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DOI:
10.1049/iet-map.2017.0478

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Composite right/left-handed based compact and high gain leaky-wave antenna using complementary spiral resonator on HMSIW for Ku band applications

ISSN 1751-8644
doi: 0000000000
www.ietdl.org

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Abstract: In this communication, a novel compact high gain composite right/left-handed (CRLH) based leaky-wave antenna (LWA) is presented at Ku band. Half-mode substrate integrated waveguide (HMSIW) incorporating with suitably oriented Complementary Quad Spiral Resonator (CQSR) is used to achieve a CRLH LWA. The uni cell is realized by a CQSR in such a way that orientation of spirals exhibit higher leakage loss having minimum cross coupling between them. The antenna is capable to scan backward to forward along with broadside direction in visible space. The proposed configuration is just length of 4.85λ_0 which can scan within the frequency range of 13.5-17.8 GHz having beam scanning range of 86° (-66° to 20°) and maximum gain of 16 dBi. The simulated reflection coefficient of the proposed antenna is below -10 dB throughout the working frequency range with a side-lobe level of below -10 dB. **The designed prototype is much compact in nature having high gain, fair scanning range, good cross-polarization level along with simpler design methodology and tuning capability to enhance the gain as well as radiation efficiency maintaining fixed size.** The proposed antenna could be a promising candidate in Ku-band applications like Fixed Satellite Services (FSS) and Broadcast Satellite Services (BSS) etc.

1 Introduction

Leaky-wave antennas (LWAs) are affiliated to the travelling-wave antenna family based on transmission line with periodic radiating elements [1]. For the past few decades, numerous LWAs have been reported with the development of various microwave transmission lines (TLs) including rectangular and circular waveguides [2], [3], various planar waveguides like parallel plate waveguides [4], microstrip lines [5], [6], co-planar waveguides [7], dielectric slabs [8] etc.. However, at high frequency it is not feasible to use microstrip based LWAs because of high conductor loss and low efficiency. Though the waveguide based LWAs can be used at higher frequencies but the integration with other planar structure is quite difficult. On the other hand, substrate integrated waveguide (SIW) is a good compromise between dielectric filled waveguide and microstrip and becomes popular in the recent years for its significant advantages such as low profile, low cost, light weight, easy integration [9]. Along with the above advantages, due to their high efficiency, high gain, narrow elevation beamwidth make SIW an attractive candidate for LWA and opens a new avenue for frequency beam scanning application. Several SIW based LWAs have been reported in [10–13]. Over the last few years, the use of half-mode substrate integrated waveguide (HMSIW) in designing leaky-wave antennas has also increased extensively. It avails all the advantageous features of SIW structures along with the size reduction by the factor of half [14], [15]. But these suffer the problem of open stopband (OSB) where broadside radiation is a null. Therefore, backward-broadside-forward frequency beam scanning is not achieved. To mitigate this problem, several techniques are developed and employed in designing LWA with full-space scanning [16–18]. In [19], a quarter-wave transformer, or alternatively a matching stub is proposed to eliminate the stopband in designing 1-D conventional periodic microstrip LWA at K-band but it suffers narrow bandwidth as well as smaller beam scanning range and this concept is validated experimentaly in [20]. Some other techniques are also proposed to

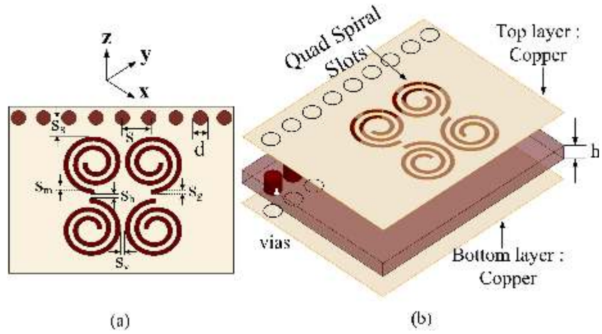

Fig. 1: (a) Layout of the proposed CQSR unit cell in HMSIW. The parameter values are: S=1.6mm, d=0.8mm, S_m=0.22mm, S_h=0.24mm, S_v=0.24mm, S_g=0.18mm, S_s=1.23mm. (b) 3-D view of the proposed CQSR unit cell structure.

