Circularly Polarized Aperture Coupled Zeroth Order Resonance Antenna for mm-Wave Applications

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Abstract **—** In this work, a millimetre-wave aperture–coupled antenna was offered by mode synthesis. In order to increase half power beam width (HPBW), the present study united a TM_{010} mode, which was generated by aperture slot of substrate integrated waveguide (SIW), and a zeroth-order resonance (ZOR) mode of a metamaterial polarizer mushroom. The results were confirmed by an equivalent circuit and a full wave simulator. Moreover, to prevent mm-wave fed loss, the study made use of a SIW feed with a slot which reduced the interference between a feed and a mushroom metamaterial structure as a polarizer antenna. The proposed antenna was fabricated and measured at 35 GHz. The simulated results roughly agreed with the measured results.

Index Terms **—** Aperture–coupled antenna, Substrate Integrated Waveguide (SIW), Zeroth-Order Resonance (ZOR).

I. INTRODUCTION

In recent years, having a high data rate and a small size, millimetre-wave systems have provided access to less-crowded spectrums. They are a good alternative to be considered for future telecommunication systems. To enhance the accessibility and capability of mm-wave systems, the use of antennas with broad bands and high gain has become widespread [1–8]. Two main issues in designing mm-wave antenna must be considered. The first problem in designing an antenna in mm-wave application is high atmospheric debilitation, which reduces communication range [1–8]. Thus, the antenna in mm-wave application must follow either beam forming or broadside patterns. As is well known, designing beam forming feed network antenna is difficult and the risk of losing in the feed network is very high. On the other hand, despite that ordinary patch antennas follow a broadside pattern, their large size and low gain are important problems of these antennas. Second problem in designing mm-wave antennas is the high loss of feed network. Although, using a substrate with high relative permittivity can compensate microstrip radiation loss, it cannot compensate for higher losses. Additionally, making the overall system more resistant in the case of multipath propagation is another problem which must be considered in mm-wave application. Hitherto, to overcome each of the mentioned problems, previous studies have reported different methods. In [2], for example, an E-shape patch antenna was proposed for millimetre wave frequencies. This technique, though causing the bandwidth and broadside pattern to increase, has lower gain. On the other hand, the microstrip feed line and patch created in same layer lead to loss increase. Combining an aperture–coupled and a zeroth-order resonance, [3] reported a mode of a metamaterial antenna at the millimetre-wave frequency band that could almost solve the problem faced in [2]. However, there are still the problems of multipath effect and radiation loss of microstrip line. [4] employed the metallic waveguide fed microstrip array to alleviate the radiation loss. The substrate integrated waveguide (SIW) fed planar microstrip array has been introduced in [5] to overcome the limitations of conventional metallic waveguide, where a higher radiation efficiency was obtained. Nevertheless, the feed network and patch array were integrated onto the same layer; hence, the spurious radiation from the feed network was not completely suppressed [6]. The present study attempted to solve the mentioned problems by offering a solution for each. In order to have a broad side pattern with a high gain and low loss feed network, the study used a SIW cavity backed slot antenna which combining novel circularly polarizer mushroom and parasitic ring, produced a zeroth order resonance (ZOR) that was capable of increasing the broadside and gain even at a small size.

II. ANTENNA STRUCTURE

Figure 1 displays the geometry of the proposed CP SIW ZOR antenna. The antenna consists of two substrates isolated by a ground with an aperture. The top substrate is RT/Duroid 5880 with relative permittivity of ε_r=2.2,

