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Bandwidth enhancement using modified L-probe fed slotted patch antenna for WLAN and UMTS applications

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Abstract

In this paper, two different radiating structures fed with modified L-probe, are reported using a circuit theory concept. The proposed antennas are operating in wireless local area network (WLAN) and universal mobile telecommunications system (UMTS) frequency bands. In the first design, an E-shaped patch is studied to increase the bandwidth. It is observed that the bandwidth is directly proportional to notch dimensions. In the second design, E-shaped patch is modified to reduce the antenna size up to 30% with high bandwidth. In the first design, measured bandwidth and gain achieved are 32.68% (1.92–2.67 GHz) and 8.43 dBi while in second design it is 34.19% (1.94–2.74 GHz) and 8.39 dBi, respectively. Radiation patterns for both the antennas are symmetrical and broadside in nature. The proposed antennas are fabricated and measured results compare well with the theoretical and simulated results.

Introduction

Many attempts have been made to overcome the low gain and narrow bandwidth of the microstrip antenna [1]. In this process, several measures were taken such as modifying the shape of radiating elements [2] and modifying the ground plane structure [3], use of low dielectric constant material, stacked patch antenna, use of impedance matching networks etc. [4–6] and creating defect in the ground plane [7]. Different feeding techniques such as L-strip proximity feeding, T-shape probe feeding etc. are also used to increase the bandwidth [8, 9]. In the literature some theoretical analysis of strip line fed, L-strip/L-probe fed patch antennas have also been reported for the wideband operation [10, 11]. When the strip line feeding is modified as L-strip/L-probe it provides better gain and directivity suitable for wideband operation and antenna array design. In such feeding an additional capacitance is created between the radiating patch and the feed which cancels the inductance of the vertical portion of the L-probe. Consequently, this effect broadens the impedance bandwidth of the antenna [12].

In this paper, symmetrical slots are etched in the radiating patch and modified L-probe feeding is used to increase the antenna bandwidth. The present design gives better impedance bandwidth and gain as compared with the previously reported results [13–16]. The antenna parameters are calculated using high frequency structure simulator (HFSS) and verified with IE3D simulators. The proposed designs are also analyzed using circuit theory concept. The theoretical and simulated results are validated with measured results.

Antenna geometry and theoretical analysis

In conventional design of L-probe feeding, probe is required to bend which makes the design process complicated. But, in the present paper a narrow horizontal conducting strip is printed on the upper side of the substrate of thickness t and the radiating patch on the lower side. In order to avoid the bending procedure, a vertical thick probe is soldered to the horizontal strip. A small portion of patch is removed so that unnecessary drilling of the patch can be avoided. The present structure looks like a capacitively loaded L-probe fed patch antenna. Figure 1 shows the top view and side view of the proposed antennas. In antenna-I, two symmetrical notches of dimensions $l_1 \times w_1$ are incorporated along one of the radiating side of the patch resulting an E-shaped structure. In antenna-II, two additional notches of dimensions $l_2 \times w_2$ and a center notch of dimensions $l_3 \times w_3$ are introduced in the opposite radiating edge. For the wideband operation air substrate of thickness H is used between ground plane and the patch. The design specification of the proposed antenna is presented in Table 1.

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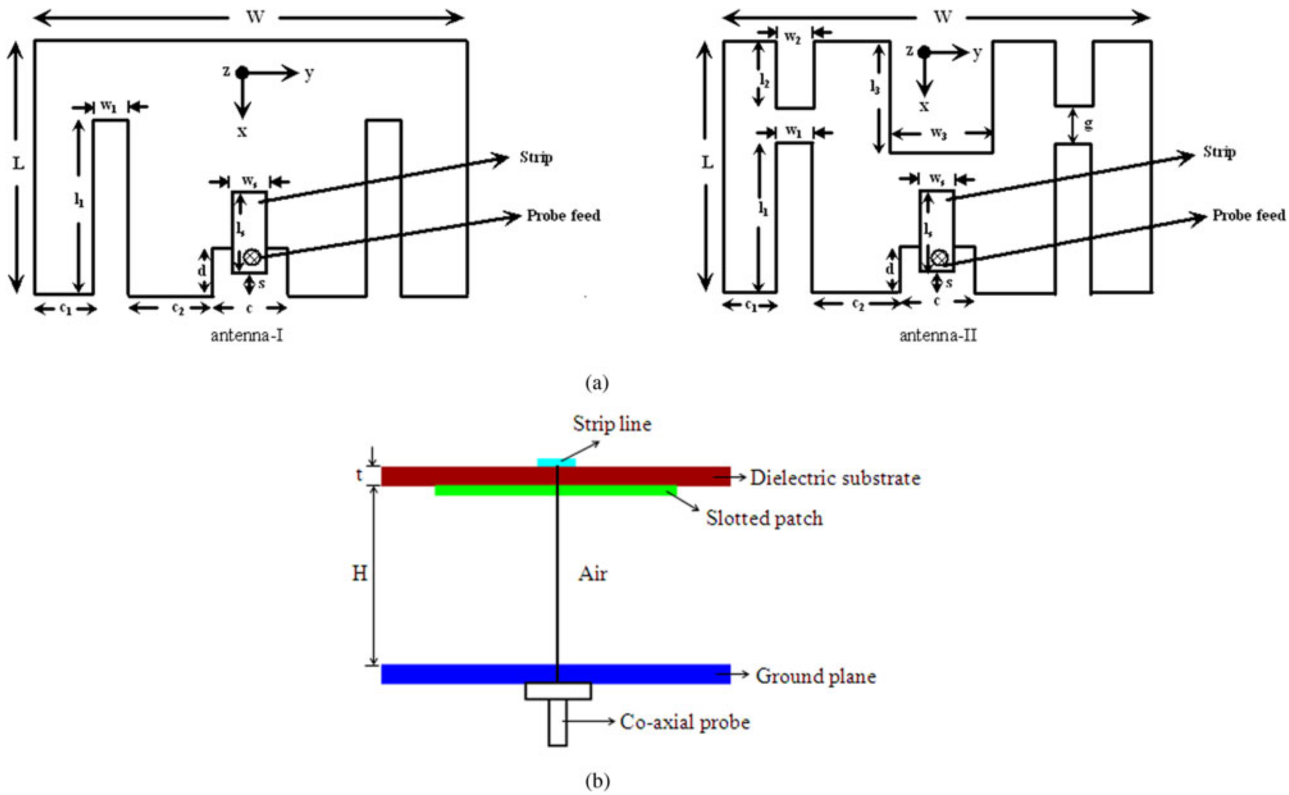

Fig. 1. Geometry of antenna (a) Top view (b) Side view for antenna-I and antenna-II.