eliminate the stopband, like unit cells are placed in transversal asymmetry [21], lattice-network based TL model [22], using π-matching network [23] in designing periodic leaky-wave antennas. In [24], a self-matched periodic LWA is designed in microstrip environment using U-stub and an interdigital capacitor which is compact in size with suitable scanning range but has maximum gain of 10.6 dBi. Recently, several CRLH based antennas have been reported using microstrip, SIW, CRLH rectangular waveguide etc. They own some unique features not available for conventional microwave structures such as it supports both backward and forward waves and this is applicable for LWAs for achieving continuous beam steering in visible space. CRLH based LWAs with detailed dispersion analysis in microstrip environment is reported in [25]. A K-Band frequency-scanned LWA based on CRLH TLs is reported in [26]. In [27], the CRLH SIW based LWA with low cross polarization level is

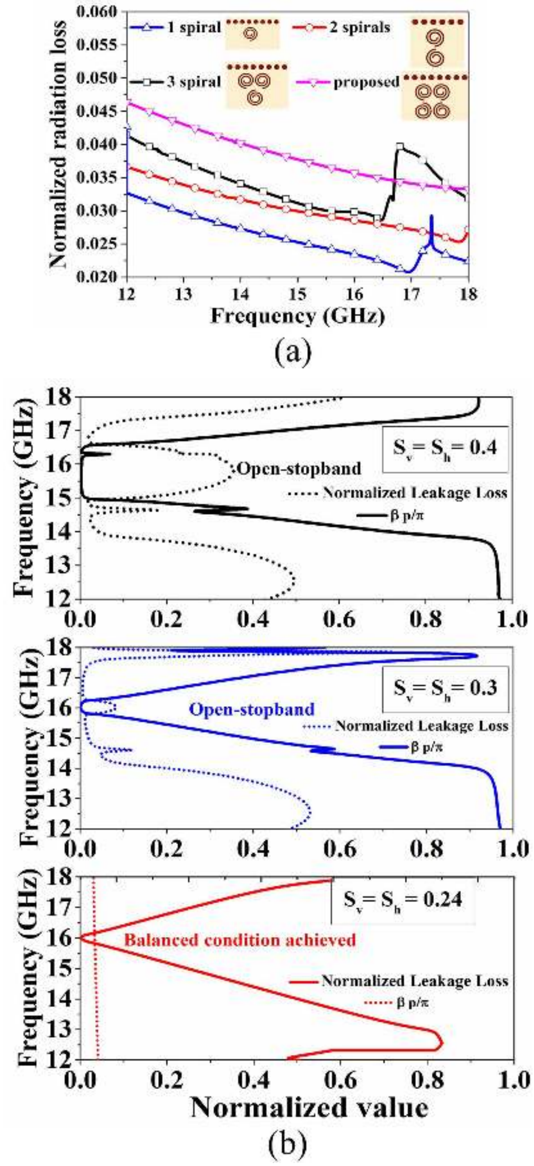

Fig. 2: **Variation of normalized leakage loss for changing (a) the number of spirals, (b) gap between the spirals with open-stopband elimination.**

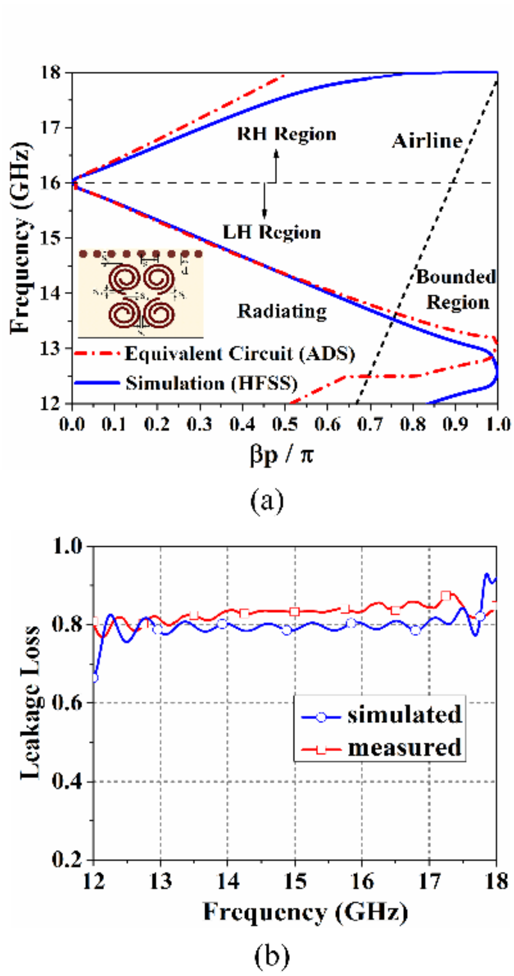

Fig. 3: (a) Comparison between simulated and optimized dispersion diagram of balanced CQSR unit cell of Fig. 1; (b) simulated and measured leakage loss of the unit cell.

