

Design and Analysis of 15.4dBi, 8-Element Phased Array Dipole Antenna resonating at 2.0GHz for Mobile Communication Applications

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Abstract:

This paper presents the design and simulation of the proposed 15.4dBi, 8-element phased array dipole antenna at 2.0GHz for mobile communication applications using MOMNEC2 and arraycalV2.4 simulators. The paper considered the effects of changing the magnitude and phase excitation of the signal to each element while keeping the geometrical configuration of the antenna uniform. The optimized results show that the simulated value of directivity D is 15.4dBi which is 0.3 higher than the chosen design value of 15.1dBi. Also, there is a slight variation in the -3dB beamwidth values; the simulated value is 15° while the designed value is 12.7°. The analysis also show that the minimum computed SWR is 1.36dB, and the computed reflection coefficient and return loss are respectively 0.15 and 16.3dB, which show that the impedance bandwidth of the antenna is 1575 – 2460MHz.

Keywords: 8-Element Phased Array Dipole, Directivity, Standing wave Ratio, Mobile Communication.

1.0 Introduction

Traditionally, most wireless communication systems make use of the conventional dipole antenna to receive and transmit radio signals. This is due to the facts that they are simple to design, construct and well suited for wireless communications applications. Though they are omnidirectional, they have poor antenna gain. Similarly, conventional reflector (popularly called dish) and horn antennas have been employed for satellite communication due to directional narrow beam and high gain. There are certain applications where they cannot be used such as where the size and rigidity of the conventional antenna constrain the performance of radio system. For instance, since they are relatively large in size they protrude from the mobile phone handsets or laptop computers. The problem of obstructive and space demanding structure also complicate any effort taken to equip a mobile communication devices with multiple antennas. In recent years, significant increase in the use of compact communication devices has placed demand on the application of phase array antenna, PAA technology, which in turn has intensify research in the area [1 – 4].

The PAA is made up of the array of radiating elements and the RF feed network (beamforming and

power distribution networks). The basic concept of the PAA is to excite the multiple radiating elements from the same source such that the components of the radiating fields will constructively add in a particular direction and destructively cancel in others, and hence by increasing the number of elements in the array high directivity is realized. PAA has the capability to communicate with several targets which maybe anywhere in space simultaneously and continuously because the main beam of the antenna can be directed electronically to a specified direction. This particular benefit help curtail the energy wastage in undesired direction as found with the conventional omnidirectional antennas. Also, with the main beam steering of the PAA the multipath or pathloss or fading effects of the communication link are reduced and hence, the channel capacity is increased [2, 4, 5]. A secondary aspect of PAA is the ability to improve the security of the signals by adding null in the direction of unwanted signal as an example a hacker trying to gain access to a private network or an enemy trying to jam the communication system.

The sensitivity of the beam forming network for the PAA system brings about calibration and fabrication process that is costly and complex thereby limiting the impressive potentials of the PAA, due to the high cost of production. Although, later on, solid state active PAA became centre of attraction due to their low weight and power dissipation and high output power and reliability, the cost of production was not reduced, which limited its usage to the military sector only for radar applications. The major limitations of the active PAA are circuit complexity and very high cost of production, which have hindered the widespread use of the antenna system in the public sector. Recent advances in PAA technology have made the dream of low-cost and high performance integrated phased array antenna a reality [6 – 9]. This study presents the design and simulation of the proposed 8-element phased array dipole antenna at 2.0GHz for mobile communication applications using

MOMNEC2 and arraycalV2.4 simulators. The paper focuses on the effects of changing the magnitude and phase excitation of the signal to each element while keeping the geometrical configuration uniform.

The circuit complexity and cost which are two major challenges facing the designer of PAA at LBand are addressed, by investigating the performance of using a linear dipole antenna array configuration and passive analogue reflective phase shifter with lumped element model of the quadrature hybrid 3dB coupler.

2.0 Design and Simulation

2.1 The Array Configuration

The radiating element of the array antenna is composed of 8 dipole elements with the operating frequency f_0 centered at 2.0GHz. The array is uniformly excited over a ground plane with the dipoles placed 0.25λ above the ground. The wavelength, λ is obtained from the fundamental electromagnetic equation:

$$c = f_0 \lambda, \quad (1)$$

Where c = speed of light in vacuum.

A linear array with its peak at θ in degrees (beamwidth) can also have other peak values subject to the choice of the element spacing d in meter. A criterion for determining the maximum element spacing d for an array scanned to a given angle θ_0 at frequency f_0 for grating lobe not to occur is to set d to satisfy the condition [4, 5]:

$$\frac{d}{\lambda} \leq \frac{1}{1 + \sin \theta} \quad (2)$$

In this study θ_0 is chosen as 15° , while d varied from 0.7λ to 0.5λ , and from the simulation result the optimized

value of 0.5λ is obtained for d . The length of the array is given by:

$$L = nd \lambda \quad (3)$$

Where n = number of elements in the array, $d\lambda$ = element spacing in terms of wavelength. The computed length is $L = 4.0\lambda$. The width, W of the array corresponds to the length of the $\lambda/2$ dipole which is 0.5λ , which in turn is the array length in the X-direction and L is the array length in the Y-direction.

3dB beamwidth (BW)

For a uniformly illuminated array the 3dB BW is given as [10]:

$$\theta_{3dB} = \frac{0.886\lambda}{L} \times \frac{180^\circ}{\pi} \quad (4)$$

Using Eqn. 4, yields: $\theta_x = 101.5^\circ$, $\theta_y = 12.7^\circ$. Where θ_x and θ_y are the 3dB BW in the X and Y directions respectively. In this study, since the polarization is horizontal, the 3dB BW in the Y direction will be used as the 3dB BW of the array.

Directivity (D)

The directivity, D in dBi of the array can be expressed as [5]:

$$D = 10 \log_{10} \left(\frac{41,253}{\theta_x \theta_y} \right) \quad (5)$$

The chosen and computed design specifications of the proposed PAA configuration are summarized and displayed in the Table 1.

Table 1: Summary of Array Configuration

Array type	Linear, vertical
Array element	Half-wave dipole
Polarization	Horizontal
No. of elements	8
Scanned angle	15°
Frequency	2.0 GHz
Vertical 3dB BW	12.7°
Side-lobe suppression	-20dB 1 st side-lobe w.r.t main beam
Directivity	15.1 dBi

2.2 Elements (Amplitude) Excitation

One of the major advantages of PAA is that the array excitation can be controlled to produce extremely low side lobe patterns in very accurate approximations of chosen radiation patterns. There are various procedures that have been developed for synthesizing array factors, namely (i) Synthesis of various patterns that are usually mainly wide beamwidth, (ii) Synthesis of low side lobe, and narrow beam patterns, and (iii) Synthesis that optimize array parameters such as gain and signal to noise ratio subject to some constraint on the side lobe level or the existence of outside noise sources.