Table 1. The design specification of the proposed antennas.

Parameter	Values
$W \times L$	72.0×50.0 mm^2
Air gap (H)	10.0 mm
Thickness of dielectric substrate (t)	1.59 mm
Substrate used	Rogers RT/duriod 5880 ($\varepsilon_r = 2.2$)
$l_1 \times w_1$	26.0×3.0 mm^2
$l_2 \times w_2$	20.0×3.0 mm^2
$l_3 \times w_3$	30.0×20.0 mm^2
Conducting strip ($w_s \times l_s$)	4.0×12.5 mm^2
$c \times d$	12.0×7.0 mm^2
c_1	11.0 mm
c_2	16.0 mm
g	4.0 mm
s	3.0 mm

Equivalent circuit for the modified L-probe

The modified L-probe can be considered as two sections, one is vertical probe and another is horizontal strip. Lumped elements of each section contribute in the total impedance of the antennas.

Vertical probe: From Fig. 2(a), the vertical probe can be analyzed as a series combination of resistance R_v and inductance L_v and can be given as [17]

$$R_v = \frac{\sqrt{\pi f \mu / \sigma}}{d_p}(H+t), \tag{1}$$

where μ = permeability of the probe conductor, f = frequency in GHz, d_p = diameter of probe and

$$L_v = 2.032(h+t)\left[\ln\frac{(H+t)}{d_p} + 0.2235\frac{d_p}{(H+t)} + 1.193\right]nH. \tag{2}$$

Horizontal strip: A series combination of distributive resistance R_s and inductance L_s is developed due to horizontal conducting strip and can be given as [18].

$$L_s = 0.2(H+t)\left[\ln\left\{\frac{2(H+t)}{w_s+t_s}\right\} + 0.2235\left\{\frac{w_s+t_s}{(H+t)}\right\} + 0.5\right](nH), \tag{3}$$

$$R_s = 4.13 \times 10^{-3}(H+t)\frac{\sqrt{f\rho_0}}{(w_s+t_s)} \text{ with } f \text{ in GHz}. \tag{4}$$

Here, w_s = width of the strip, t_s = thickness of the strip, ρ_0 = ratio of the specific resistance of strip and copper.

The distributive capacitance C_{sp} between the horizontal strip and the radiating patch can be given as:

$$C_{sp} = \frac{\varepsilon_r \varepsilon_0 l_s w_s}{t}. \tag{5}$$

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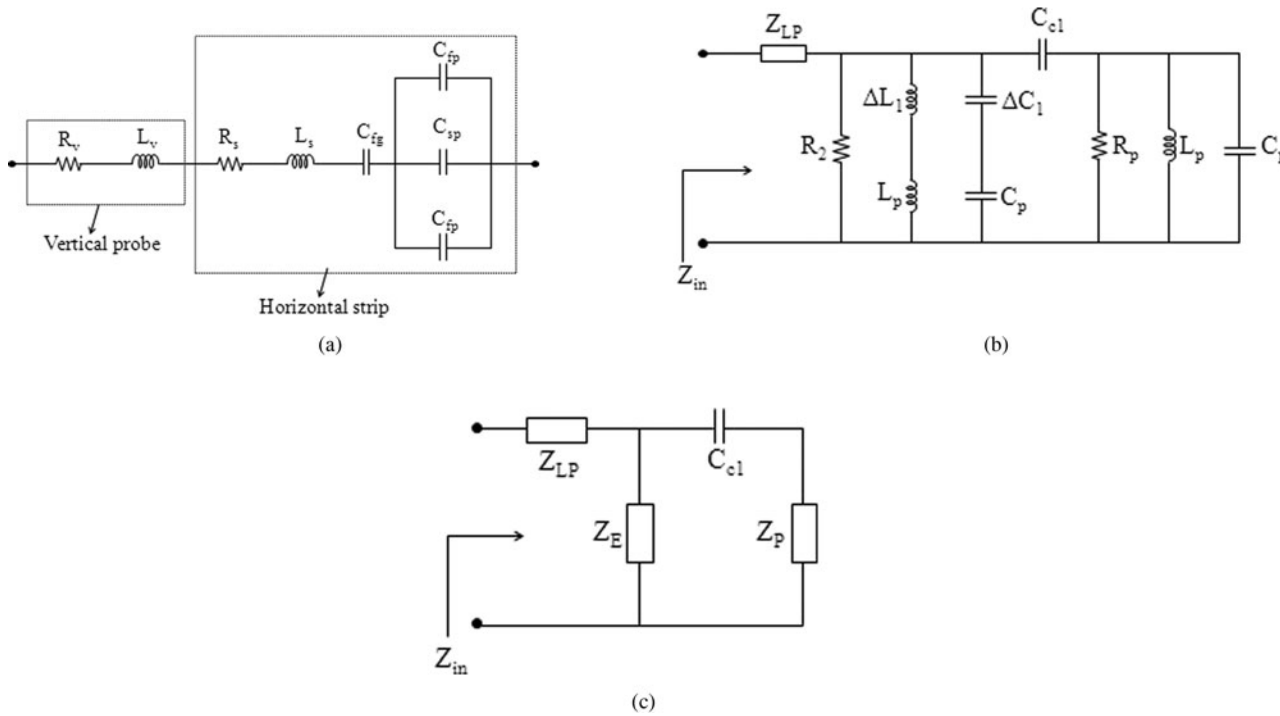

Fig. 2. (a) Equivalent circuit of modified L-probe (b) equivalent circuit of antenna-I and (c) modified circuit of (b).

