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# Design of Compact Transverse Slot Array Antenna Using Corrugated H plane Horn

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*Abstract*—**A compact, high gain, series fed transverse slot array antenna is designed on the top plate of a shorted corrugated H-plane horn. Corrugations are used to reduce the wavelength inside the shorted horn in order to suppress the grating lobe problem associated with the transvers slot arrays. By etching eight transverse slots separated by one guided wavelength, the array is formed and a proper broadside radiation is successfully achieved. The proposed structure can also be interpreted as a flat horn with directive radiation which is also more compact compared to the conventional H-plane horns as the guided wavelength inside it is reduced. The antenna is simulated by HFSS simulator and optimized for maximum gain and best matching condition. The simulation results show that the reflection coefficient is less than -10 dB around 15.15 GHz with about 1.3 % fractional bandwidth and the gain and SLL are about 19 dBi and -13 dB, respectively. Also, the cross polar discrimination of the antenna is better than 25 dB in both E- and H-planes. Simulation results were confirmed by comparing them with those obtained by CST software.**

***Keywords—slot; transverse; corrugation; compact; graiting lobe.***

## I. INTRODUCTION

Nowadays, high rate and long-range communication links are popular and the key component of these links is a high gain and efficient antenna. Several technologies are proposed to design high gain antennas including microstrip, SIW and hollow waveguide structures. Substrate based antennas are compact and lightweight, however they suffer from dielectric losses which limits their efficiency and applications. Therefore, hollow waveguide antennas are used widely for low-loss applications.

Slotted waveguide antennas are widely used in modern communication systems due to the growing demand of high gain flat antennas with high-power handling capability. In [1-6] some designs for this class of antennas can be found. Slot arrays can be implemented either in series-fed or parallel-fed forms. The parallel fed slot arrays require a corporate feed network which makes them quite complicated for large arrays. However, the series fed slot arrays benefit from very simple structure with lower number of layers but at the cost of narrower bandwidth compared to the parallel-fed counterparts.

The arrangement of radiating slots in slot array antennas can be longitudinal [1-2], [7-8], inclined [9] or transverse [6, 10-12] according to the feeding mechanism and desired radiation performance. In series fed arrays with transverse slots dealing with the grating lobe problem is usually a challenge as the spacing between the adjacent slots have to be one guided wavelength which is more than the free space wavelength in conventional waveguides. Different techniques such as using substrate-based waveguides

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[6, 10] or applying external horns [12] have been used to deal with the grating lobe problem.

In in this paper, an H- plane horn is used to feed an array of transverse slots to form a series fed slot array antenna. As the slots are located on the broad-wall of an H-plane horn, we can interpret the proposed structure as a flat horn with directive radiation. With conventional horns, the directive radiation and high gain can be achieved with a bulky structure. Our design is a proof of concept for development of a flat horn with directive radiation and high gain but at the cost of increased cost and complexity relative to the conventional bulky horns. So, the proposed flat horn is more compact compared to the conventional bulky horns but at the cost of higher complexity and cost.

In the proposed structure corrugations are used [13] in the feeding waveguide horn in order to reduce the distance between radiating transverse slots and consequently resolve the grating lobe problem. This antenna can be used in radar systems or in point-to-point communication links. This paper is organized as following.

In section II, making use of corrugations inside the H-plane horn is discussed. Moreover, the design of radiating part including the transverse slots on the top plate of the horn is presented. In section, III we optimize the proposed antenna for the best radiation and impedance matching condition. In section IV, we present the simulation results which show a high gain antenna with appropriate SLL and return loss. Finally, in section V, some concluding remarks are discussed.

## II. ANTENNA STRUCTURE

Fig. 1 shows the proposed structure and its geometrical parameters. We see that the transverse slots are etched on the top plate of a shorted end H-plane horn. The transverse slots are laid in *x*- direction while the array is formed along *y*- direction. The design guidelines for dimensions of feeding H-plane horn can be found in [14].

As shown in Fig. 1, in this design eight transverse slots with the lengths of $l_{si}$ ($i$=1, 2, …,8) in $x$ direction and the widths of $w_s$ are used as a linear array in $y$ direction. In order to achieve a broadside radiation, the slots must be excited in phase. Therefore, the distance between slots should be equal to $\lambda_g$ where $\lambda_g$ is the guided wavelength of the propagating wave inside the horn. As the electromagnetic wave propagates through the flare part of the horn, its field distribution is changing from $TE_{10}$ to Quasi-TEM mode due to the enhancement of propagation constant (i.e., the reduction of $\lambda_g$) caused by the gradual expansion of the waveguide width. On the other hand, in order to suppress the grating lobe effect of the array factor, the spacing between the neighboring radiating slots, $d_{si}$ ($i$=1, 2, …,7), should be less than 0.8 $\lambda_0$, where $\lambda_0$ is the free space wavelength. It should be noted that as $\lambda_g$ is gradually changing inside the horn flare part, the $d_{si}$ value changes as well.