shown. The CRLH SIW based LWA with polarization flexibility is shown in [28]. **It is to be noted that, the side-wall vias of HMSIW structures provides an effective shunt inductance and hence, to implement the CRLH medium, only series capacitance is required. By utilizing these benifits of HMSIW, several CRLH HMSIW based LWA have been reported in past [14], [29], [30], [31] where full space scanning is achieved by solving the open stopband issues for unbalanced transmission line. However, most of the previously reported antennas are not exhibiting a high gain with significantly decreasing the overall profile of the structure**.

Our main aim in this work is to propose a compact low profile leaky-wave antenna having higher gain and improved radiation efficiency which takes advantage of all the characteristics of HMSIW structures. This is facilitated by introducing the novel concept of complementary spiral resonators which help in obtaining better control of design parameters thereby providing easier tuning. The proposed design is also compared with other reported works. To the best of author's knowledge, such type of CRLH based leaky-wave antenna with compact size, high gain, improved side lobe level and more flexible in tuning has not been earlier proposed in literature. The proposed antenna is simulated as well as optimized in Ansoft High Frequency Structure Simulator (HFSS) and the same is fabricated and tested. Along with the good frequency scanning range and high gain of the proposed prototype, the antenna is advantageous in terms of compactness, simple single layered fabrication process that can be easily implemented on printed circuit board.

2 Unit Cell Design and Analysis

2.1 *Design of Complementary Quad Spiral Resonator (CQSR) Unit Cell*

The top and perspective views of the proposed CQSR unit cell in HMSIW are shown in Fig. 1. The proposed structure is designed on RT/Duroid 5880 substrate with dielectric constant (ϵ_r) = 2.2, loss tangent (tanδ) = 0.0009 and height (h) of 0.787 mm. Shunt capacitance and series inductance are distributed along the geometry, which correspond to an equivalent model of ground plane and top wall of the HMSIW structure respectively. A group of four complementary spiral slots are etched on the top wall of the HMSIW which provide series capacitance. Whereas, metallic vias are realized as distributed shunt inductance. Thus, combined effects support the property of left-handed region. The length of the unit cell is chosen by maintaining the homogeneity condition i.e. unit cell length $\ll \lambda_g/4$. By satisfying the balanced condition of the proposed unit cell, backward to forward frequency beam scanning together with broadside radiation is obtained. This implies the elimination of band-gap