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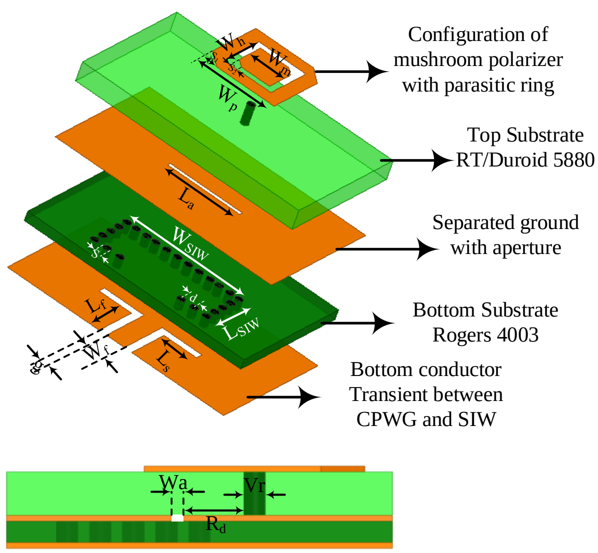
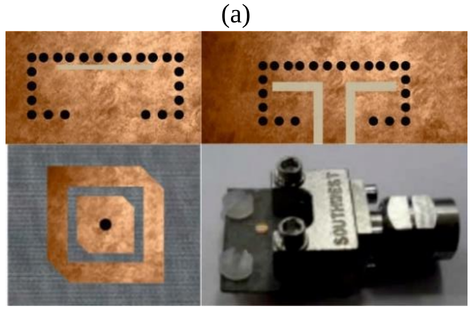
loss tangent of tanδ=0.0009, and thickness of h_1=0.508 mm. It contains the ZOR polarizer radiating element, while the other substrate, with the SIW fed line, is Rogers 4003 (ε_r=3.55, tanδ=0.002). The polarizer includes a mushroom antenna and a parasitic ring patch, the two opposite corners of each are etched by S_1 and S_2, respectively, to attain CP feature. The antenna produces two modes of TM_{01} and TM_{10} with a directional radiation pattern leading to the emergence of CP characteristic, and a ZOR mode with an omnidirectional radiation pattern, respectively. The optimal dimensions of the antenna are as follows: W_p=2 mm, W_h=1.3 mm, radius of the via V_r=0.1 mm, L_a=2 mm, and W_a=0.1 mm. The parameters were derived from full wave simulation (ANSYS HFSS). The size of the patch antenna (W_p) and that of the etched hole (W_h) are determined to set the resonance frequency at 35 GHz, while the size of the aperture (L_a) and the radius of via (V_r) are set to optimize the radiation efficiency of the mushroom antenna [3].



(a)



(b)

Fig. 1. (a) Configuration of antenna structure, and (b) fabricated photo.

III. EQUIVALENT CIRCUIT

The equivalent circuit of the antenna is shown in Fig. 2. It consists of an aperture (C_1, C_a, L_a, and R_a), a parasitic ring patch antenna (C_p, L_p, and R_p), and a mushroom (C_R, L_L, and R_m). Γ_m, Γ_p, and Γ_{pm} are the coupling coefficients between an aperture and a mushroom, an aperture and a parasitic ring patch, and a mushroom and a patch, respectively. The equivalent circuit is calculated as follows in Ref. [1-4].

The first step is to model the mushroom with a RLC network. The formulas to represent a resonator as a parallel resonant circuit can be found in [1-4] and [9] when the resonator is coupled to the excitation source:

$$Z_{surface} = \frac{j\omega L}{1-\omega^2 LC}, \tag{1}$$

where ω=2πf and f define the angular frequency and frequency of the wave, respectively. The equivalent sheet inductance, L_L, and the equivalent sheet capacitance, C_R, are given as follows:

$$L_L = \frac{\eta_s}{\omega}\tan(\beta h);\ and\ \ C_R = \frac{w\varepsilon_0(\varepsilon_{r1}+\varepsilon_{r2})}{\pi}\cosh^{-1}\left(\frac{D}{g}\right). \tag{2}$$

The properties of the characteristic parasitic patch can be obtained by using:

$$L_P = \frac{\frac{\mu_0}{2} l_{avg}}{4}\left[Ln\left(\frac{l_{avg}}{w}\right) - 2\right], \tag{3}$$

where μ_0 is the vacuum permeability, l_{avg} is the average strip length calculated over all the rings and,

$$C_P = 2\varepsilon_0\varepsilon_r^{sub}\frac{2w+\sqrt{2}g}{\pi}\operatorname{arccosh}\left[\frac{2w+g}{g}\right];\ and\ \varepsilon_r^{sub} = 1 + \frac{2}{\pi}arctg\left[\frac{h}{2\pi(w+s)}\right](\varepsilon_r - 1). \tag{4}$$

Where in (3) and (4), g, s and w are the gap between patch and parasitic patch (W_h - W_p), width of parasitic patch (W_p - W_h), and patch width (W_p), respectively. The second step is to find the input impedance of the slot. When the transmission line is terminated by a stub length, the input impedance is simply put under the rectangular waveguide supposition and this result is added as series reactance, X. The total impedance is then:

$$Z_{slot} = Z_c\frac{2R}{1-R} + X. \tag{5}$$

As is well known, the impedance of an SIW, Z_0, can be calculated by:

$$Z_0 = \frac{\frac{h}{w}\eta}{\sqrt{1-\left(\frac{\lambda}{\lambda_c}\right)}}\ and\ \eta = \frac{120\pi}{\sqrt{\varepsilon_r}}, \tag{6}$$

where Z_c is the characteristic impedance of transmission line and R is voltage reflection coefficient.