In order to modify the guided wavelength inside the horn, corrugations have been adopted in the flare part of the horn [13, 15]. The corrugations and their geometrical parameters are shown in Fig.1. We see that the corrugations are laid transversely relative to the wave propagation inside the horn. The modified guided wavelength inside the corrugated area of the horn, $\lambda_g'$, can be calculated by the following equation:

$$\frac{1}{{\lambda'_g}^2} = \frac{1}{{\lambda_g}^2} + \frac{1}{{\lambda_y}^2} \qquad (1)$$

where $\lambda_g$ is the guided wavelength inside the horn without corrugations and $\lambda_y$ is the solution of the following equation:

$$\frac{P}{P+a}\frac{\lambda_y}{\lambda_g}\frac{\tan\left(\frac{2\pi h_1}{\lambda_g}\right)}{\tanh\left(\frac{2\pi(h-h_1)}{\lambda_y}\right)} = 1 \qquad (2)$$

The variations of $\lambda_g'$ with respect to the dimensions of corrugated surface are depicted in Fig. 2 at the middle of the horn (width=36mm). It is obvious from this figure that the wavelength inside the corrugated horn can be well controlled by corrugations. In this design, the dimensions of the corrugations are selected as $P$=3.1mm, $a$=1.2 mm and $h_1$=3 mm so that $\lambda_g'$ becomes less than 0.8 $\lambda_0$. It is worth noting that these values are selected with respect to the fabrication restrictions.

Fig. 3 shows the field distribution of the propagating wave inside the H-plane horn for the two cases of with and without corrugations. We see that the guided wavelength inside the corrugated one is shortened which reveals that the corrugated one is presenting more compact structure. In Fig. 3 it can also be observed that at the middle of the corrugated horn, the wavelength is about 14 mm which is compatible with the contours shown in Fig. 2. This value satisfies the grating lobe suppression condition, hence we expect to design a high gain antenna with proper radiation performance if we place the transverse slots at the null locations of electric field distribution with respect to Fig. 3-b. It should be noted that at the nulls of a $TE_{10}$/QTEM wave electric field distribution, the current vectors are perpendicularly cut by the transverse slots resulting in efficient radiation through the slots.

The applied transverse slots are non-resonant and their lengths are long enough to cover the whole waveguide width in order to cut the maximum surface current distribution on the top plate leading to maximum directivity and aperture efficiency. As the waveguide width is expanding in the horn flare part, the lengths of the slots are increasing as well. So, the slots lengths are dependent to the horn flare angle. The width of the slots controls the magnitude of radiated power through each slot by which we can control the amplitude taper across the antenna aperture in E-plane (i.e., the y-z plane in Fig.1). Here, the slot width is considered to be the same for all radiating slots for design simplicity. Our goal is to achieve a uniform amplitude distribution across the aperture in E-plane in order to get maximum directivity and gain.


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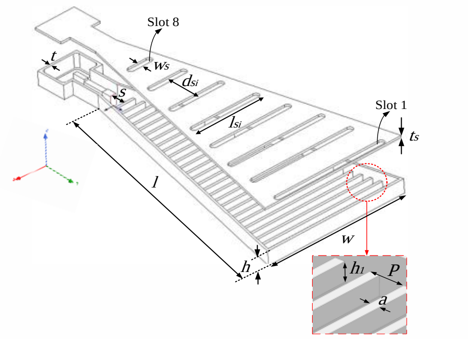


Figure 1. 3D distributed view of the proposed antenna topology.

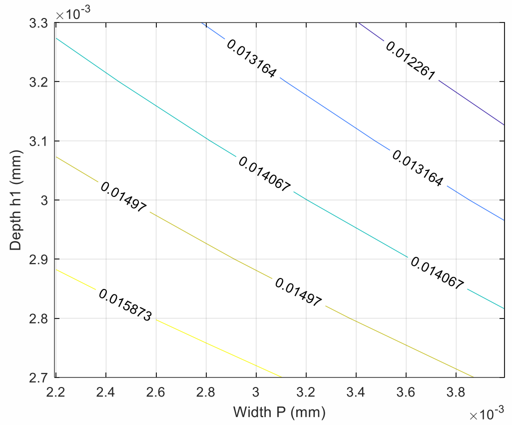


Figure 2. Guided wavelength $\lambda_g'$ at the middle of corrugated horn as a function of the corrugated surface dimensions at 15 GHz.

