad (5)$$

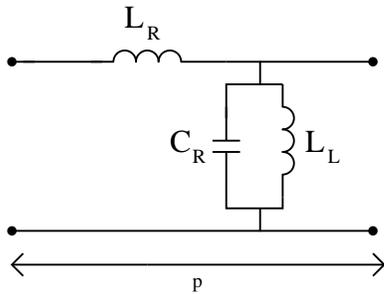


Fig. 2. Equivalent circuit model of the Inductor-Loaded TL unit-cell.

The idea here is to diminish the resonance frequency by raising the values of  $C_R$  and/or  $L_L$ .  $C_R$  is proportional to the patch area and to the relative permittivity of the substrate and inversely proportional to the thickness of the substrate. As we want to miniaturize an antenna, increasing the area is not an option. The via inductance is directly proportional to the length, or, in other words, to the thickness of the substrate. Therefore, decreasing substrate thickness to improve  $C_R$  would cause almost no effect to the resonant frequency at all, because it lowers  $L_{via}$  approximately the same amount it raises  $C_R$ . This is the reason why it is more convenient to raise the inductance instead of capacitance [3].

One option to do this is to put in series with the short-circuit via a spiral slot-inductor on the ground plane. The proposed geometry with the spiral slot-inductor is shown in Fig. 3. In that case,  $C_R$  is still the coupling capacitance between patch and ground, but  $L_L = L_{via} + L_{slot}$ , and depending upon the slot inductance we can considerably reduce the resonance frequency [5].

To design a slot-inductor for given dimensions, it is important to have knowledge about the value of the  $C_R$  capacitance. If we know  $L_L = L_{via}$  and simulate the ZOR structure, we can have an idea of what is this value and then design the ground inductance to shift the resonance frequency to 960MHz.

The software utilized to design the antenna was Agilent Advanced Design System (ADS), being Method of Moments (MoM) the method utilized for the printed inductor simulations, and Finite Element Method (FEM) for the antenna geometry simulations.

#### A. Unit-cell without slot inductor

The via inductance can be estimated using the following approximations for a round wire conductor [6]:

$$L_{via} = 0.002\ell \left[ \ln \left( \frac{4\ell}{d} \right) - 1 + \frac{d}{2\ell} + \frac{\mu_r T(x)}{4} \right] (\mu H) \quad (6)$$

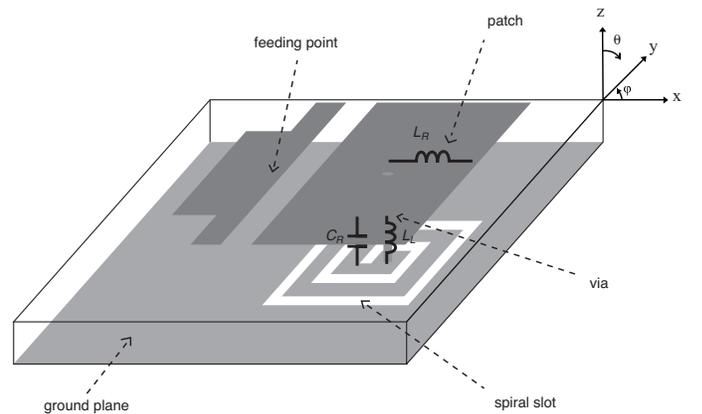


Fig. 3. One unit-cell MTM based antenna with slot inductor loading on the ground plane.

where,

$$T(x) \approx \sqrt{\frac{0.87311 + 0.00186128x}{1 - 0.278381x + 0.127964x^2}} \quad (7)$$

$T(x)$  is a compensation for the AC effects, and

$$x = \pi d \sqrt{\frac{2\mu f}{\sigma}} \quad (8)$$

Here:

$d$  = diameter in cm

$\ell$  = wire length in cm ( $\ell = t$ , the thickness of the substrate)

$f$  = frequency in Hz

$\sigma$  = conductivity of the material in S/m

Calculating and simulating for comparison, the inductance of the via is approximately  $L_L = L_{via} \approx 0.55\text{nH}$ .

The return loss simulation result for the non-loaded one unit-cell version of this structure is shown in Fig. 4. As we can see, the zeroth-order resonance occurs approximately at 3GHz, so we can extract  $C_R \approx 5.15\text{pF}$ .