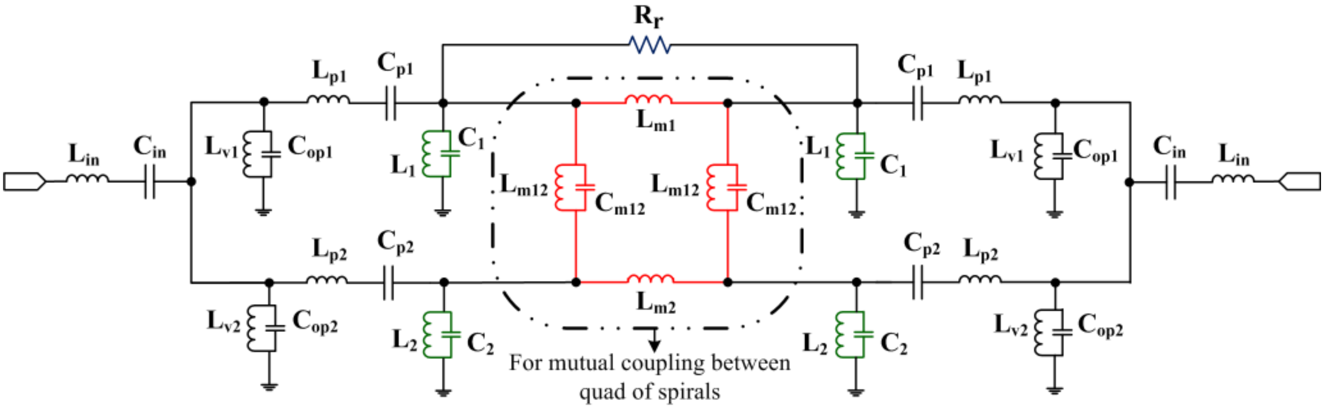

Fig. 4: Equivalent Circuit Model of HMSIW CQSR unit cell. The extracted circuit element values are L_{in}=0.367 nH, C_{in}=0.486 pF, L_{v1}=2.06 nH, C_{op1}=0.205 pF, L_{v2}=0.697 nH, C_{op2}=0.171 pF, L_{p1}=2.974 nH, C_{p1}=2.924 pF, L_{p2}=1.132 nH, C_{p2}=0.122 pF, L_1=0.482 nH, C_1=0.1 pF, L_2=0.887 nH, C_2=0.942 pF, L_{m1}=2.99 nH, L_{m2}=2.97 nH, L_{m12}=1.905 nH, C_{m12}=0.452 pF, R_r=1.416 ohm.

between left-handed and right-handed regions. The balanced condition for the unit cell is established by tuning the values of S_m, S_h, S_v, S_g, S_s at a fixed cut-off frequency and the left-handed region only gets affected by these parameters. The quad of spirals with suitable orientation is exploited as a unit cell of the proposed antenna in such a manner that **radiation loss in terms of normalized leakage constant is enhanced and showing a smooth variation within desired frequency band. The dependency on number of spirals on selection of unit cell is clearly depicted in Fig. 2(a) from where, it is clear that the proposed configuration with four spirals shows most suitable performance among the depicted four cases (in Fig. 2(a)). Though, similar leakage behavior is achieved from unit cell having two spirals but proposed unit cell is the better selection as side lobe level is also reduced by $\sim$ 5.41 dB compared to the unit cell containing two spirals. Moreover, the number of turns of spirals also helps in enhanced leakage loss resulting in improved efficiency and gain**. For better understanding, dispersion analysis is done and described next.

2.2 *Dispersion Analysis*

For better realization of the behavior of the leaky-wave antenna, dispersion characteristic of the CQSR unit cell is studied. It is done through driven-mode solution which is more time efficient than eigen-mode simulation. In the full wave simulation of unit cell, wave-port excitation is carried out. By considering the effect of periodicity, the unit cell is simulated in HFSS and the S-parameters are extracted to study the dispersion characteristics. During simulation, the mutual coupling effect between the unit cells is neglected. The dispersion equation for the periodic transmission line can be written as (1) [26]

$$cos(\beta p) = \frac{(A + D)}{2} \tag{1}$$

where A and D are the components of a transmission matrix of TL. To extract the dispersion curve from full wave simulation, the equivalent S-parameters of the periodic transmission line can be calculated as (2)

$$\beta p = cos^{-1}\left(\frac{1 - S_{11}S_{22} + S_{12}S_{21}}{2S_{21}}\right) \tag{2}$$

where p is the periodicity of unit cell, β is the propagation constant or radiating space harmonic. In the proposed design, n = 0 harmonic is responsible for radiation. For traditional transmission line, series inductors and shunt capacitors are the distributed elements which exhibit forward wave propagation. The left-handed transmission line is the dual of right-handed TL structure and it exhibits backward-wave propagation. It is observed from Fig. 3(a), at the transition frequency (16 GHz) of the right and left-handed regions, series (ω_{se}) and shunt (ω_{sh}) resonant frequencies are equal. Thus, balance condition is achieved. At this condition, phase constant β exhibits zero value at the transition frequency (ω_0) as mentioned in (3).

$$\omega_0 = \frac{1}{\sqrt[4]{L_R C_R L_L C_L}} = \omega_{se} = \omega_{sh} \tag{3}$$

Below this frequency, the LH region appears which supports the propagation of backward-wave and just above the transition frequency is RH region which supports forward-wave. The simulated and measured leakage loss [34] of unit cell is depicted in Fig. 3(b) where it is clear that the radiation loss has higher value due to the proposed CQSR unit cell. The radiation from the unit cell in terms of leakage loss is varying with low fluctuation from 0.8-0.87 within the working frequency range which is desired for good antenna design. **In Fig. 2(b), the effect of variation in inter-spiral gap (S_h or S_v) to determine the balance condition are shown where it varies from S_h=S_v= 0.4mm (with open stop band) to S_h=S_v= 0.24mm (open stop band is eliminated)**.