For the amplitude taper, the Modified Taylor Distribution synthesis procedure is used. It is a procedure for synthesizing pattern functions with arbitrary first side lobe levels, a far side lobe level similar to that of a uniformly illuminated source and narrow beam patterns according to [10 – 16]. It is used to calculate the element excitation depending on their distance from the center of the array. Since the far-field pattern is a continuous function, rather than a summation, a standard inverse finite Fourier Transform is employed rather than the discrete resulting in an aperture distribution function given as [5]:

$$f(X) = J_0(j \cdot \pi \cdot B \sqrt{1 - x^2}) \tag{6}$$

Where, J_0 is zero order Bessel function of the first kind, x is the position with a 2 unit length aperture i.e. $-1 \leq x \leq 1$,

and Bis Taylor function parameter, which according to [17] for side lobe level 10dB and 25dB, are 0.7386 and 1.0229 respectively. The element amplitude excitations generated are shown in the Table 2.

Table 2: Element Amplitude Excitation

Element no	1	2	3	4	5	6	7	8
Amplitude(dB)	-7.07	-3.26	-1.04	0.00	0.00	-1.04	-3.26	-7.07

2.3 Reflective Type Phase Shifter (RTPS) Design and Phase Excitation

Quadrature Hybrid (3dB) Coupler

This is a four port directional coupler that is used for equal (3dB) power split. The input is split into two signals each of equal amplitude (magnitude) that are 90° apart in phase. The purpose of this block is to isolate the input and the output signals turning the phase shifting behavior of the reflective load into a tunable two-port device. It can be implemented using distributed or lumped elements. The size of a quadrature phase shifter is directly related to the frequency band of operation since the coupler uses one or more quarter-wave sections. The lumped element model gives the RTPS the smallest size at frequency below 10GHz independent of technology used [12, 13].

The coupler could be realized as branch line or lumped element couplers as shown in the Figure 1a,b. The branch line model is simple in design but the major limitation is the large size. The transformation of the branch line hybrid from distributed to lumped form is achieved using the transformation equations given as [14]:

$$C_1 = \frac{1}{\omega_o Z_o} \tag{7}$$

$$L_1 = \frac{Z_o}{\omega_o \sqrt{2}} \tag{8}$$

$$C_1 = \frac{1}{\omega_o 2L_1} - C_1 \tag{9}$$

Where $\omega_o = 2\pi f_o$, and Z_o = the antenna characteristic impedance. This conversion minimizes the number of area consumption and loss.

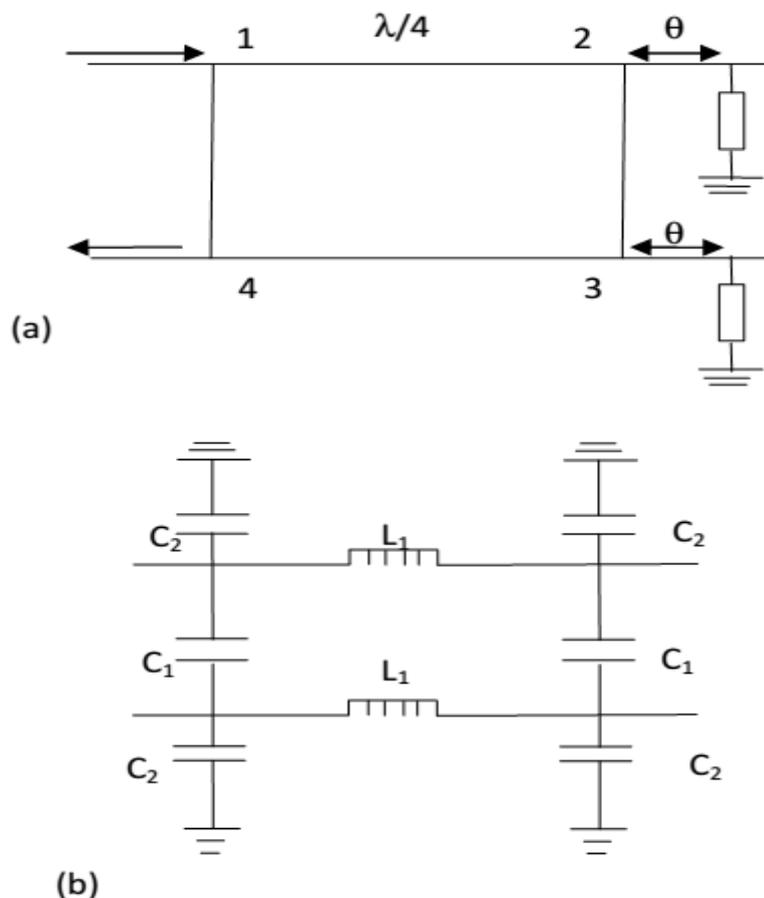


Figure 1: (a) Branch-Line Distributed Coupler; (b) Lumped Element Hybrid 90° Coupler. $C_1 = 1.59\text{pF}$, $C_2 = 0.66\text{pF}$, $L_1 = 2.81\text{nH}$.

Reflective Load

The reflective load is a one port circuit with variable phase reflection characteristic. The operation of the phase

shifter depends upon their reflective networks which are usually the set of valuable reactance terminals at the through and coupled ports. Four very common reflective networks are shown in the Figure 2a-d.