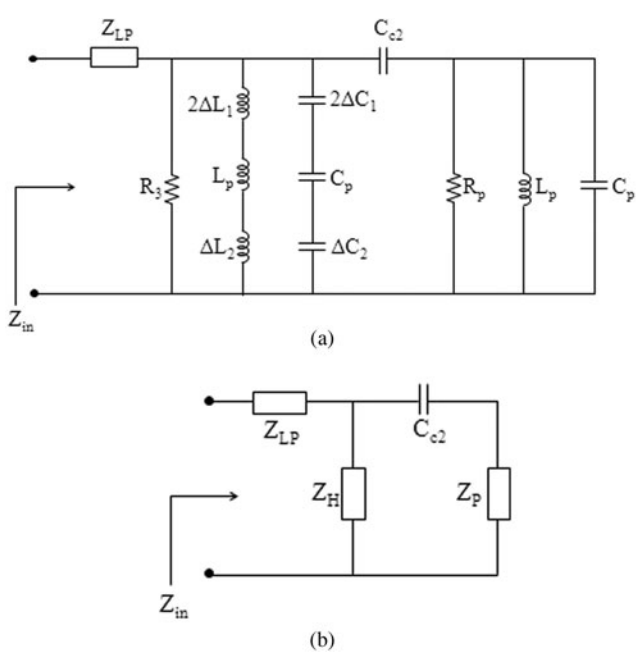

Fig. 3. (a) Equivalent circuit of antenna-II and (b) modified circuit of antenna-II.

Since the open ends of the horizontal strip above the radiating patch will have a fringing field, the effective length of the strip is increased. The increment of length will cause some extra capacitance which is fringing capacitance and it can be calculated as:

$$C_f = \frac{l_e \varepsilon_e^{-1/2}}{cZ_0} \quad (6)$$

here,

$$l_e = \frac{0.412\text{h}(\varepsilon_e + 0.3)(w_s/h + 0.264)}{(\varepsilon_e - 0.258)(w_s/h + 0.8)},$$

in which, ε_e is the effective dielectric constant for the material under conducting strip [17]. The fringing capacitance C_{fp} between the open end of the strip and the patch is calculated by putting the substrate height $h = t$ and the fringing capacitance C_{fg} can be given by putting $h = (H + t)$. The entire feeding acts as a series L–C resonant element and connected in series with the radiating patch.

Equivalent circuit for antenna-I and antenna-II

The antenna-I is realized by introducing two parallel slits in the radiating edge and symmetrical with respect to the horizontal strip. This radiating structure is considered equivalent to two resonators, one as an original patch and another as a modified patch with additional series inductor ΔL_1 and capacitance ΔC_1 with the initial inductance L_p and capacitance C_p of the origin patch. The value of ΔL_1 and ΔC_1 can be given by [19, 20]. Now, these two resonators are coupled through the coupling capacitance C_{c1} (Fig. 2b) and its value can be given as:

$$C_{c1} = \frac{-(C_p + C') + \sqrt{((C_p + C')^2 - 4C_pC'(1 - (1/C_m^2)))}}{2}, \quad (7)$$

here,

$$C' = \frac{\Delta C_1 C_p}{\Delta C_1 + C_p},$$

$$C_m = \frac{1}{\sqrt{Q_1 Q_2}},$$

where Q_1 and Q_2 are the quality factors of the two resonators.

The microstrip patch is considered as a parallel combination of resistance (R_p), inductance (L_p) and capacitance (C_p) and the corresponding values are given as [18].

$$C_p = \frac{\varepsilon_0 \varepsilon_e \mathrm{LW}}{2\mathrm{H}} \cos^{-2}\left(\frac{L_s \pi}{L}\right), \tag{8}$$

$$L_p = \frac{1}{\omega^2 C_p}, \tag{9}$$

$$R_p = \frac{Q_r}{\omega C_p}. \tag{10}$$

In the proposed geometry, the effective dielectric constant for the fed patch is calculated by [21]:

$$\varepsilon_e = \frac{\varepsilon_{rs}+1}{2} + \frac{\varepsilon_{rs}-1}{2}\left(1+\frac{12H}{W}\right)^{-1/2}, \tag{11}$$

where, "H" is the thickness between ground and slotted patch, and

$$\varepsilon_{rs} = \frac{\sum_{i=1}^{n} h_i}{\sum_{i=1}^{n} \frac{h_i}{\varepsilon_{ri}}},$$

where n is the number of stacked layers.

The equivalent circuit of the modified L-probe fed antenna-I is shown in Fig. 2(b). The input impedance for this antenna can be calculated using Fig. 2(c).

$$Z_{in} = Z_{LP} + Z'', \tag{12}$$

where,

$$Z_{LP} = R_v + j\omega L_v + R_s + j\omega L_s + \frac{1}{j\omega C_{eq}}, \tag{13}$$

in which,

$$C_{eq} = \frac{(C_{fg})(2C_{fp}+C_{sp})}{(C_{fg}+2C_{fp}+C_{sp})},$$

and

$$Z'' = \frac{Z_E Z'}{Z_E + Z'},$$

where,

$$Z' = \frac{1}{j\omega C_{c1}} + Z_p,$$

$$Z_E = \frac{1}{(1/R_2) + (1/j\omega(L_p+\Delta L_1)) + j\omega C_{eq1}},$$

$$C_{eq1} = \frac{C_p \Delta C_1}{C_p + \Delta C_1},$$

$$Z_p = \frac{1}{(1/R_p) + (1/j\omega L_p) + j\omega C_p}.$$

The circuit diagram of antenna-II is shown in Fig. 3. For antenna-II, the input impedance Z_{in} is calculated as:

$$Z_{in} = Z_{LP} + Z_1'', \tag{14}$$

where,

$$Z_1'' = \frac{Z_H Z_1'}{Z_H + Z_1'},$$

$$Z_1' = \frac{1}{j\omega C_{c2}} + Z_p,$$

and

$$Z_H = \frac{1}{(1/R_3) + (1/j\omega(L_p + 2\Delta L_1 + \Delta L_2)) + j\omega C_{eq2}},$$

in which,

$$C_{eq2} = \frac{\Delta C_1 C_2 C_p}{C_2 C_p + \Delta 2 C_1 C_2 + 2 C_1 C_p},$$

here, C_{c2} can be calculated using equation 7 and Fig. 3(a).