## III. Final Antenna Design

The initial values for geometrical parameters of the radiation layer are selected so that there is one guided wavelength (which is less than 0.8 $\lambda_0$) spacing between adjutant transverse slots. The lengths of non-resonant slots are following the width of the horn flare part. We have tuned the slots' width and spacing using full-wave simulation, in order to achieve a uniform phase and amplitude distribution across the aperture in E-plane in order to get maximum directivity and gain. The tuning was done using parametric sweep in HFSS simulator to obtain the optimum values. The error of convergence for the meshing was set to be less than 2% in two consecutive passes and the minimum number of meshing passes was set to six.

The designed antenna is excited by WR62 standard waveguide via a ridged waveguide. A stepped transition is applied to the ridge in order to provide proper matching between WR62 and the horn. Fig. 4 shows the geometrical parameters of the transition which are adjusted by full wave simulation using parametric sweep in HFSS simulator to obtain the best impedance matching. The final dimensions of designed antenna are listed in table I.

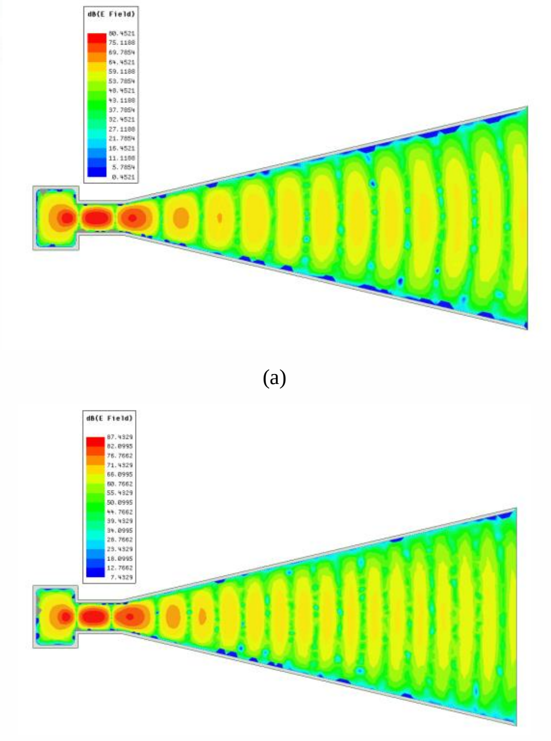


Figure 3. Electric field distribution inside an open-ended a) conventional H-plane horn b) corrugated H-plane horn.

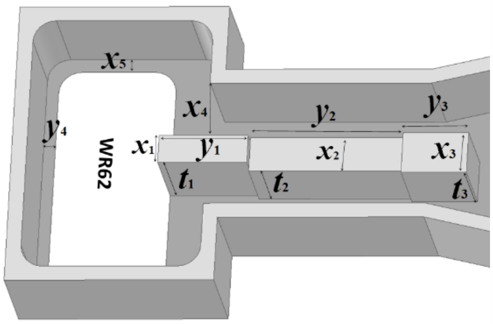


Figure 4. 3D view of the transition with its geometrical parameters.

TABLE I. GEOMETRICAL PARAMETERS OF THE DESIGNED ANTENNA

| Parameter | Value (mm) | Parameter | Value (mm) |
|---|---|---|---|
| $l$ | 127.22 | $l_{s6}$ | 24 |
| $h$ | 7 | $l_{s7}$ | 16 |
| $t$ | 2 | $l_{s8}$ | 9 |
| $w$ | 68.42 | $S$ | 4.2 |
| $t_s$ | 1 | $w_s$ | 3 |
| $d_{s1}$ | 15 | $x_1$ | 1.9 |
| $d_{s2}$ | 14 | $x_2$ | 2.4 |
| $d_{s3}$ | 14 | $x_3$ | 2.85 |
| $d_{s4}$ | 14 | $x_4$ | 3.35 |
| $d_{s5}$ | 15.4 | $x_5$ | 0.9 |
| $d_{s6}$ | 16 | $y_1$ | 6 |
| $d_{s7}$ | 15 | $y_2$ | 10.3 |
| $l_{s1}$ | 52 | $y_3$ | 4.4 |
| $l_{s2}$ | 48 | $y_4$ | 0.8 |
| $l_{s3}$ | 45 | $t_1$ | 6 |
| $l_{s4}$ | 37 | $t_2$ | 5 |
| $l_{s5}$ | 31 | $t_3$ | 5.3 |



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## IV. RESULTS

The antenna shown in Fig. 1 with the dimensions listed in table I was simulated using HFSS and the results are shown through Figs. 5 to 11.