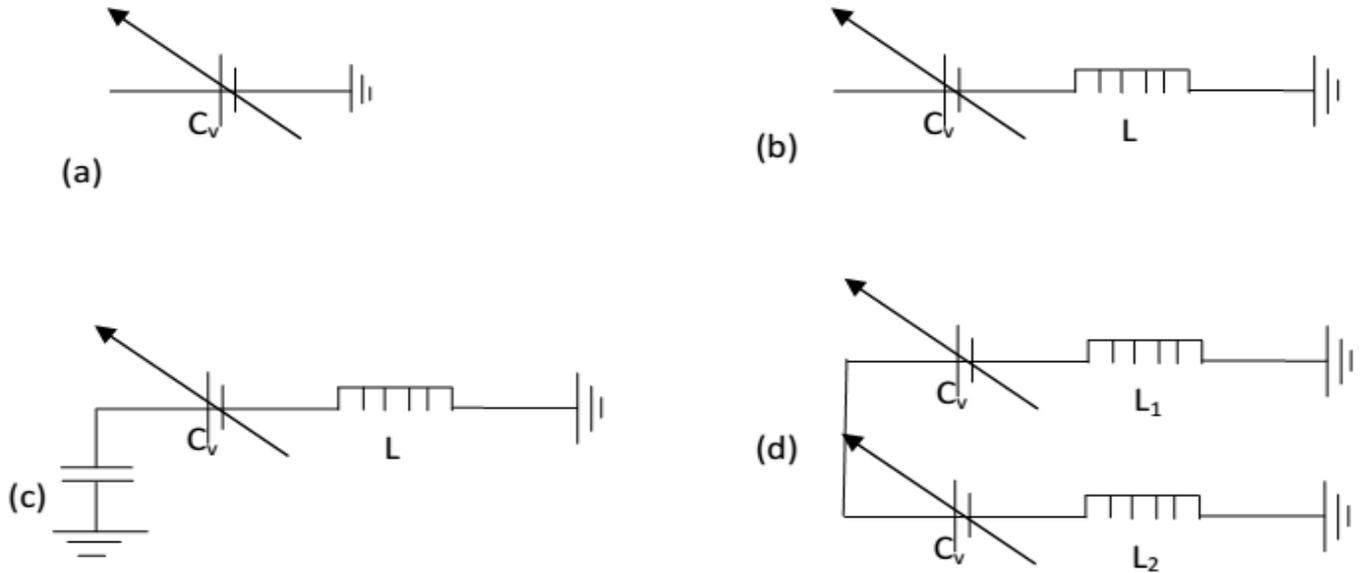


Figure 2: (a) Varactor; (b) Single resonated load (SRL); (c) Transformed single resonated load (TSRL); (d) Dual resonated load (DRL).

The most basic and simplest reflective network, a single varactor is employed in this study. With the single varactor, Figure 2a, phase shift less than 90° could be obtained practically. To increase the phase shift tuning range of the single varactor, a series inductor is added which would resonate with the varactor at the operating frequency as shown in Figure 2b. This network has also reduced circuit complexity.

To obtain the phase control, the capacitance \$C_v\$ of the varactors are varied. \$C_v\$ ranges from \$C_{min}\$ to \$C_{max}\$. The capacitance control ratio, \$r_c\$ and \$C_v\$ are given as:

$$r_c = \frac{C_{max}}{C_{min}} \tag{10}$$

$$C_v = \frac{1}{\omega_o Z_o}, \tag{11}$$

$$C_{max} = C_v \sqrt{r_c}, \tag{12}$$

$$C_{min} = \frac{C_v}{\sqrt{r_c}} \tag{13}$$

\$r_c\$ is chosen as 16, and the resistive parasites are neglected, \$R_L = R_v = 0\$, \$C_v = C_1 = 1.59\text{pF}\$, \$C_{max} = 6.4\text{pF}\$, \$C_{min} = 0.4\text{pF}\$.

The phase control range can be sufficiently increased by resonating the capacitance of the varactor with an inductor \$L\$. The varactor resonated with inductor designed such that:

$$\frac{Z_v}{Z_o} = 2, \tag{14}$$

Where, \$Z_v\$ is the reactance of \$C_v\$, and \$L\$ the load inductor is given as:

$$L = \frac{Z_v}{Z_o} \left(\sqrt{r_c} - \frac{1}{\sqrt{r_c}} \right), \tag{15}$$

The maximum phase control range

$$\Delta\beta = \tan^{-1} \left[\frac{Z_v}{Z_o} \times \frac{1}{2} \left(\sqrt{r_c} - \frac{1}{\sqrt{r_c}} \right) \right], \tag{16}$$

So that for number of elements, the phase control range will be

$$\beta = knd \sin \theta, \tag{17}$$

Where \$k = 2\pi/\lambda_o\$. From equation 17, taking \$d = 0.5\lambda\$ the phase control ranges were calculated and the values obtained are shown in the Table 3.

Table 3: Element (Phase) Excitation

Element no	1	2	3	4	5	6	7	8
Phase Excitation	0.00	46.59	93.17	139.76	186.35	232.94	279.52	326.11

2.4 Feed Network

Types of Feed Network

The very common feeding techniques are namely; Series Feed, Corporate Feed and Space Feed Techniques. For this work, the corporate feed technique (Figure 3) is used to design the beam scanning network for the proposed linear phased array antenna. The feed network is designed to combine the output from the array element

and phase shifters to achieve the tunable progressive phase shift required for the beam steering of the array. Corporate Feed Technique is common in arrays of dipole, open-end guide and patches.

The name is derived after the structure of organization charts, where the feed divides into two or more paths, and then each path divides and so on. The feeds are usually binary but sometimes the divider tree includes 3 way or 5 way dividers depending on the number of array element. It uses equal length feed network and they provide equiphase signal distribution for wide band array. The combined effect of planning the feeder lines in parallel with each other reduces the effective resistance and involves the use of quarter wave transformers to match each section to a 50Ω line source.

Also the width of feed lines can be varied to achieve different magnitude of power division [12].

Feed Implementation Technique

The most commonly used transmission lines and wave guide for implementing feed network are coaxial cables, microstrip lines, rectangular wave guide and strip line. The microstrip due to reasons of low cost and ease of manufacture will be considered and the simple transmission line model.

Power Divider/Combiner Design

A reciprocal power divider known as a power combiner or power splitter is a device for dividing and/or combining the wireless signals. The power divider is an interdependent device calculating a vector sum total of at least two signals. When used as a power combiner, they are used as devices to couple electromagnetic energy from multiple input ports to an output port in a specified manner. The 8-way power combiner would be employed for this work. The corporate feed technique using basic series of two-way power combiners will be cascaded to increase the number of output ports [12 – 16].

The impedance was calculated in accordance to [17] as follows from Table 1. The power (Pwr) distributions were also computed and shown in Table 4.

Table 4: Power Combiner Ratio

Element	1	2	3	4	5	6	7	8
Pwr (dB)	-7.07	-3.26	-1.04	0.00	0.00	-1.04	-3.26	-7.07
Pwr (Lin)	0.1963	0.4721	0.7871	1.000	1.000	0.7871	0.4721	0.1963
Npwr (Lin)	0.0400	0.0961	0.1603	0.2036	0.2036	0.1603	0.0961	0.0400

The normalized power ratio gives the value of radiofrequency power from the antenna, and phase shifter to be received at the 8-way inputs of the power combiner. As shown in the Figure 4, the linear power combined ratio can be calculated as follows:

$$\text{Linear Power combine Ratio } LPR = \frac{Z_1}{Z_2} \tag{18}$$

$$Z_1 = LPR + Z_2 \tag{19}$$

$$Z_1 = Z_c \left(\frac{1+LPR}{LPR} \right) \tag{20}$$

For the ¼ wave matching transformers:

$$Z_{match} = \sqrt{Z_{reg}} \times Z_{load} \tag{21}$$

Where, Z_{load} = is the impedance to be transformed, Z_{req} = required impedance, Z_{match} = matching impedance, Z_c = the line impedance matching transformers = 25Ω.