3 Equivalent Circuit Model

In this proposed LWA, CRLH structure is realized in HMSIW and the equivalent circuit for the proposed structure is modeled by Advanced Designed System software where dispersion characteristic is analyzed through optimization. **Primarily, initial values of the equivalent circuit are predicted from S-parameter responses of unbalanced unit-cell (HFSS) by fitting the circuit model in ADS. Further, the equivalent model parameters are tuned in ADS to achieve balance condition. The change in circuit elements values from the initial prediction gives an insight into the direction of required changes in physical parameters of unit cell in HFSS design in order to achieve balanced condition. For example, the change in values of L_{m12}-C_{m12} indicates the corresponding change in physical parameter S_h. Similarly, change in L_{m1} and L_{m2} can be related to corresponding change of S_v. The width and turns of the complementary spiral can be modeled by the equivalent circuit elements L_1-C_1 (for spirals close to the via-wall) and L_2-C_2 (for spirals close to the magnetic wall)**. The detailed and precise equivalent circuit for the proposed HMSIW CQSR unit cell is shown in Fig. 4 where encircled area of the circuit is showing the effect of mutual coupling between the single quad of complementary spirals, neglecting the diagonal mutual coupling. L_1, C_1, L_2, C_2 are coming for quad of spirals which are represented by four L-C tank circuits. The parallel inductors (L_{v1} and L_{v2}) and capacitors (C_{op1} and C_{op2}) are coming due to the metallic vias and considering the fringing field at the magnetic wall respectively. All the optimized inductors and capacitors values are mentioned in Fig. 4. The comparison between simulated and optimized dispersion characteristic

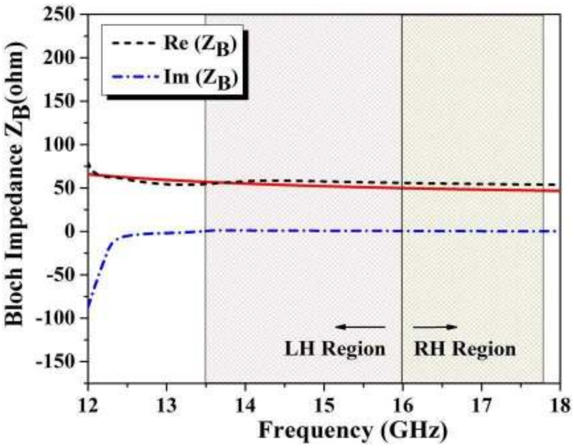

Fig. 5: Real and imaginary part of Bloch impedance Z_B Ω for the proposed structure as a function of frequency. The red line is the impedance at the port of unit cell.

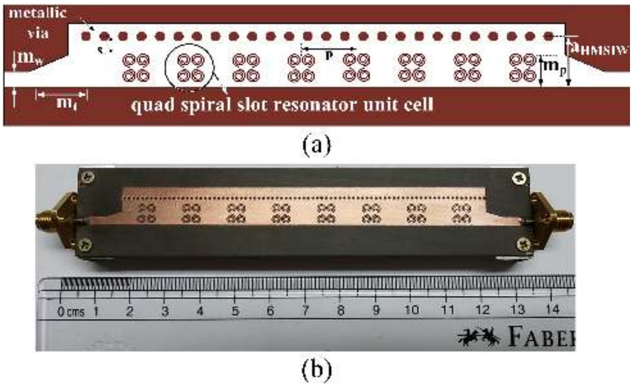

Fig. 6: Microstrip line fed proposed HMSIW based CRLH CQSR leaky-wave antenna with tapered microstrip transition. The parameter values are: a_{HMSIW}=7.47mm, m_p=3.61mm, m_t=4.35mm, m_w=2.29mm, S_a=1.6mm, p=12mm.
a layout of the proposed antenna
b fabricated prototype with 8 unit cells