Now, the return loss and voltage standing wave ratio (VSWR) can be obtained for both the antennas as:

$$\text{Reflection coefficient } (\Gamma) = \left|\frac{Z_{in}-Z_0}{Z_{in}+Z_0}\right|, \tag{15}$$

$$\mathrm{VSWR} = \frac{1+|\Gamma|}{1-|\Gamma|}, \tag{16}$$

$$\text{Return loss (dB)} = 20\log\left|\frac{Z_{in}-Z_0}{Z_{in}+Z_0}\right|. \tag{17}$$

Antenna fabrications

The radiating patch is printed on substrate RT Duroid 5880 with the thickness 1.59 mm. The patch is separated by air gap spacing of 10 mm from the ground plane. The fabricated antenna is shown in Fig. 4. The patch is electromagnetically coupled by a conducting strip connected with the probe. This conducting strip of dimensions $l_s \times w_s$ is printed above the radiating patch and both are separated by a dielectric substrate of thickness t = 1.59 mm. Such feeding method is more convenient than the conventional L-strip fed patch. The perturbed structure introduces some additional current to flow in the patch which causes the

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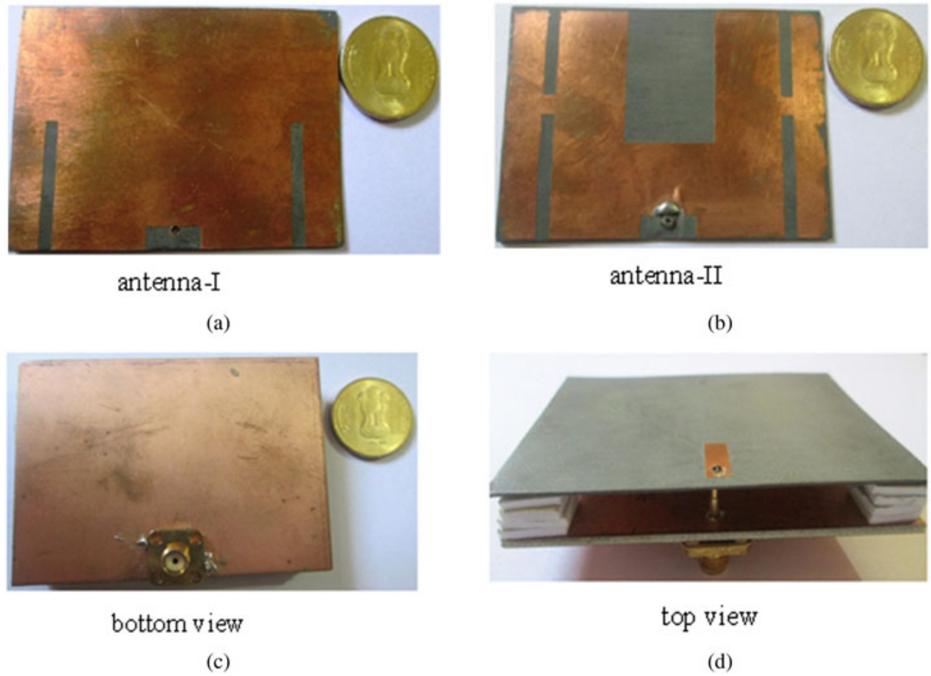

Fig. 4. Prototype of the fabricated antenna (a) antenna-I (b) antenna-II (c) bottom view (d) top view.

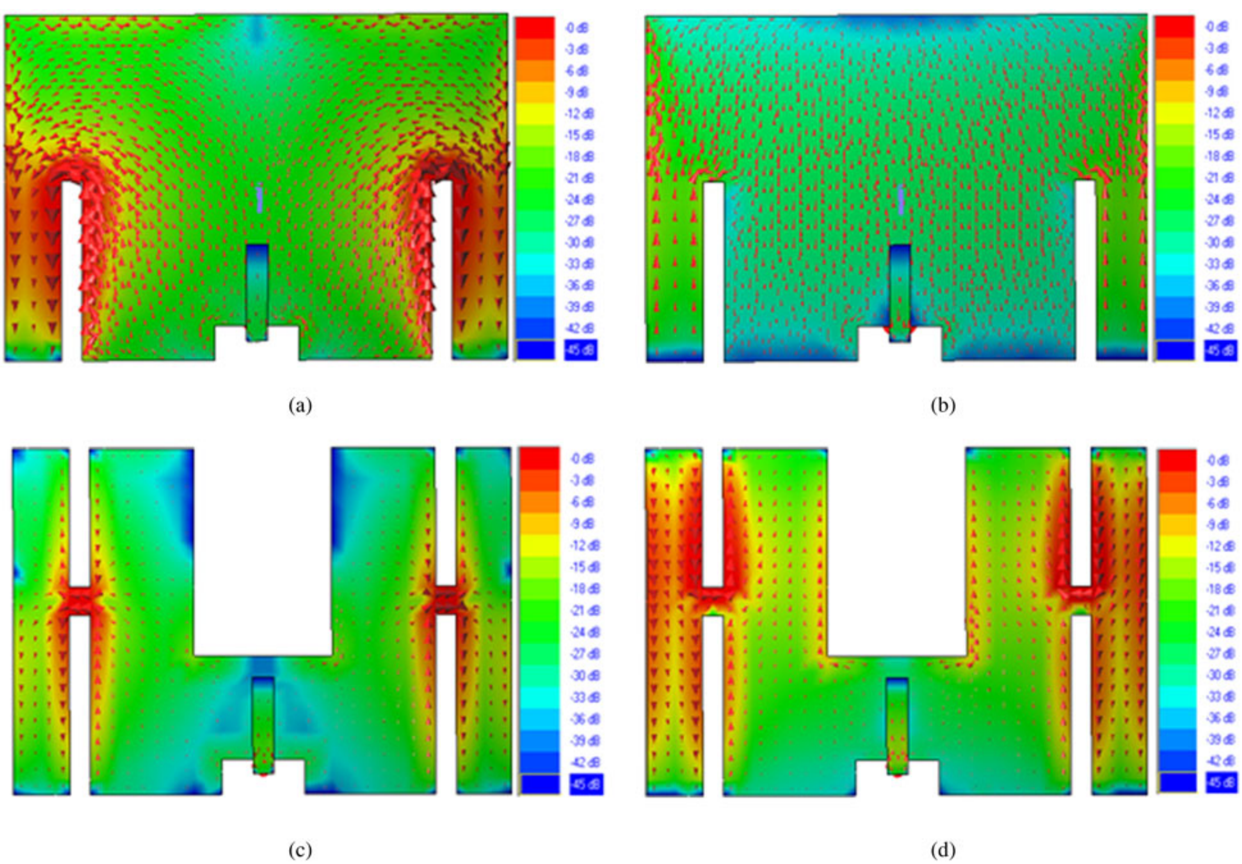

Fig. 5. Current distribution using IE3D at two resonant frequencies 2.1 and 2.5 GHz for antenna-I (a) and (b), and for antenna-II (c) and (d).