The 8-way power combiner is made up of 4 basic series 2-way power combiner. The design of the 8-way combiner would incorporate four steps of same manner of calculation of each 2-way power combiner as shown in the Figure 3.

Z_o = characteristic impedance (50Ω) is transformed to the junction impedance Z_1 and Z_2 by the ¼ wave line section $Z_{1(match)}$ and $Z_{2(match)}$ respectively.

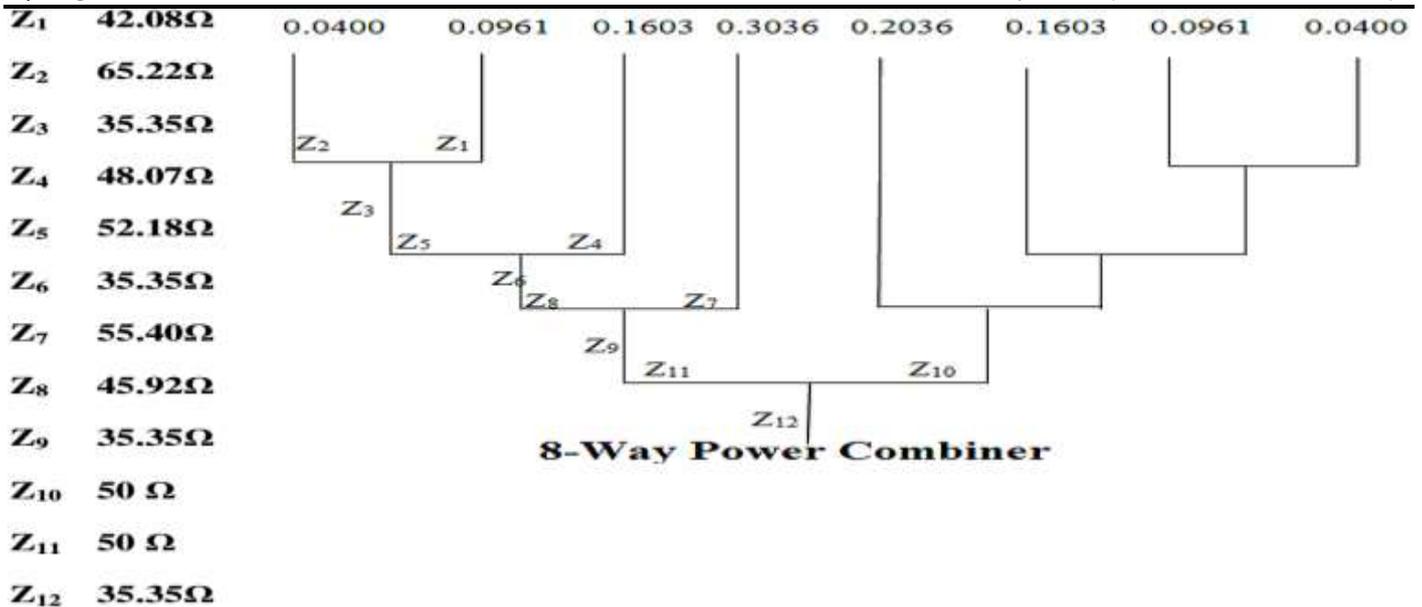


Figure 3: 8-way power combiner

3.0 Simulation Tool

Antennas are usually designed and simulated using powerful 3D electromagnetic field solvers (CEM) which are used to investigate antenna parameters performance such as: radiation pattern, directivity, gain, input impedance (matching) and 3dB beamwidth. Electromagnetic simulation (CEM) is a process of modeling the interaction of electromagnetic fields with physical objects and their environments. CEM deals with solving Maxwell's equation with numerical simulations of electromagnetic fields. These numerical simulations have predictive powers and capability of dealing with complex structures. The simulation design for antennas analytically predicts their performance, calibrate antenna systems, and estimate co-site interference of multiple antennas on a platform. There are various simulation tool which could be used for system antenna simulation. For this employed.

The array calc V2.4 [17] is a phased array design software based on MATLAB. The tool box uses a graphical method to compute the array patterns. Each array element is placed in any 3D locations and orientations using a global co-ordinate system. It then calculates the distance and direction from each element to the required points on the surface and summing the field contributions produces the radiation patterns. For effective simulation using the array calc V2.4 tool box, all elements in the array must be of the same type, the effect of mutual coupling is neglected, no voltages, current or impedance is considered and only far-field patterns are calculated.

NEC2 is a computer code for analyzing the electromagnetic response of antennas structure in free space or over a ground plane. The analysis is accomplished by the numerical solution of integral equations for induced currents. The principles of antenna analysis used by NEC2 is based on the method of moments (MoM), a mathematical technique that subdivides an antenna element into segments, calculate the current properties

and then combines the results to provide a set of results for the array of elements. The excitation could be due to an incident plane wave or voltage source on a wire. The output generally are current and charge density, electric or magnetic field in the vicinity of the structure and the radiated fields. This approach avoids many of the simplifying assumptions required by other solution methods and versatile tools for electromagnetic analysis [1].

4.0 Results and Discussion

The proposed 8-element linear phased array antenna at 2.0GHz in this study is completely modeled and simulated using array calc V2.4 and NEC2. The analysis of the performance of the antenna as determined and controlled by its geometry, coupling interaction between the array elements and the field networks using the software simulation is presented. Figure 4 shows the physical layout of the radiating element in 3D geometric coordinates. Figure 5 shows the radiation pattern of the passive radiating element, when there is no application of any form of excitation. The direction of the maximum radiation is at $\theta=0^\circ$ since there is no phase excitation. The first side lobe level (FSL) with respect to the main beam is 13dB below the main lobe and there is no grating lobe. This is quite far from the main lobe indicating low level of interference from unwanted sources, but does not correspond to the design orientation of 20dBi below main lobe. The simulated values of the -3dB beamwidth obtained is 8.52° while the directivity is 17.76dBi.

The radiation pattern obtained with the effect of amplitude excitation only gives a wide 3dB beamwidth of 23° and the side lobes are relatively large and for most applications these side lobes are undesirable. High side lobes can increase interference or result in spurious signal reception. Therefore, it is necessary to reduce the beamwidth and side lobes level by using a taper in the amplitude of the element [4, 5]. Using arraycalc simulation with the modified Taylor distribution for the amplitude

taping, the pattern in the Figure 6 is obtained. The radiation pattern indicates beamwidth of 9.10° , a value higher than the value obtained in the Figure 5, and reduction in the side lobes. It is also observed that there is a reduction in the value of the directivity from 17.76dBi to 17.19dBi . However, the FSL is 19.8dBi below the main beam which is very close to the design specification of 20dBi .