In Fig. 5 the simulated reflection coefficient is depicted. In this figure the simulated result obtained by CST is also shown for comparison and good agreement between both results obtained by CST and HFSS can be observed. We see proper impedance matching ($S_{11}$<-10dB) over 15.08-15.28GHz frequency range which shows about 1.3% impedance bandwidth. In the proposed structure making use of series fed simple slots as narrowband radiating elements along with using transverse corrugations along the feeding wave path which tends to suppress the wave propagation inside the waveguide has resulted in hard impedance matching and consequently narrow bandwidth.

Fig. 6 shows the electric field distribution inside the antenna structure. We can see that all radiating slots are excited quite in phase which results in broadside radiation of the antenna. It can also be observed that as the wave propagates through the horn, the guiding wavelength and consequently the distance between the slots are decreased as the $TE_{10}$ mode is converted to TEM mode. It can be predicted by the following equations [16] that as the wave is propagating along the horn, the exciting power for each slot is decreased as a part of feeding wave power is radiated trough the previous slots:

$$P_{rad\_i} = \eta(1-\eta)^{8-i}P_0 \quad i=1,2, \ldots, 8. \tag{3}$$

$$P_{out} = (1-\eta)^{9-i}P_0 \tag{4}$$

where $P_{rad\_i}$ and $P_{out}$, are the radiation power and residual power associated with $i$-th slot respectively. The parameters $\eta$ and $P_0$ are the radiation efficiency of each slot and the input power, respectively. The $\eta$ value is mainly controlled by the slot width. In this design, as all the slots have the same width, the $\eta$ value is constant for all of them.

The antenna radiation pattern in H- and E- planes is shown in Fig. 7. In this figure the results obtained by HFSS are compared with that achieved by CST and good agreement between them can be observed. We see that, the designed antenna has a proper radiation performance in which the grating lobes are successfully suppressed. The SLL is about -13 dB, as theoretically expected for a quite uniform illuminated aperture. The front to back ratio (FTBR) is better than 25 dB and a broadside radiation can be observed which confirms that all radiating slots are excited in phase. It should be noted that our goal is to obtain maximum directivity and gain which theoretically leads to the side lobe level of -13 dB when the aperture is uniformly illuminated. So, the obtained SLL of -13 dB reveals that in the designed antenna, the radiating slots are uniformly illuminated by the feeding wave in E-plane.

Fig. 8 shows the 3D radiation pattern of the proposed antenna. We see that, the proposed flat horn can provide a broadside, directive radiation pattern in 3D space. This can be regarded as an interesting feature when compared with conventional directive horns which are bulky structures.

Fig. 9 shows the co-and cross-polar radiation patterns of the antenna in both E & H planes at the desired frequency. We see that, the cross polar discrimination is better than 25 dB which indicates high purity of the radiated wave by the designed antenna. To improve this feature, one can adopt edge at the top plate of the designed antenna to suppress the edge effect.

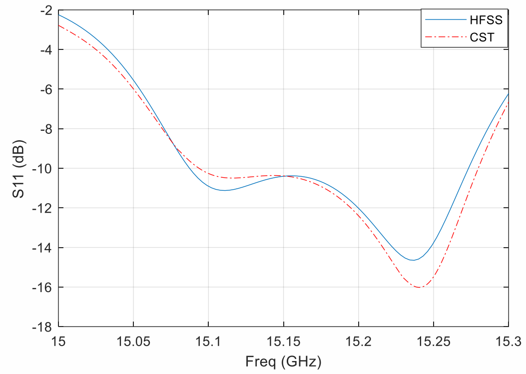



Figure 5. the reflection coefficient of the designed antenna.