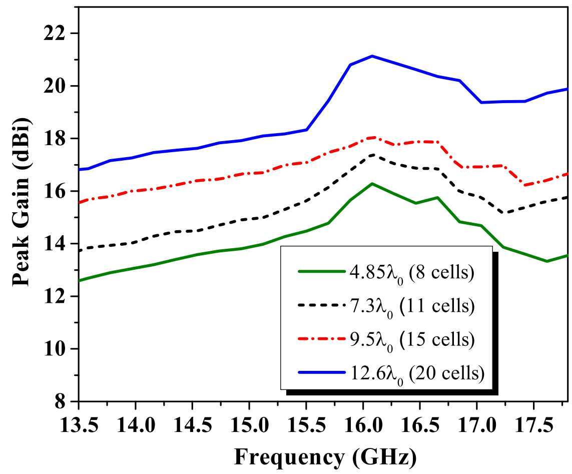

Fig. 7: **Variation of simulated gain with frequencies for different lengths.**

are depicted in Fig. 3(a) and it is showing a good agreement.

$$Z_B = Z_L\sqrt{\frac{\left(\frac{\omega}{\omega_{se}}\right)^2 - 1}{\left(\frac{\omega}{\omega_{sh}}\right)^2 - 1} - \left(\frac{\omega_L}{2\omega}\left[\left(\frac{\omega}{\omega_{se}}\right)^2 - 1\right]\right)^2} \tag{4}$$

where,

$$Z_L = \sqrt{\frac{L_L}{C_L}} \qquad \omega_{se} = \frac{1}{\sqrt{L_R C_L}}$$
$$\omega_{sh} = \frac{1}{\sqrt{L_L C_R}} \qquad \omega_L = \frac{1}{\sqrt{L_L C_L}} \tag{5}$$

$$Z_B = \pm Z_0\sqrt{\frac{(1+S_{11})^2 - S_{21}^2}{(1-S_{11})^2 - S_{21}^2}} \tag{6}$$

The *Bloch* wave analysis can be performed through full wave analysis by considering infinitely long structure. Here, *Bloch* impedance is extracted from N number of CRLH unit cell i.e. by finite unit cell approach as [32], [33] and is shown in Fig. 5. Mutual coupling is taken into account during analysis. By neglecting the radiation resistance, the *Bloch* impedance Z_B for symmetric CRLH unit cell is expressed as [32], [34] and mentioned in (4)-(6). From Fig. 5 it is clear that the real part of the *Bloch* impedance is almost matched with the characteristic impedance Z_0 of the transmission line within the desired frequency range (LH and RH regions are depicted shaded in figure) of the proposed antenna. In that region, the reactive part of Bloch impedance is almost zero, signifies good matching throughout the working frequency region of the antenna.

4 Leaky-wave Antenna Design

Periodic LWAs are generally realized by incorporating periodic perturbations to the guided mode of the structure such that n = 0 becomes radiating harmonic in nature. The unit cell shown in Fig. 1 is placed periodically in series with the periodicity p (maintaining the homogeneity condition i.e. $p \ll \lambda_g/4$) in such a manner that the scanning range of the leaky-wave antenna will be in the fast wave region (13.5-17.8 GHz) where n^{th} space harmonic phase constant (β_n) increases from negative (β_n = -k_0) to positive (β_n = k_0) values with increment of frequencies. Moreover, the average cell size p must be substantially smaller than the guided wavelength λ_g to satisfy the homogeneity condition. In the fast wave region, β_n is less than the free space wave number i.e. $|\beta_n|$ <k_0 which is necessary radiation condition of LWA. Fig. 6(a) depicts the layout of the proposed LWA having eight unit cells. In order to match the structure with the 50Ω microstrip line, taper-line transitions are used at both ends. The m_t and m_p are optimized for the purpose of good matching. The phase constant of the n^{th} spatial harmonic β_n determines the direction of radiated main beam measured from broadside direction to both sides of the visible space. The angle of maximum radiated beam direction (θ_m) can be calculated in dominant mode n = 0 as [18].

$$\theta_m = sin^{-1}\left(\frac{\beta_n}{k_0}\right) = sin^{-1}\left(\frac{\beta_0}{k_0} + \frac{n\lambda_0}{p}\right) \tag{7}$$

The equation (7) shows that a full space scanning (-90° to 90°) can be achieved if β_n varies with the range (-k_0, k_0). Fig. 7 shows the variation of simulated peak gain with frequencies for different radiator lengths where gain is increasing with radiator length and is maximum at broadside. The proposed antenna stands as a very good candidate in terms of compactness, frequency scanning range and gain for the applications in this frequency range.