Figure 7 shows the effect of the exciting each element in the array with a different phase thereby steering the beam in a particular direction with the amplitude taper on the radiation pattern. The beamwidth increases as the array is scanned. When the beam is scanned to the angle of 15° at the resonant frequency of 2.0GHz , the entire pattern is displaced from broadside pattern. Also, the presence of grating lobes is observed. To eliminate the effect of the grating lobes, the element spacing (d) is reduced to a value of 0.5λ from 0.7λ . The Amplitude and the phase taper values were slightly changed thereby causing changes in both directivity and 3dB beamwidth. The optimized characteristics obtained for the proposed antenna are 15.4dBi Directivity, 15.0° 3dB beamwidth and FSL of 20dBi as shown in the Figure 8.

To match the 50Ω characteristics impedance (Z_0) of the antenna with the transmission line, a quarter wave impedance transformer was employed for the impedance transformation and the values were calculated using the equations (18 – 21) with the aid of NEC2. The computed input impedances plotted on the Smith chart are displayed in the Figure 9.

Figure 10 shows the plot of the simulated standing wave ratio (SWR) versus the frequency of the proposed antenna. The minimum computed SWR obtained is 1.36dB . The computed reflection coefficient and return loss are respectively 0.15 and 16.3dB , which show that the impedance bandwidth is $1575 - 2460\text{MHz}$. The complete-Integrated model comprising of array with finite ground plane and feed networks is shown in figure 4.13. The simulated value of directivity D is 15.4dBi which is 0.3 higher than the design value of 15.1dBi . Also, there is a slight variation in the 3dB beamwidth values; the simulated value is 15° while the designed value is 12.7° . This may be due to the influence of inter-element effect not being taken into consideration in the designed calculation.

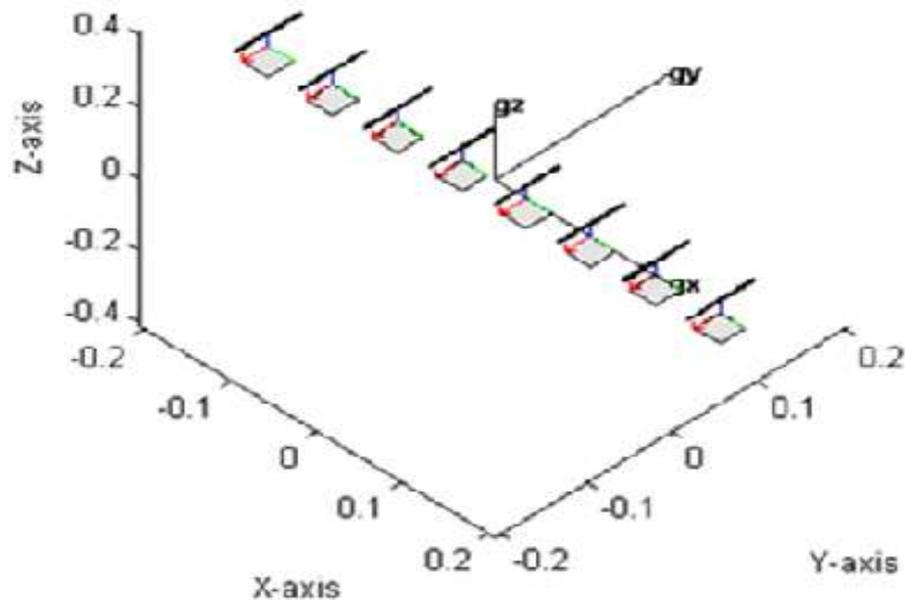


Figure 4: Showing the physical layout of the radiating elements of the proposed antenna in 3D geometric coordinates.

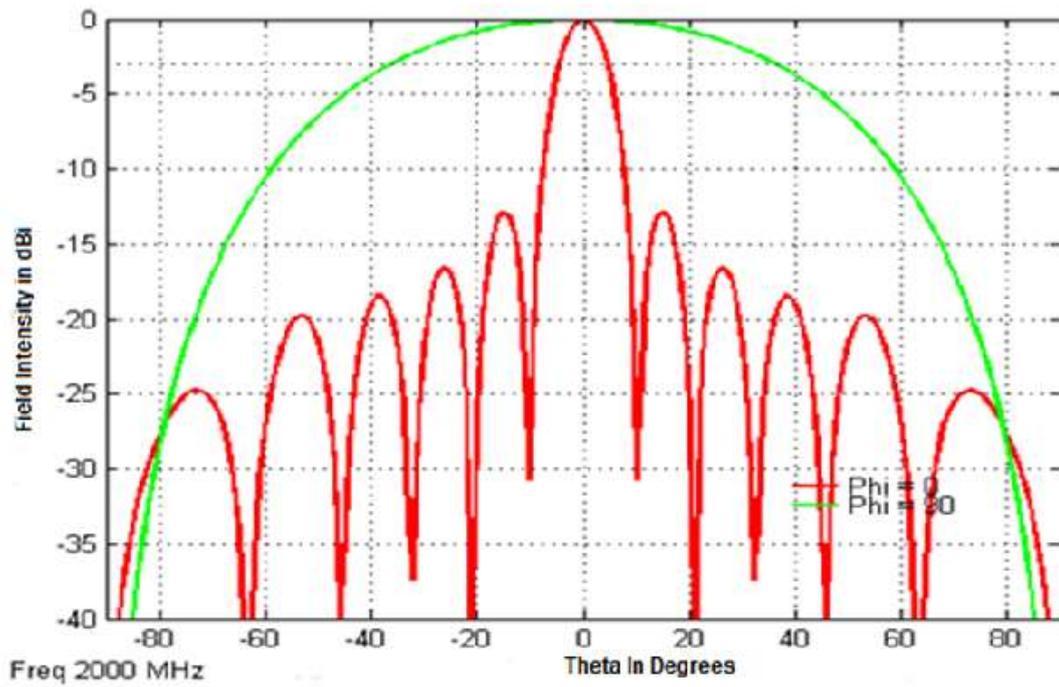


Figure 5: Showing the radiation pattern of the passive radiating element without application of any form of excitation.