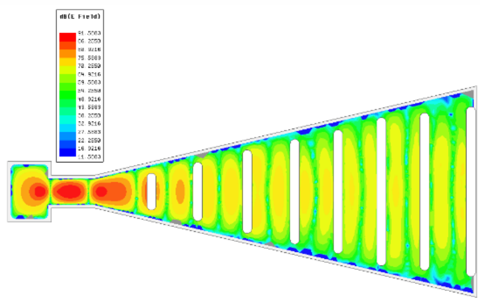
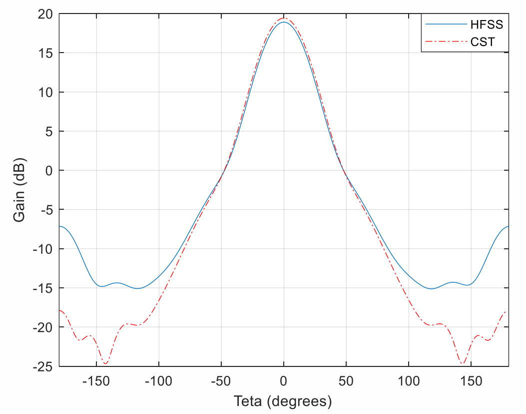


Figure 6. electric field distribution inside the proposed antenna.




(a)



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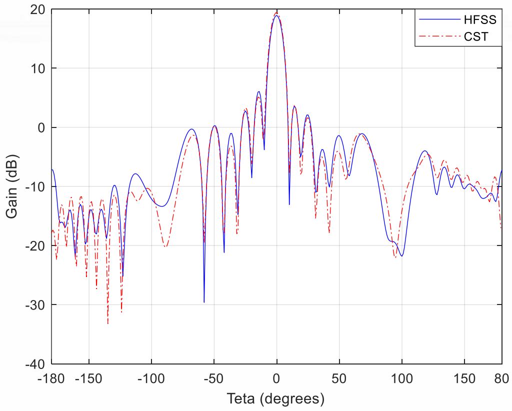



(b)
Figure 7. a) H-plane and b) E-plane radiation pattern of the proposed antenna at 15.15 GHz.

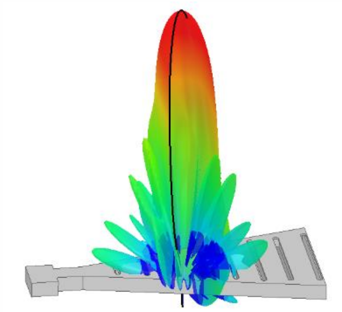
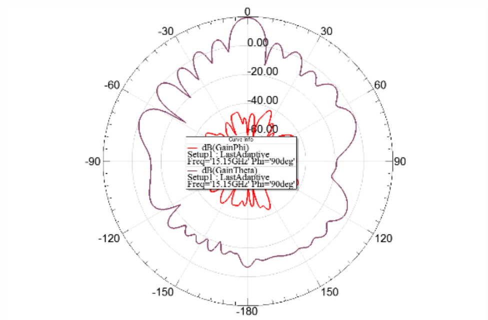
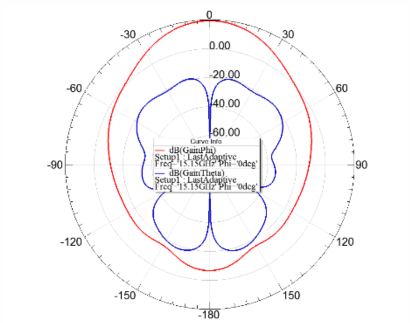


Figure 8. 3D radiation pattern of the proposed antenna



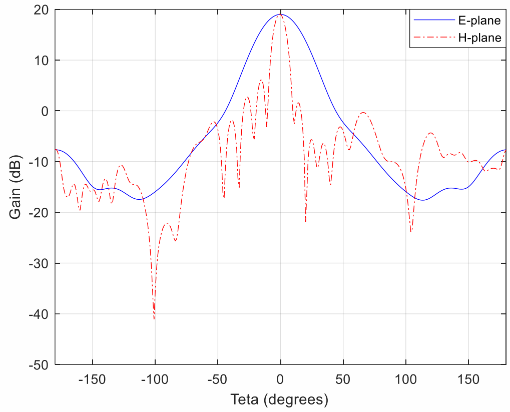
(a)



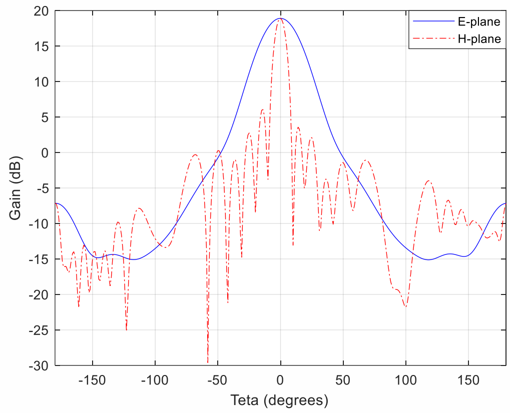
(b)
Figure 9. Co- and Cross polar radiation patterns in a) E and b) H planes.