higher modes to shift closer and hence a wideband characteristics of the antenna is observed. Therefore, the current distributions on the two antennas are shown in Fig. 5 at two resonant frequencies 2.1 and 2.5 GHz, respectively. From this figure, it is clear that the current length can be optimized by introducing some notches along the radiating edges. The dimensions of the notches are taken in such a way so that the wideband nature of the antenna can be realized. The current distribution curves indicate that the

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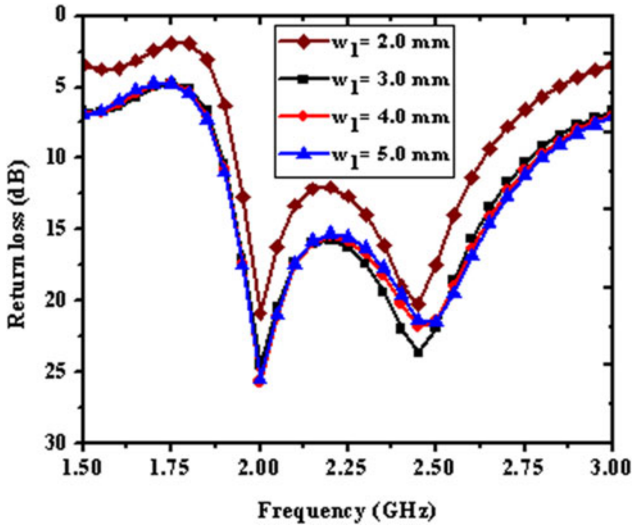

Fig. 6. Variation of return loss with frequency for different value of w_1 ($l_1 = 26.0$ mm is fixed) for antenna-I.

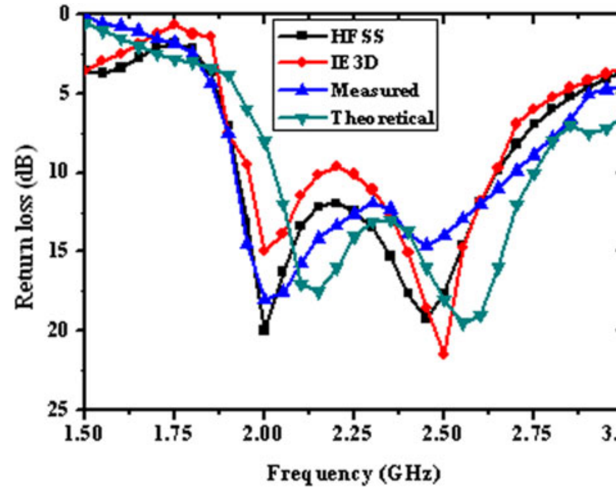

Fig. 8. Comparative graph of return loss with frequency for antenna-I, ($w_1 = 3.0$ mm and $l_1 = 26.0$ mm).

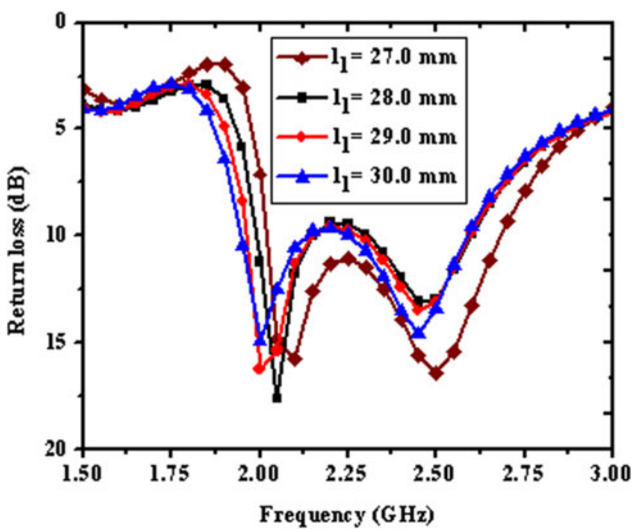

Fig. 7. Variation of return loss with frequency for different values of l_1 ($w_1 = 3.0$ mm is fixed) for antenna-I.

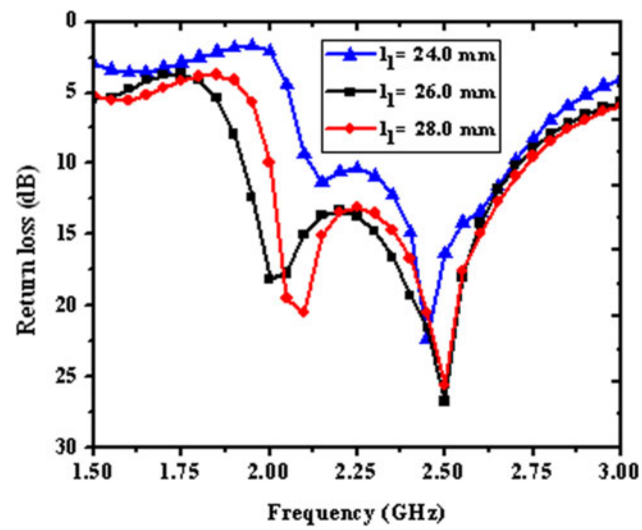

Fig. 9. Variation of return loss with frequency for different values of l_1 ($l_2 = 20.0$ mm, $w_1 = w_2 = 3.0$ mm) for antenna-II.

lower resonance is produced by two parallel notches in the patch while the higher one is created due to the combined effect of the initial patch and the notches.

Results and discussion

It is noted that all the antenna parameters are calculated using HFSS simulator for antenna-I and antenna-II. For antenna-I, the variation of return loss with frequency for different values of notch width (w_1) is shown in Fig. 6. From the figure, it is observed that the bandwidth increases as w_1 increases. It is found that as notch length l_1 increases (for a given value of $w_1 = 3.0$ mm) the operational frequency band shifts towards the lower side with a slight increase in bandwidth as shown in Fig. 7. The operational frequency band for antenna-I, calculated using HFSS, extends from 1.91–2.66 GHz and bandwidth obtained is 32.82%. The fabricated antenna is measured using Vector Network Analyzer with model type-Agilent N5230A. The measured bandwidth for antenna-I is found to vary from 1.92–2.67 GHz (32.68%). Theoretical and simulated return loss results are compared with the measured results which are in good agreement (Fig. 8). Return loss is also plotted for antenna-II in Fig. 9 and it is observed that the bandwidth significantly increases with the increasing value of l_1. Similar results are also found in case of increasing value of l_2 (Fig. 10).