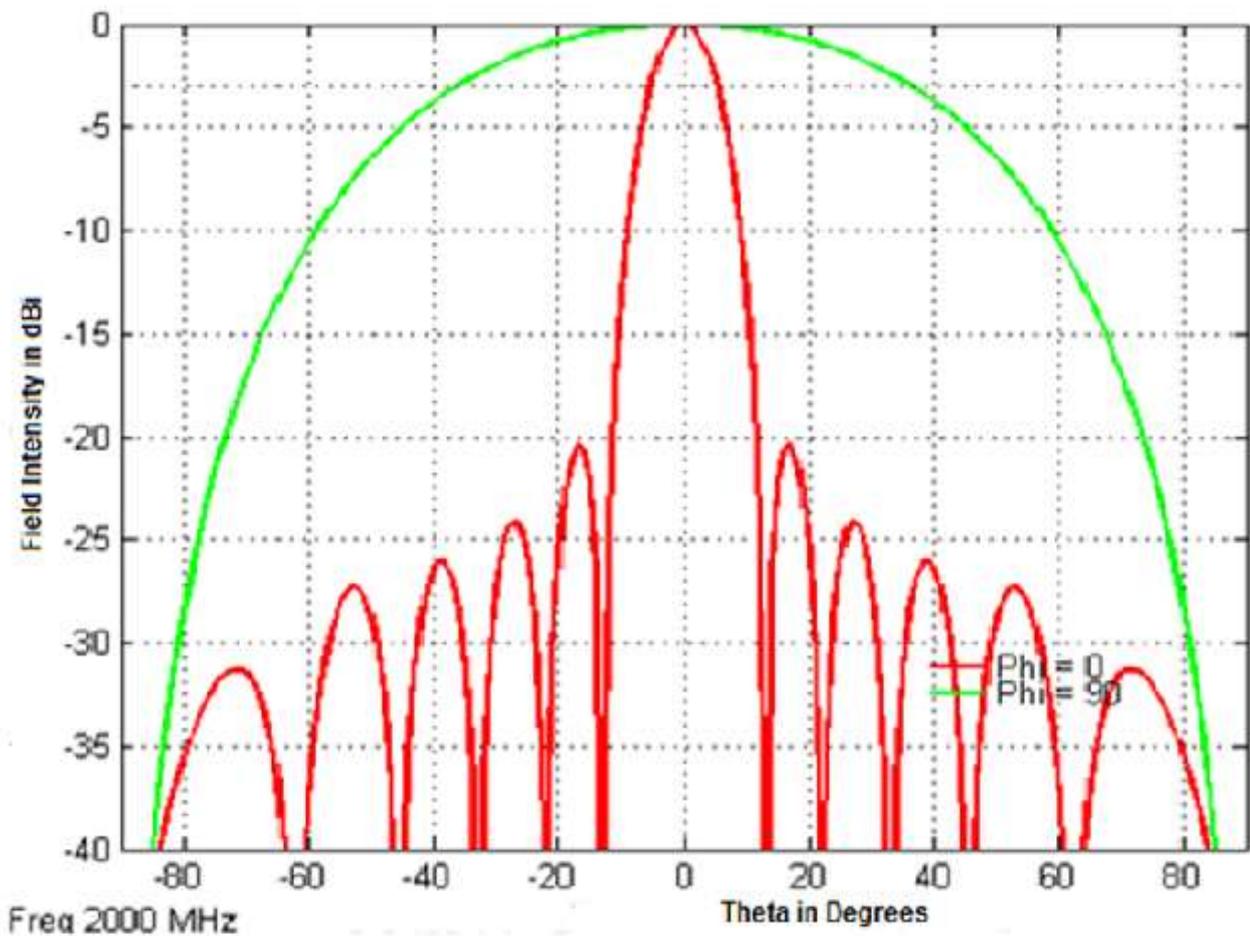


Figure 6: Showing the radiation pattern when amplitude excitation with tapping is applied.

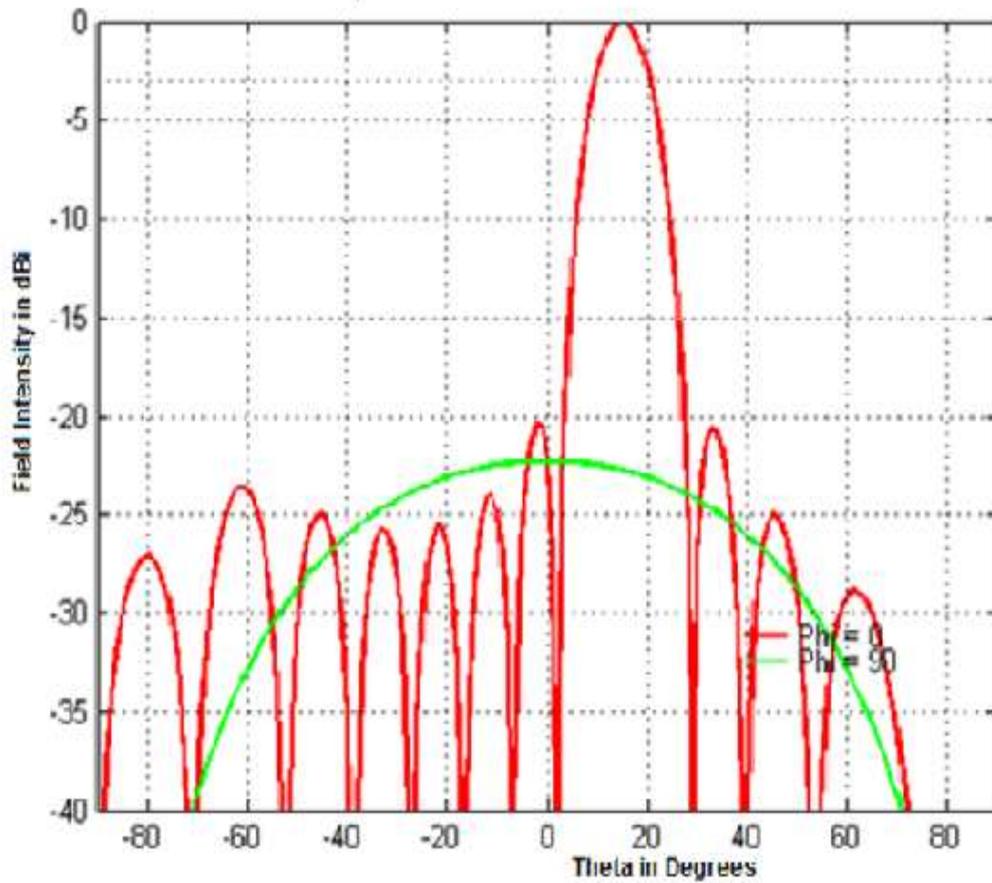


Figure 7: Showing the effect of amplitude taper excitation with a different phase on the radiation pattern.