In Fig. 10, the patterns are shown at three frequencies over the operating bandwidth. We see that the designed antenna provides broadside and directive radiation over its operating bandwidth with the SLL of about -13 dB.

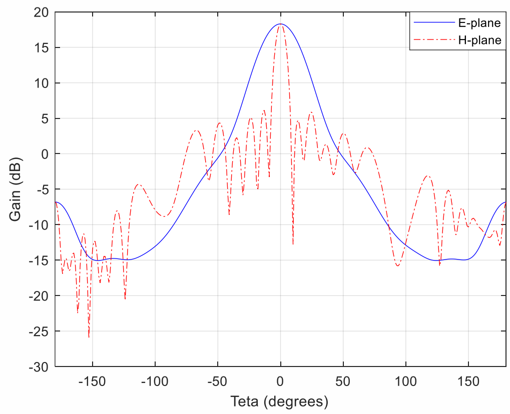
In Fig. 11 the simulated gain plots obtained by both CST and HFSS are depicted and good agreement between them can be observed. Both results show the maximum gain of about 19.2 dBi with the variation of lower than 1.5dB over the operating bandwidth, i.e., from 15.08 to 15.28GHz.




(a)




(b)




(c)
Figure 10. Radiation pattern of the designed antenna at a) 15.08 GHz b) 15.15 GHz c) 15.25 GHz.



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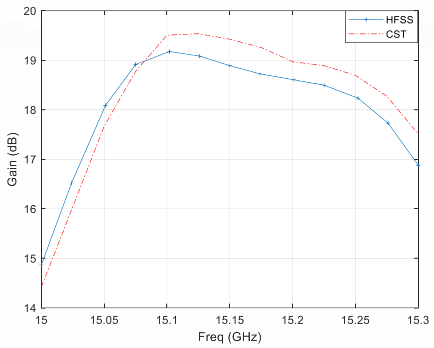


Figure 11. Simulated gain of the proposed antenna versus frequency.

In Table II the proposed antenna performance and features are compared with some recent works presented in the literature. We see that compared to the other one-dimensional arrays, *i.e.*, those presented in [6, 11, 12], the proposed structure in this paper introduces higher gain, with respect to its dimensions, compared to those presented in [6, 11] but at the cost of lower bandwidth. It is noticeable that the proposed antenna can be used in high power applications but the antennas presented in [6] and [11] have power limitation problem as they are dielectric based planar structures.

TABLE II. COMPARISON OF THE PROPOSED ANTENNA PERFORMANCE WITH SOME RECENT WORKS

| Ref. | Technology | Center frequency | Dimensions($\lambda^2$) | No. of elements | Number of layers | Gain (dBi) | BW (%) | Evaluation method |
|---|---|---|---|---|---|---|---|---|
| [6] | SIW | 10 GHz | 1.28λ×0.73λ | 1×4 | 1 | 9.2 | 1.7 | Measurement |
| [11] | SIW | 34.5 GHz | 4.18λ×0.93λ | 1×5 | 3 | 12.6 | 5.54 | Measurement |
| [12] | GGW | 28 GHz | 6.62λ×1.2λ | 1×5 | 1 | NA | 3.5 | Simulation |
| This work | C-Horn | 15 GHz | 1.67×5.23λ | 1×8 | 1 | 19 | 1.3 | Simulation |

BW: band width
NA: not available

## V. CONCLUSION

A series fed transverse slot array antenna was designed using an H-plane horn as the feeding waveguide. To decrease the antenna size and suppress the grating lobes, corrugations were adopted inside the feeding H plane horn. The transverse slots were etched on the top plate of corrugated horn at the null locations of electric field distribution to perpendicularly cut the surface currents leading to efficient radiation through the slots. A ridged waveguide with stepped transition is used for impedance matching of the antenna with the standard WR62 waveguide. The designed antenna was optimized to achieve maximum gain and directivity using HFSS simulator. Simulation results show a high gain antenna with pencil beam radiation pattern and without grating lobes which confirms that the proposed structure can be considered as a good candidate as a low loss flat antenna for realization directive radiation.

Further works can be done to increase the bandwidth of the proposed high gain antenna.

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