### B. Slot Inductor Design

Loading the antenna with the slot inductor, the total inductance  $L_L$  becomes a series connection between the two inductances. We want the antenna to resonate in the frequency  $f_0 = 960\text{MHz}$ , so we have to design an inductor with:

$$L_{slot} \approx \frac{1}{(2\pi f_0)^2 C_R} - L_{via} \quad (9)$$

This equation gives us a value of  $L_{slot} \approx 4.8\text{nH}$ . A good closed formula to design a spiral inductor is given by [7]:

$$L_{slot} = \xi d_{out}^{\alpha_1} w^{\alpha_2} d_{avg}^{\alpha_3} n^{\alpha_4} s^{\alpha_5} \quad (10)$$

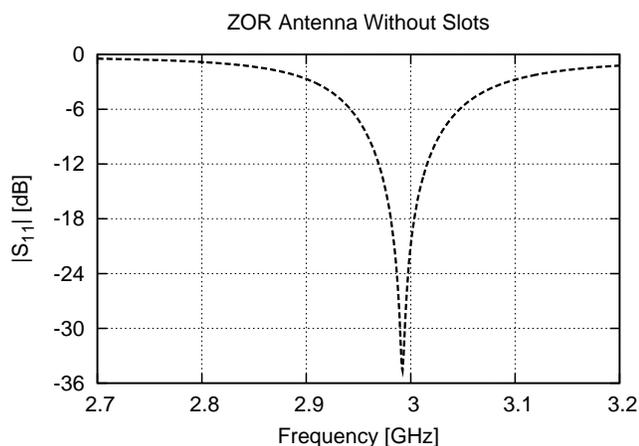


Fig. 4. Simulated reflection coefficient of the ZOR antenna without slots.

TABLE I. Coefficients for the Calculation of Inductance [7].

Layout	$\xi$	$\alpha_1(d_{out})$	$\alpha_2(w)$	$\alpha_3(d_{avg})$	$\alpha_4(n)$	$\alpha_5(s)$
Square	$1.62 \times 10^{-3}$	-1.21	-0.147	2.40	1.78	-0.030
Hexagonal	$1.28 \times 10^{-3}$	-1.24	-0.174	2.47	1.77	-0.049
Octagonal	$1.33 \times 10^{-3}$	-1.21	-0.163	2.43	1.75	-0.049

in which  $d_{avg}$  is the average diameter, and its given by:

$$d_{avg} = \frac{1}{2}(d_{out} + d_{in}) \quad (11)$$

and  $\rho$  is the filling factor:

$$\rho = \frac{d_{out} - d_{in}}{d_{out} + d_{in}} \quad (12)$$

where  $w$  is the width of the conductor,  $n$  is the number of turns,  $s$  is the space between two conductors,  $\xi$  and  $\alpha_n$  are coefficients, whose values are shown in table I. As the choice of geometry has a negligible impact to the quality factor (Q) of the inductor, we choose the square shape [8].

The slot inductor was simulated and optimized for the desired inductance. The quality factor Q of the inductor was also optimized in order to minimize the losses on the operating frequency. The curves of simulated inductance and quality factor are shown in Figs. 5 and 6. Due to the electromagnetic coupling between the components of the structure and the shift from 3GHz to 960MHz on the resonance frequency who changed the capacitance to a smaller value of  $C_R \approx 3\text{pF}$ , a higher inductance value had to be utilized on the optimization process ( $L_{slot} \approx 8.95\text{nH}$ ).

## IV. RESULTS

The square spiral slots was loaded to the ground plane and the ZOR antenna with one unit-cell was experimentally implemented on the FR4 substrate ( $\epsilon_r = 4.47$ ,  $\tan\delta \approx 0.02$ ), and it is shown on Fig. 7. The simulated gain was  $G = -12.75\text{dBi}$ .

The reflection coefficient of the antenna was measured using a XYZ vector network analyser, and it is shown in Fig. 8. As

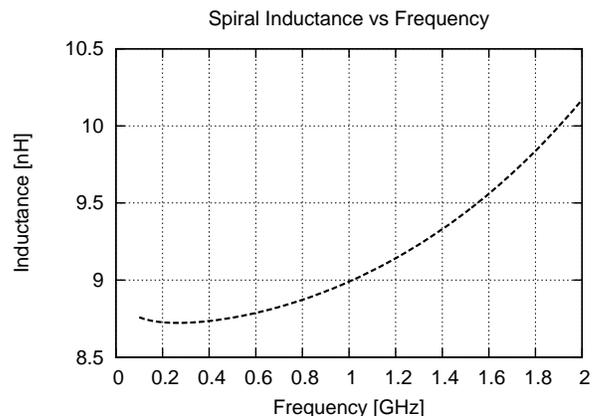


Fig. 5. Simulated inductance of the spiral slot inductor.

TABLE II. Comparison between miniaturized antennas.

	This work	[5]	[9]	[10]	[11]
Dimensions (L×W)	$0.036\lambda_0 \times 0.044\lambda_0$	$0.14\lambda_0 \times 0.23\lambda_0$	$0.021\lambda_0 \times 0.017\lambda_0$	$0.091\lambda_0 \times 0.091\lambda_0$	$0.096\lambda_0 \times 0.138\lambda_0$
Gain[dBi]	-12.75	-0.28	-38	-1.302	-4.7
Operating Frequency[GHz]	0.96	2.3	0.4	0.91	3.59

we can see, the resonant frequency has been shifted to the right by some  $\Delta f$  (90MHz). The electrical size is calculated to be  $0.036\lambda_0 \times 0.044\lambda_0 \times 0.005\lambda_0$  in the operating frequency.