The mutual inductance between the microstrip patch and the slot is:

$$M = \left(\frac{\mu_0 W_a}{2\pi}\right) Ln\,(sec\ \theta_0)\ ;\ \theta_0 = arctan\left(\frac{L_a}{2h}\right), \tag{7}$$

where W_a, L_a and h are the slot width, the slot length, and the substrate height, respectively.

The mutual inductance between the microstrip patch and parasitic patch is:

$$M = \frac{\mu_0 x_1}{2\pi}\left[0.467 + \frac{0.059w^2}{x_1^2}\right], \tag{8}$$

where w is the patch width(W_p) and x_1 is the effective parasitic patch length.

From the equivalent circuit, the total incident power to the proposed antenna can be expressed by:

$$P_{inc-t} = P_{inc-ZOR} + P_{inc-TM010}, \quad (9)$$

where $P_{inc\text{-}ZOR}$ and $P_{inc\text{-}TM010}$ are the incident powers to the inner mushroom ZOR antenna and the outer patch antenna, respectively. Hence, $P_{inc\text{-}ZOR}$ and $P_{inc\text{-}TM010}$ can be written by [10]:

$$P_{inc-ZOR} = \frac{1}{2} R_m |i_1|^2, \quad (10)$$

$$P_{inc-TM010} = \frac{1}{2} R_p (|i_2|^2 + |i_3|^2). \quad (11)$$

By using the injected power to each antenna, the incident power ratio (P_r)can be defined as:

$$P_r = \frac{P_{inc-ZOR}}{P_{inc-ZOR} + P_{inc-TM010}}. \quad (12)$$

Consequently, in (1)-(12), the equivalent elements of the antenna are extracted as follows: C_1=12.63 fF, C_a=126.47 fF, L_a=163.91 nH, R_a=3126 Ω, C_p=49.61 pF, L_p=842.26 nH, R_p=884 Ω, C_R=84.62 pF, L_L=276.01 nH, R_m=796 Ω, Γ_m=0.146, Γ_p=0.103, and Γ_{pm}=0.078.

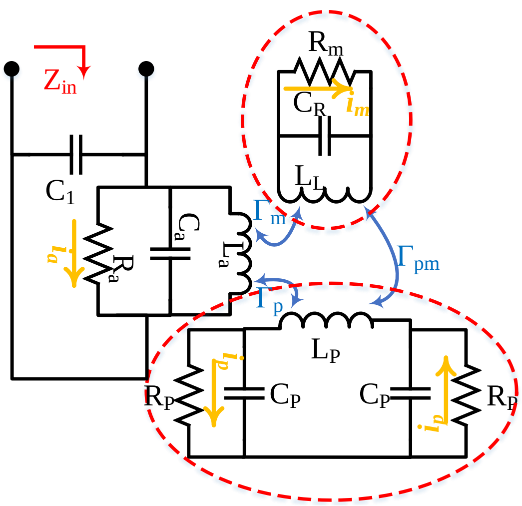


Fig. 2. Equivalent circuit of the proposed antenna.

IV. RESULTS AND DISCUSSION

All parameters of the antenna have been optimized by using commercial software high frequency structure simulator (HFSS ver. 14). In order to investigate the effect of each parameter, two parameters of Rd and WR=Wp-Wh have been studied. By changing Rd value from 0.5 to 0.55 mm, inductance effect increases and impedance matching changes, but by changing Rd value from 0.55 to 0.6 mm, the capacitance effect increases again and causes a decrease in impedance bandwidth. Results of these changes are demonstrated in Fig. 3.