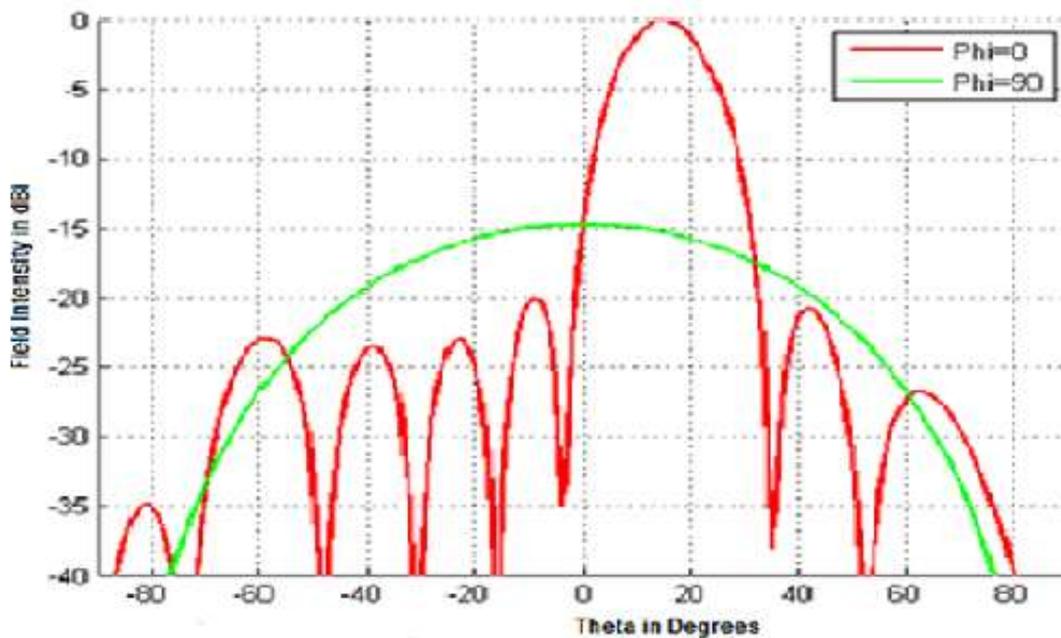


Figure 8: Showing the optimized radiation pattern obtained for the proposed antenna.

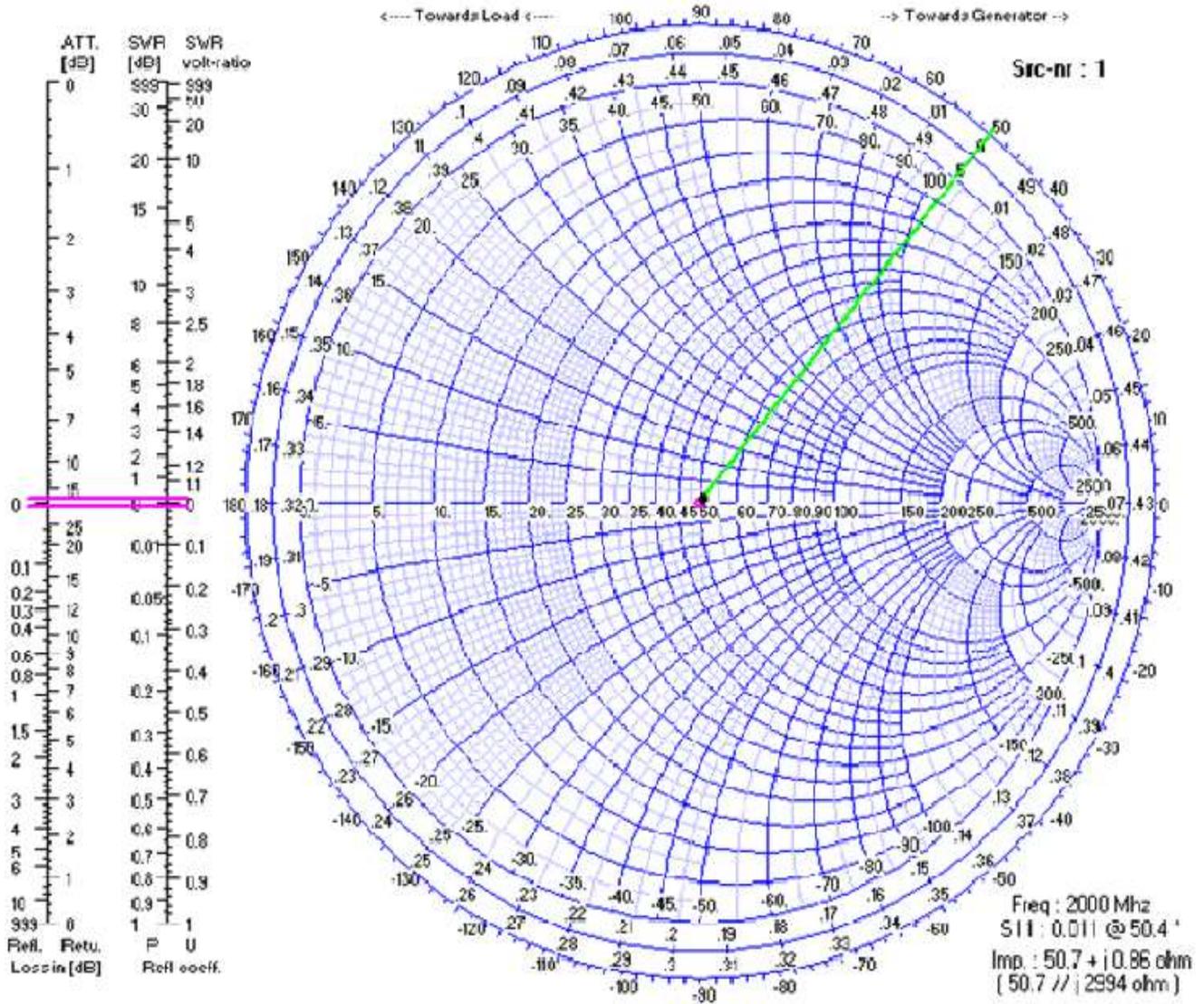


Figure 9: Showing the computed input impedances of the proposed PAA.