The inductance of the spiral inductor is given by the complicated equation (10) but, however, in order to get some insight about the sensitivity of the inductance with the variations of the fabrication process, let us consider a very crude zeroth-order estimation which is suitable for quick hand calculations [8]:

$$L \approx \mu_0 n^2 r \quad (13)$$

where  $r$  is the radius of the spiral. This equation has a accuracy of approximately 30%.

Differentiating to take the variation with  $r$ , we get:

$$\frac{dL}{dr} = \mu_0 n^2 \quad (14)$$

Yielding a sensitivity of  $\approx 3\text{nH/mm}$  (we have used one and a half turns).

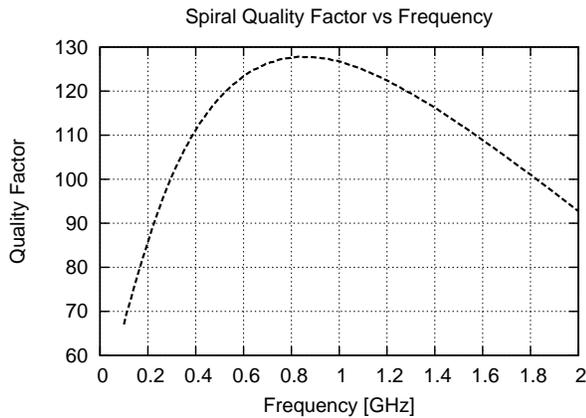


Fig. 6. Simulated quality factor of the slot spiral inductor.

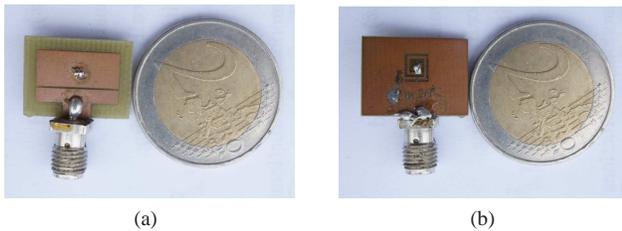


Fig. 7. Photograph of Antenna Physical Implementation (a) Upper view (b) Bottom view

If we differentiate equation (5) to estimate the sensitivity of frequency in relation to inductance  $L_L$ , we obtain:

$$\frac{\partial f_0}{\partial L_L} = \frac{-C_R}{4\pi(C_R L_L)^{3/2}} \quad (15)$$

That gives us a sensitivity of 55MHz/nH. As an example, if the fabrication process has a slight total variation on the geometry of only a half millimeter, the total variation of resonance frequency is approximately 82.5MHz. This is a reasonable explanation for the shift on the frequency.

The low gain is a consequence of the tradeoff between the efficiency and the size of an antenna, but also due to the high losses of the FR4 substrate.

Table II gives us a comparison between our work and other miniaturized antennas based on metamaterials. On [5], [10] and [11], a relatively high gain was obtained, but the miniaturization achieved is smaller. Reference [9] has achieved an extremely high miniaturization, but the gain is also considerably small.

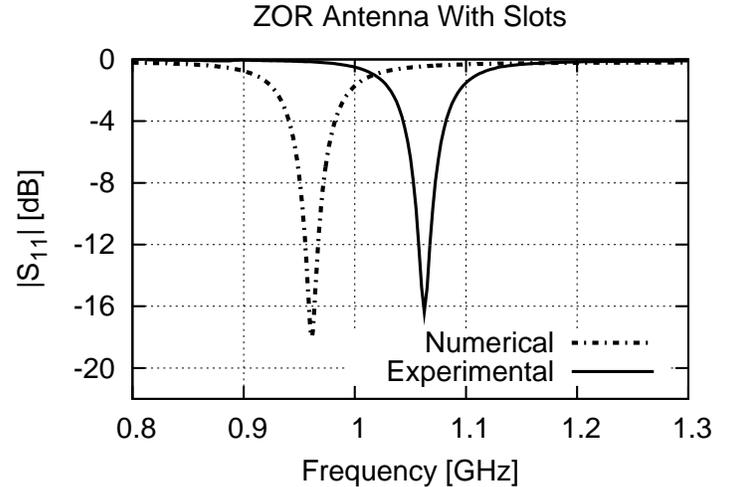


Fig. 8. Simulated and experimental reflection coefficients of the ZOR antenna.

## V. CONCLUSION

Miniaturization is a primary concern today in communication and biomedical systems. A highly miniaturized antenna with electrical length  $0.036\lambda_0 \times 0.044\lambda_0 \times 0.005\lambda_0$  ( $11.25\text{mm} \times 13.80\text{mm} \times 1.55\text{mm}$ ) was implemented based on

the zeroth-order resonance approach. The experimental resonance frequency presented a slight shift to the right due to the limited resolution of the fabrication process. The detailing of the design methodology presented here can be a useful guide to antenna designers who are working on miniaturized antennas.

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