5 Experimental Verification and Discussion

The proposed HMSIW CRLH CQSR based leaky-wave antenna are fabricated on Rogers RT/duroid 5880 substrate having the dielectric constant of 2.2, loss tangent (tanδ) of 0.0009 and thickness of

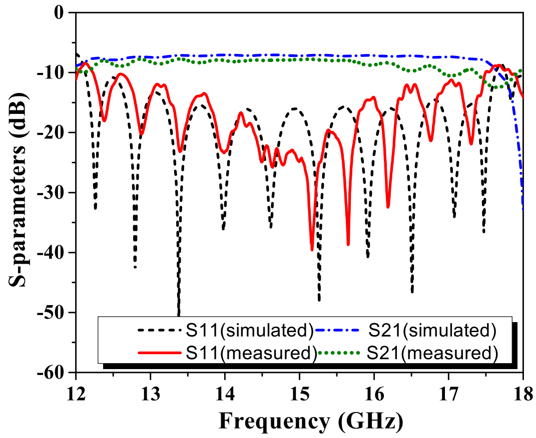

Fig. 8: Comparison of simulated and measured S-parameters with frequency

Table 1 COMPARISON WITH OTHER REPORTED DESIGNS

Ref	Radiator Length	Bandwidth (GHz)	Scanning Range	Peak Gain (dBi)
[19]	$22.82\lambda_0$	26 - 27.5 (5.6%)	10°	16
[20]	$11.57\lambda_0$	8 - 11.4 (35.05%)	35°	21
[24]	$7.44\lambda_0$	5 - 7 (33.33%)	155°	10.6
[26]	$12.95\lambda_0$	20 - 30 (40%)	75°	14
[27]	$5.74\lambda_0$	13.5 - 17.8 (27.47%)	87°	9.4
[30]	$6.6\lambda_0$	8.4 - 11.4 (30.3%)	86°	11
[34]	$5.67\lambda_0$	8.6 - 10.3 (17.98%)	86°	11.5
This Work	**4.85 λ_0**	**13.5 - 17.8 (27.47%)**	**86°**	**16**

λ_0: Wavelength at center frequency

0.787 mm shown in Fig. 6(b). The overall dimensions of the proposed antenna (including feeding network) are 121.5 x 24.5 x 0.787 mm^3 ($4.85\lambda_0$ long) which is much more compact compared to other reported designs with promising property of broadband scanning range at Ku-band. Fig. 8 shows the simulated and measured S-parameters of the proposed antenna which exhibits an impedance bandwidth of 27.47% (simulated) and 25.80% (measured) within the radiating fast wave region. The passband before 13.5 GHz signifies the bounded wave propagation and there is a stopband beyond 17.8 GHz. Fig. 9 shows the comparison between simulated and measured normalized **H-plane (y-z plane)** radiation patterns with frequencies where all of the beam patterns reveals the property of balanced CRLH LWAs, backward to forward continuous frequency beam scanning. In Fig. 8, the radiation patterns at 13.5 GHz to 15 GHz correspond to the LH radiation regions, 16 GHz corresponds to broadside radiation and up to 17.8 GHz correspond to RH radiation regions. **By changing the periodicity p, the transitional frequency of broadside radiation can be tuned keeping a fair scanning range on both sides from broadside direction. Thus, scanning range can be modified in RH region due to increase in positive angles. The measured minimum cross-polarization level are achieved as -20.04 dB at 13.5 GHz, -26.89 dB at 14 GHz, -29.26 dB at 14.5 GHz, -19.82 dB at 15 GHz, -25.45 dB at 16 GHz and -23.05 dB at 17.8 GHz for the radiated main beam directions of -66°, -43°, -30°, -17°, 0° and 20° respectively**. The 3-D radiation pattern of the proposed antenna is shown in Fig. 10 where radiated fan beam is scanning with the range of 86° in visible space (y-z plane) with frequencies. Variation of measured peak gain and simulated radiation efficiency with frequencies of the above mentioned antenna are shown in Fig. 11. Measured antenna gain varies from 12 - 16 dBi and the estimated radiation efficiency [25] is varied from 82% to 85% within the desired frequency range. The proposed antenna is advantageous in terms of compactness and gain with good scanning angle capabilities than other reported designs that concluded in Table 1.