From Fig. 11, it is found that the bandwidth marginally increases as the length of center notch l_3 increases, however; the patch size is reduced upto 30% as compared with the conventional rectangular patch. Thus, for a particular dimension, one can achieve a miniaturized antenna with appreciable bandwidth. The results for antenna-II obtained from HFSS, IE3D, and proposed theory are in agreement with the measured results (Fig. 12). The operational frequency band for this antenna, calculated using HFSS, is varying from 1.93 to 2.83 GHz and the bandwidth obtained is 37.82%. The measured frequency band for antenna-II is found to vary from 1.94 to 2.74 GHz (34.19%).

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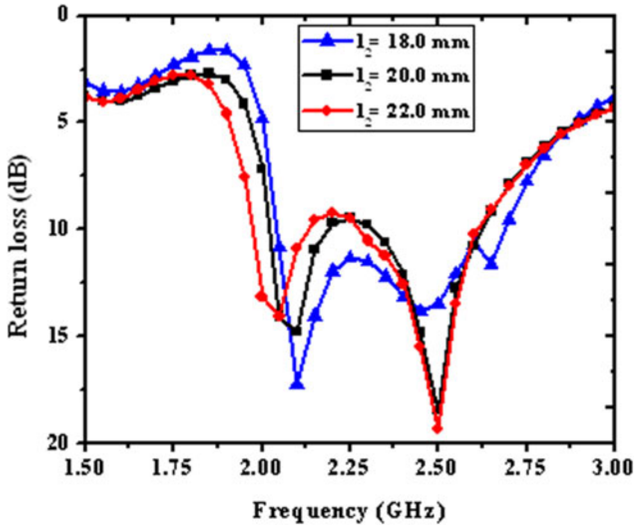

Fig. 10. Variation of return loss with frequency for different values of l_2 (l_1 = 26.0 mm, w_1 = w_2 = 3.0 mm) for antenna-II.

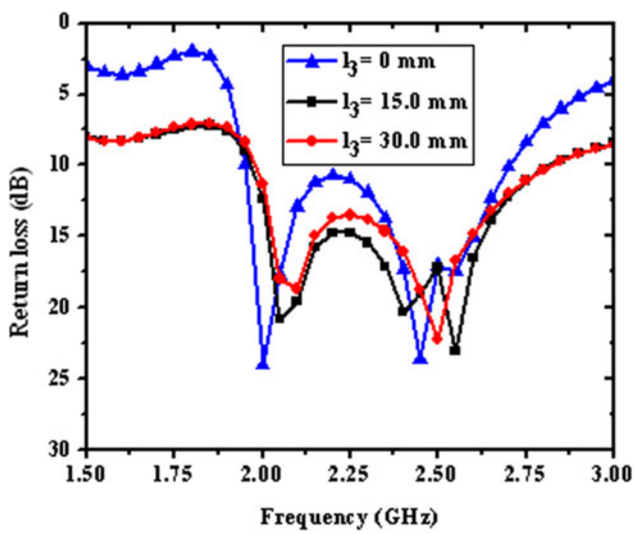

Fig. 11. Variation of return loss with frequency for different values of l_3 (w_3 = 20.0 mm is fixed) for antenna-II.

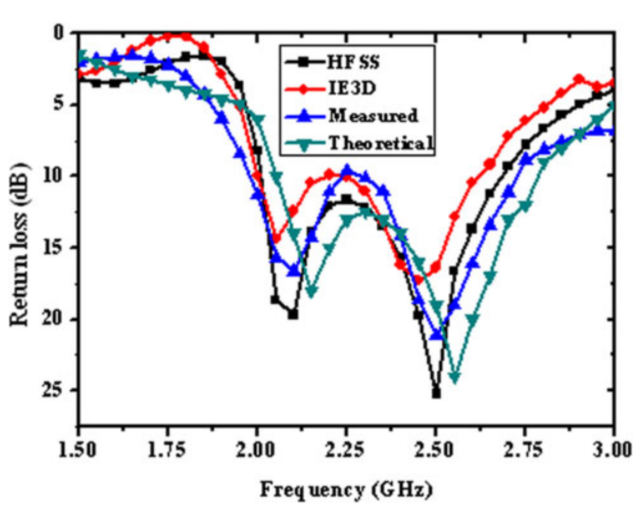

Fig. 12. Comparative plot of return loss with frequency for antenna-II (l_3 = 30 mm, w_3 = 20.0 mm, l_1 = 26 mm, w_1 = w_2 = 3 mm).

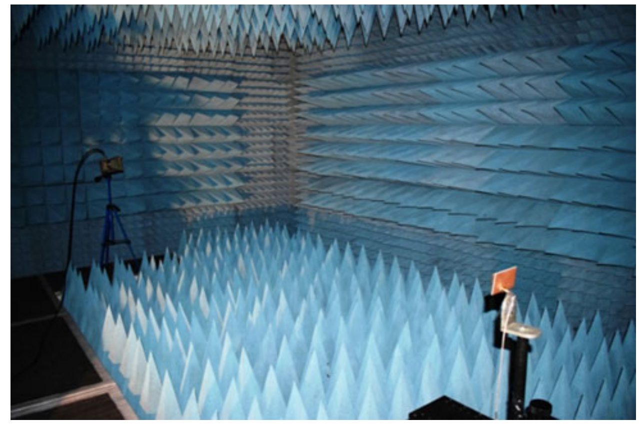

Fig. 13. Gain measurement setup.

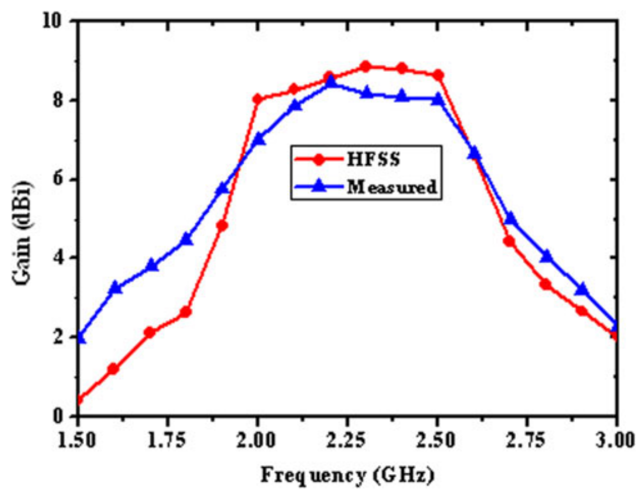

Fig. 14. Measured gain for antenna-I.