As mentioned, another parameter which has important effect on the antenna results is WR. In Fig. 4, by changing this parameter from 0.2 to 0.7, the capacitance effect is reduced which leads to an increase in impedance bandwidth but by increasing inductance effect via changing W_R from 0.7 to 1.2 mm, impedance matching was disturbed.

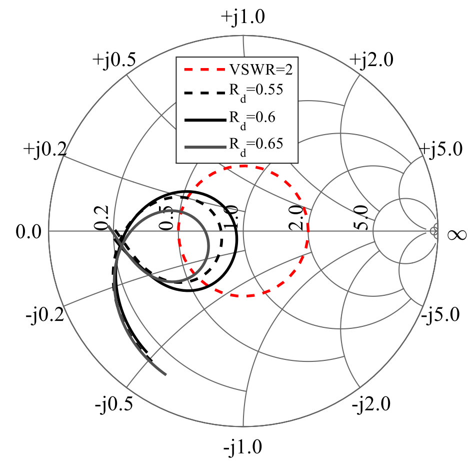


Fig. 3. Studying the effect of changing parameter R_d on impedance matching.

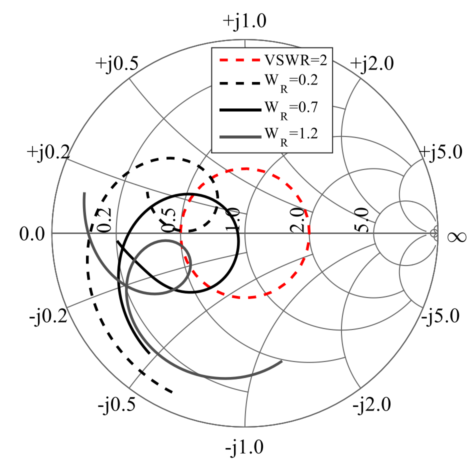


Fig. 4. Studying the effect of changing parameter W_R on impedance matching.

The proposed SIW fed CP ZOR antenna was fabricated and measured. The scattering parameters of the proposed antenna were performed with an Agilent 8510XF (E7340A) vector network analyzer. As specified in Fig. 5, the measured result for the S_{11} is in reasonable agreement with the simulated one, except for a little frequency shift [7-8] for the most ranges of the impedance bandwidth. From the measured results, a good impedance bandwidth over the frequency range of 33.82-36.37 GHz can be attained (as shown in Fig. 5).

As is well known, since an equivalent horizontal magnetic loop current is generated inside the mushroom structure at the ZOR mode, the power radiates with an omni-directional pattern [3]. If the directional radiation pattern of the combination of TM_{01} and TM_{10} modes and the omni-directional radiation pattern of the ZOR mode are combined at the same frequency, it is possible for the HPBW of the total radiation pattern to improve. On the other hand, since TM_{01} and TM_{10} are two modes with the same magnitudes and 90-degree phase difference, the proposed antenna radiated as CP antenna. In order to prove the performance principle of the design theory of the antenna, the study displayed the simulated surface current distribution on the antenna peripheral as well as the E-field distribution at the center frequency for different phases (Fig. 6). The direction of the wave in the presented antenna configuration leads to the counter clockwise rotated current distribution which results in right-handed CP (RHCP) radiation. Left-handed CP (LHCP) radiation can be obtained by changing chamber location of the presented configuration. The comparison between simulated and measured radiation patterns of the proposed antenna is demonstrated in Fig. 7. As proved in Fig. 7, this antenna with a broadside pattern radiates RHCP in +Z direction and because the loss of feed line is reduced and antenna is operated with SIW cavity backed slot, the gain is increased to 9.2 dB.

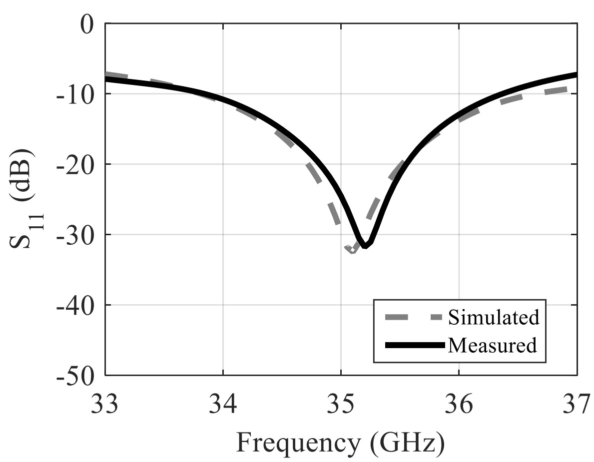


Fig. 5. The comparison between simulated and measured S_{11}.