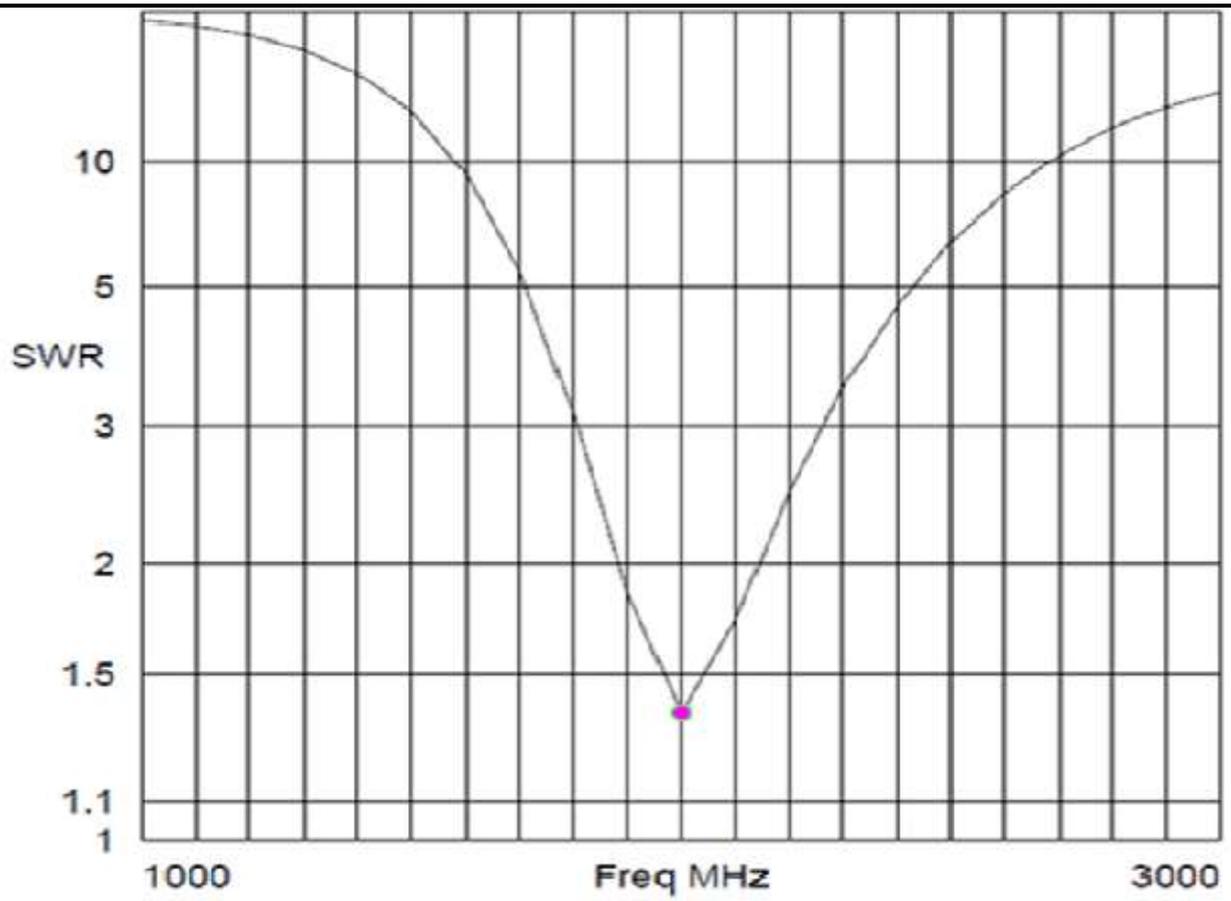


Figure 10: Showing the plot of the simulated standing wave ratio (SWR) versus the frequency (Freq) of the proposed PAA

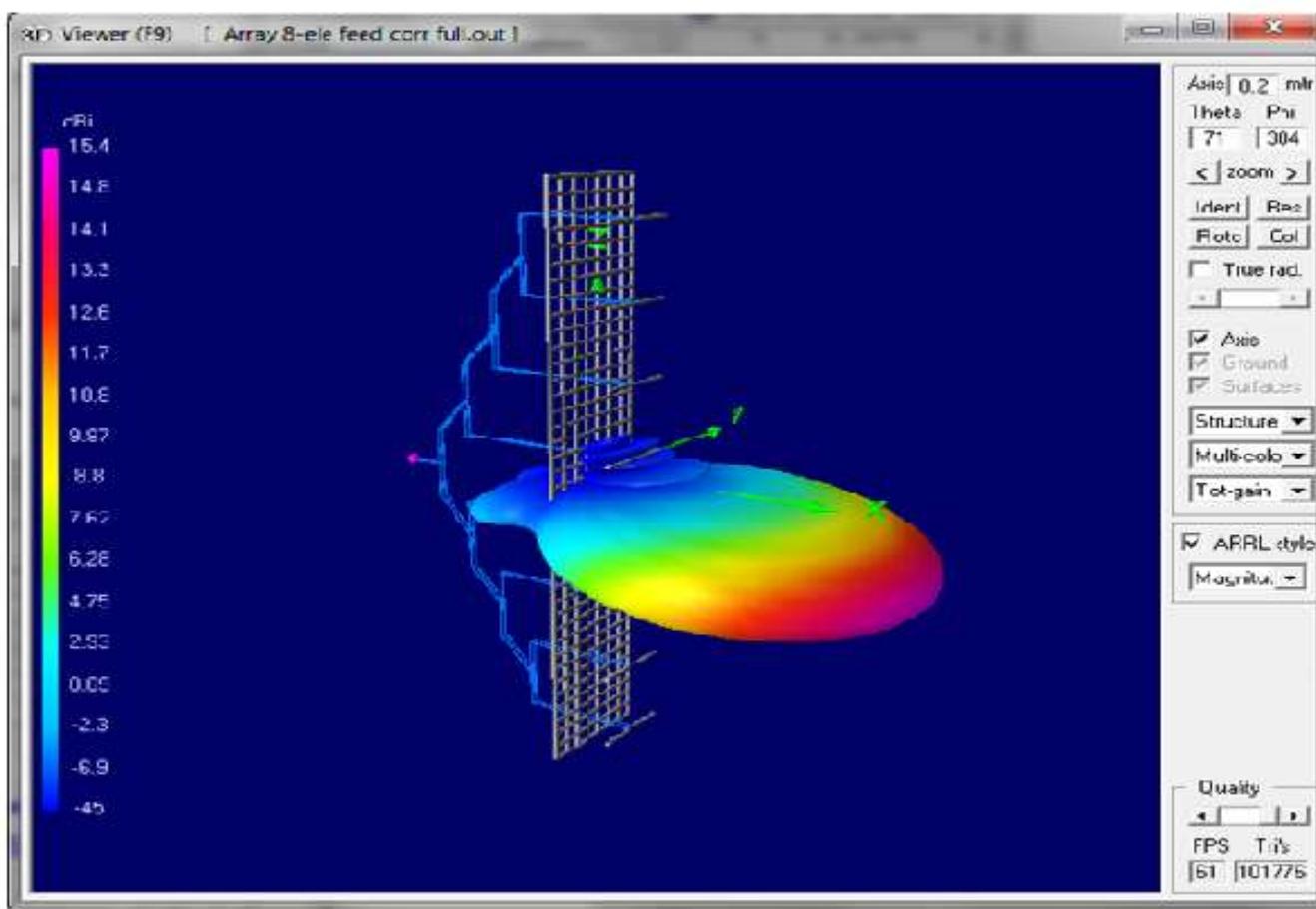


Figure 11: Showing the complete model of the proposed antenna designed in this study.

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