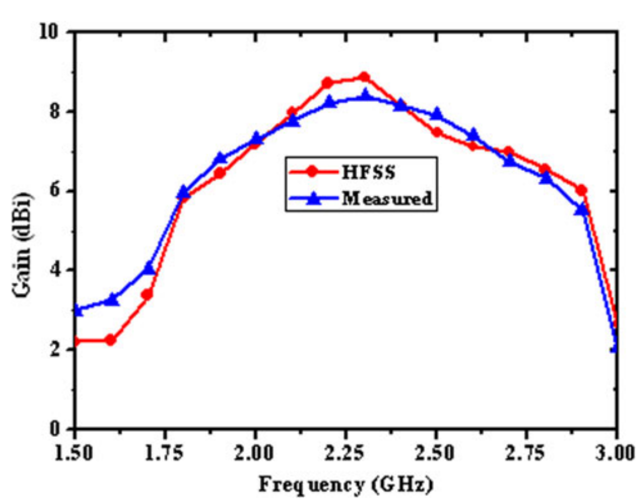

Fig. 15. Measured gain for antenna-II.

Table 2. Comparison of the proposed antennas with earlier reported antennas.

Ref. No.	Patch dimensions (mm^2)	Bandwidth (GHz)	Bandwidth (%)	Antenna Gain (dBi)
[13]	72 × 50	1.92–2.51	26.50	7.8
[14]	120 × 60	1.17~0.89	26.70	7.0
[15]	76.4 × 76.4	2.02–2.88	36.00	2.0
[16]	56.7 × 56.7 (Parasitic), 48.6 × 48.6 (Radiating)	1.66–2.27	30.00	7.5
Proposed (Antenna-I)	72 × 50	1.92–2.67	32.68	8.43
Proposed (Antenna-II)	72 × 50	1.94–2.74	34.19	8.39

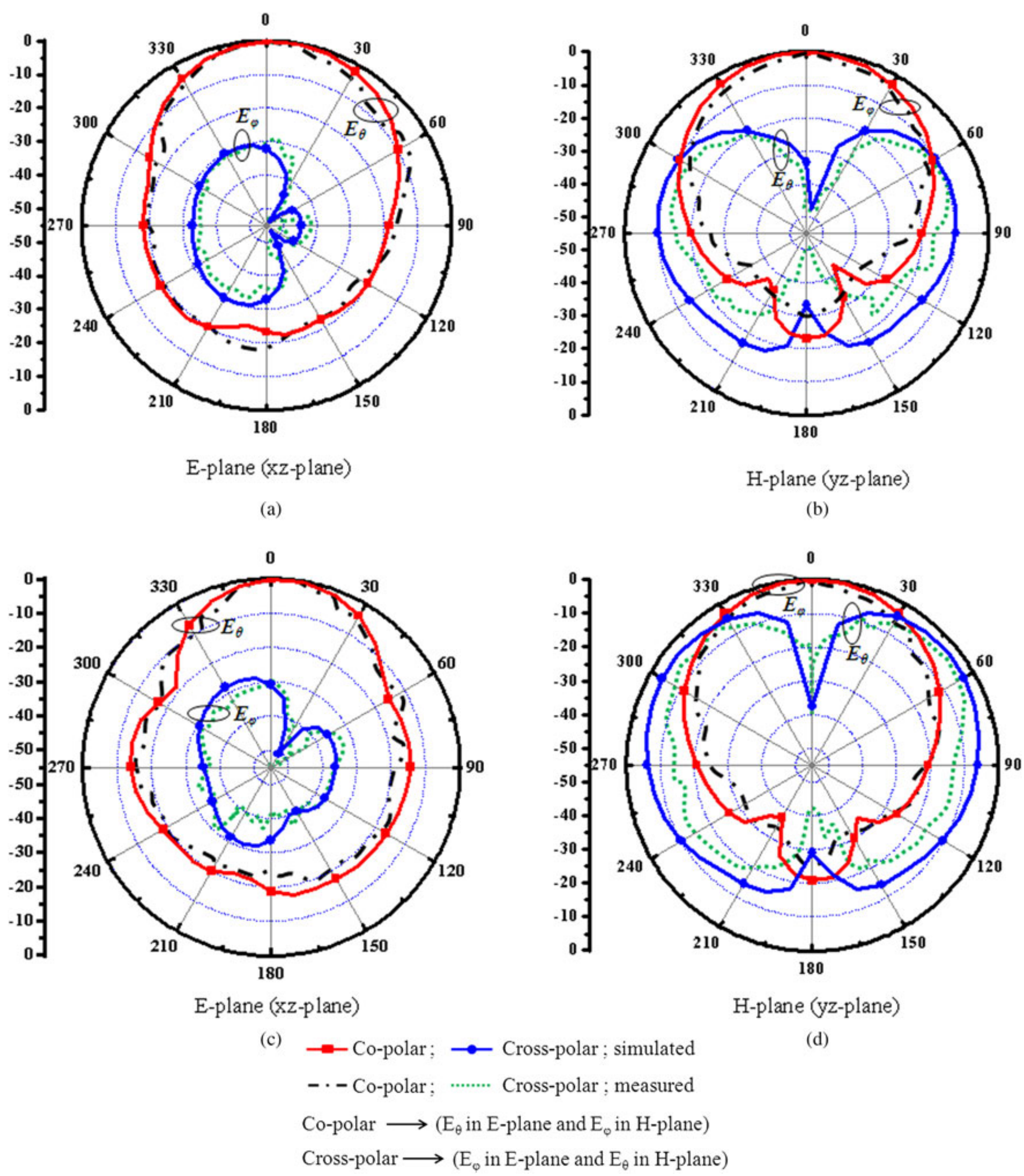

Fig. 16. Simulated (HFSS) and measured radiation pattern of antenna–I at f = 2.1 GHz (a and b) and at f = 2.5 GHz (c and d).