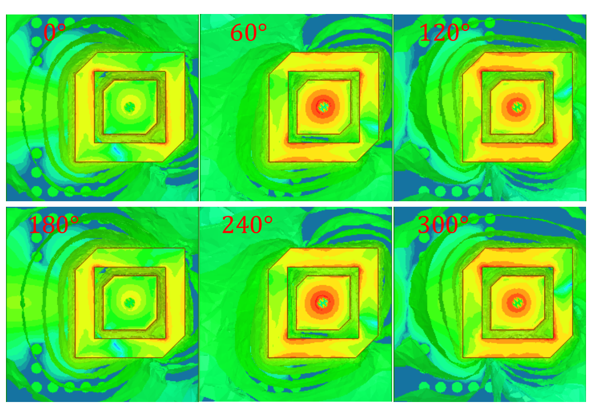


Fig. 6. Simulated current and E-field distributions on the surface of the proposed CP planar aperture ZOR antenna at 35 GHz for different phases.

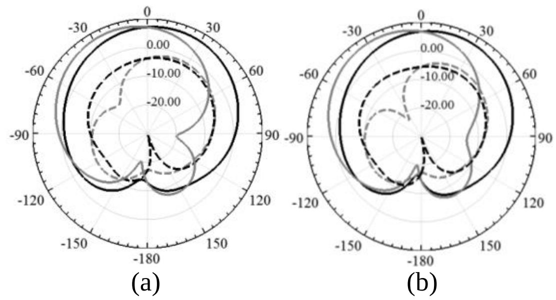


Fig. 7. The comparison between (a) simulated and (b) measured radiation patterns at 35 GHz (solid line is RHCP, dash line is LHCP, black line is φ=0°, and grey line is φ=90°).

In addition, using mushroom polarizer leads to an increase in gain with a broadside pattern. The simulated and measured half power bandwidths of antenna are 105 and 103.8 degrees, respectively. The comparison between simulated and measured results of axial ratio and gain of ZOR antenna versus frequency is illustrated in Fig. 8. The simulated 3-dB axial ratio is from 34.27 to 35.87 GHz and measured axial ratio, while matching well with simulated results, has 4.7% (34.32-35.94 GHz) bandwidth. The gain of antenna in all frequency ranges has low variant value which suggests that this antenna possesses stable pattern and results. The measured average gain of the antenna is 8.93 dBic.

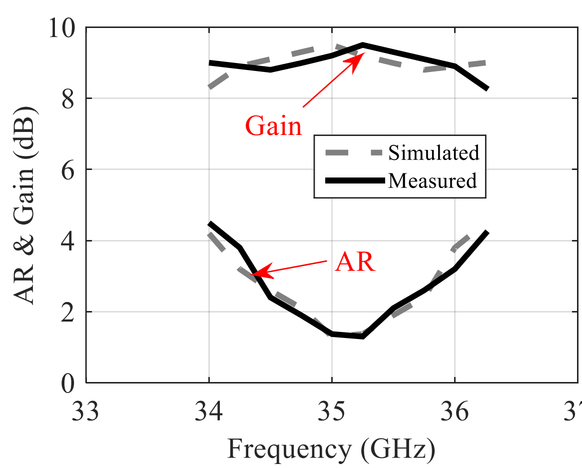


Fig. 8. The simulated and measured ARs and gains of proposed CP ZOR antenna.

V. CONCLUSION

In this work, a SIW mm-wave antenna with wide HPBW was presented. The combination between a SIW aperture-coupled antenna (with TM_{010}) and a mushroom polarizer metamaterial antenna helps develop HPBW. In order to increase the antenna performance and generate two orthogonal modes simultaneously to attain CP, the study applied a modified mushroom with etching corner on SIW slot. The antenna has the advantages such as a simple structure, and planar configuration in spite of using two radiators. In conclusion, in the present study, an antenna with 7.3% BW, 4.7% 3-dB axial ratio BW, the maximum gain of 8.93 dBic, and 103.8° HPBW was designed.

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