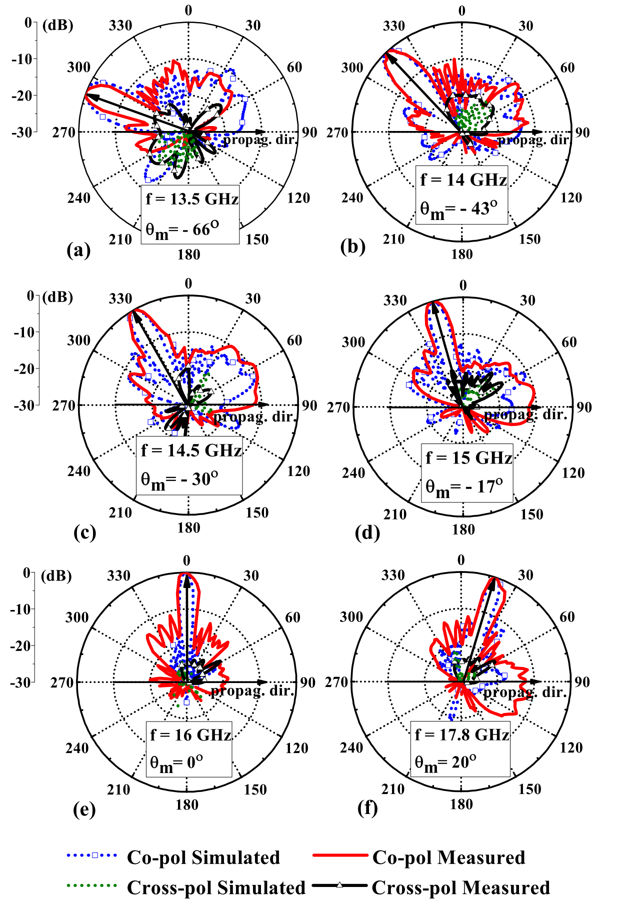

Fig. 9: **Comparison of simulated and measured normalized H-plane (y-z plane) radiation patterns at (a) 13.5 GHz, (b) 14 GHz, (c) 14.5 GHz, (d) 15 GHz, (e) 16 GHz, (f) 17.8 GHz.**

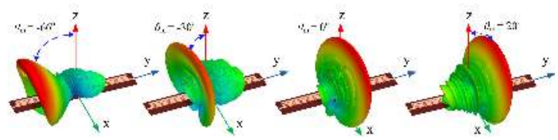

Fig. 10: 3-D fan-beam radiation patterns at 13.5 GHz, 14.5 GHz, 16 GHz and 17.8 GHz of the antenna.

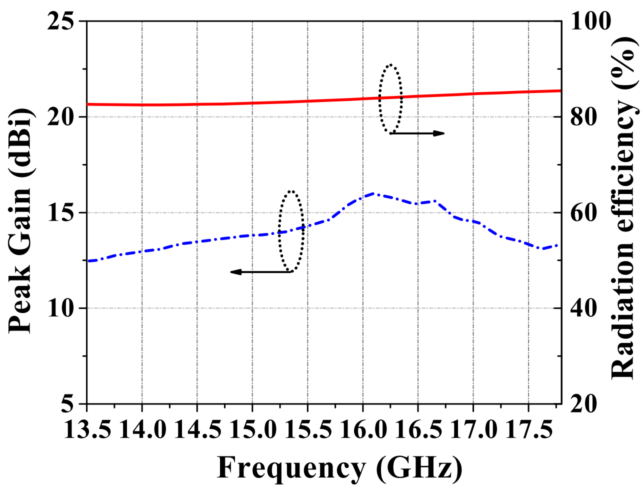

Fig. 11: Variation of measured peak gain and simulated radiation efficiency as a function of frequency for the proposed structure of $4.85\lambda_0$.

6 Conclusion

A compact frequency beam scanning high gain complementary quad spiral resonator based (balanced CRLH) HMSIW LWA is presented at Ku band. Matched balanced condition is achieved for the unit cell. Thus the proposed geometry is able to scan from backward to forward including broadside direction within a wide frequency range of 13.5-17.8 GHz with a scanning range of 86° (-66° to 20°) maximum simulated and measured gain of 16.28 dBi and 16 dBi respectively. **The simpler and compact design methodology along with easier tuning capability for further enhancing gain and radiation efficiency make this proposed antenna as an appealing candidate for many practical applications in Ku-band like Fixed Satellite Services (FSS), Broadcast Satellite Services (BSS)**.

7 Acknowledgments

The authors would like to thank Dr. Arun K. Bhattacharyya, Northrop Grumman Systems Corporation for his valuable suggestions.

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