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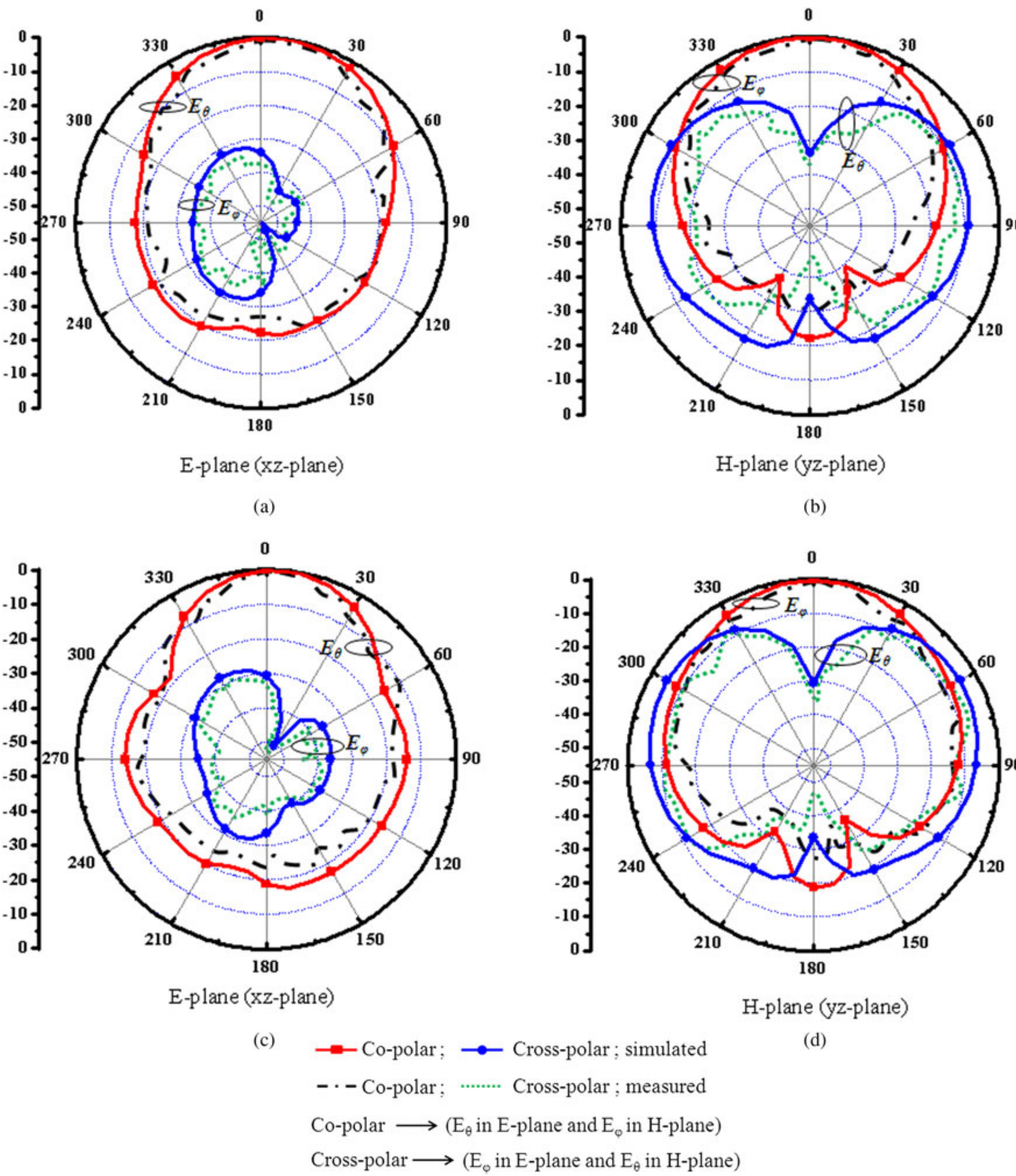

Fig. 17. Simulated (HFSS) and measured radiation pattern of antenna–II at $f = 2.1$ GHz (a and b) and at $f = 2.5$ GHz (c and d).

Some deviation in the simulated and measured results is due to fabrication imperfection and the misalignment of SMA connector. The gain measurement is performed inside an in-house anechoic chamber of physical dimensions $7.0 \times 5.0 \times 3.0\ m^3$ and shown in Fig. 13. A Dual Ridge Horn Antenna (DRH20) of size $56 \times 41 \times 18\ cm^3$ is used for the gain measurement. The gains of antenna-I and antenna-II are measured from 1.5 to 3.0 GHz with an equal frequency interval of 0.1 GHz. Measured gains for both the antennas are shown in Figs 14 and 15, respectively. The discrepancy between simulated and measured gain of antennas may be due to the fact that no near-field to far-field correction is applied for the direct measurement. Another reason is due to the use of DRH horn antenna which is somewhat not suitable for such type of correction. It is observed that the gain is always above 6.5 dBi for the entire band of operation. The maximum measured gain for antenna-I and antenna-II are 8.43 and 8.39 dBi, respectively. The measured average gain for antenna-I and antenna-II are calculated and found to be 7.23 and 7.63 dBi, respectively. The comparisons of the proposed antennas performance are illustrated in Table 2. The proposed antennas exhibit sufficient bandwidth and high gain as compared with the earlier reported antennas.

The radiation patterns measured for antenna-I and II are plotted in Figs 16 and 17 at $\varphi = 0^0$ (E plane) and $\varphi = 90^0$ (H plane) for two resonant frequencies (2.1 and 2.5 GHz), respectively. The cross-polarization level is quite low as compared with co-polarization level for both the antennas at $\varphi = 0°$ and $\varphi = 90°$. The radiation patterns of the proposed antennas are symmetrical, broadside in nature and linearly polarized. E and H plane half power beamwidths (HPBW) is measured for both the antennas

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at 2.1 and 2.5 GHz, respectively. For antenna-I, the *E* and *H* plane HPBW are 57.8° and 61° at 2.1 GHz and 45.48° and 58.68° at 2.5 GHz, respectively. For antenna-II, the *E* and *H* plane beamwidths are 58.14° and 63.05° at 2.1 GHz and 45.41° and 54.24° at 2.5 GHz, respectively. Also, it is noted that the simulated and measured values of beamwidths are in good agreement.

Conclusion

A modified L-strip feeding technique is used to study the slotted patch antennas. The circuit theory is successfully implemented to calculate the antenna parameters. From this study it is inferred that the antenna bandwidth is function of dimensions of the notches incorporated in the patch. Antenna size is also reduced along with the enhancement of bandwidth. Radiation pattern is stable and gain is acceptable across the entire operational band. The dimension of conducting strip may also be the key parameters for further improvement in the antenna performance, which is not studied in the present paper. The proposed antennas can be used for WiFi/Bluetooth, UMTS, radio frequency identification device (RFID) and other communication systems.

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