Radiation Patterns Account of a Circular Microstrip Antenna Loaded Two Annular

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Abstract
In this paper, theoretical study of circular microstrip antenna loaded two annular (CMSAL2AR) and calculation of the radiation pattern using principle equivalence with moment of method formulation of electromagnetic radiation in this these based on the bodies of revolution (BoR), which are generated by revolution a planar curve about an axis called axis of symmetry to solving the electric fields integral equation (EFIE) and magnetic field integral equation (MFIE). To find an unknown electric current density on the conductor surface ,and both unknowns electric and magnetic density current on the dielectric surface which are responsible for the generation of far fields radiation in the space for the components (E_θ ,E_φ) ,the surface currents was represented by a set of basis functions that give the Fourier series because the body has a circular symmetry property and then select a set of weighted functions to find a linear system by using Galerkin method which requires that the weighted functions are equal to the complex conjugate of the current $(W = J^*)$ .from radiation pattern calculated the Directive gain can be utilized to the directive gain increased to (G= 21.30 dB) when ($R_{g1} = 0.015\lambda$) for the ratio of (R_{ab}= 5.5), and bandwidth has been better (BW%= 19.9%) when ($R_{g1} = 0.01\lambda$) for the ratio (R_{ab}= 6.5) **.**

1. Introduction

The circular disc microsrtip antenna consists radiating circular patch generally made of high conductivity material such as gold, and copper. It is also the source of the radioactive radiation of the antenna on one side of a dielectric substrate, a layer of electrically dialectic material was hurt on one of its aspects patch and on the other side ground plane have a fixed base of the Electrical permittivity (ε_r) thickness (h) is much less than the length deployed in space wave (λ_O). And Ground Plane This part of the conductor of the same metal that enters the patch industry material consists for this segment a significant impact on the radiation structure of the microstrip antenna and so their impact on the areas of the edge (Fringing Fields) that are the basis of the radiation from the microstrip antenna, when increasing the dimensions this area of the base to a certain extent leads to increased transmitted power [1].

As for its disadvantages, microstrip antennas are inefficient and possess very narrow frequency bandwidth and low efficiency [2], there are methods of significantly reducing the effect of some of the above-mentioned, since then, researchers have devoted their studies to find new approaches, For example, Directive gain and bandwidth of microstrip antennas can be improved by increasing the height of the substrate, such changes shapes of microstrip patch elements and using loaded between two or more patches as in propose antenna circular microstrip antenna loaded two annular ring as shown figure (1).The analytic method for used in the antenna propose is principle equivalence to calculated the basis function and using Galerkin model for BOR to convert the integral equation to set of inner matrix solving by using moment method [3].

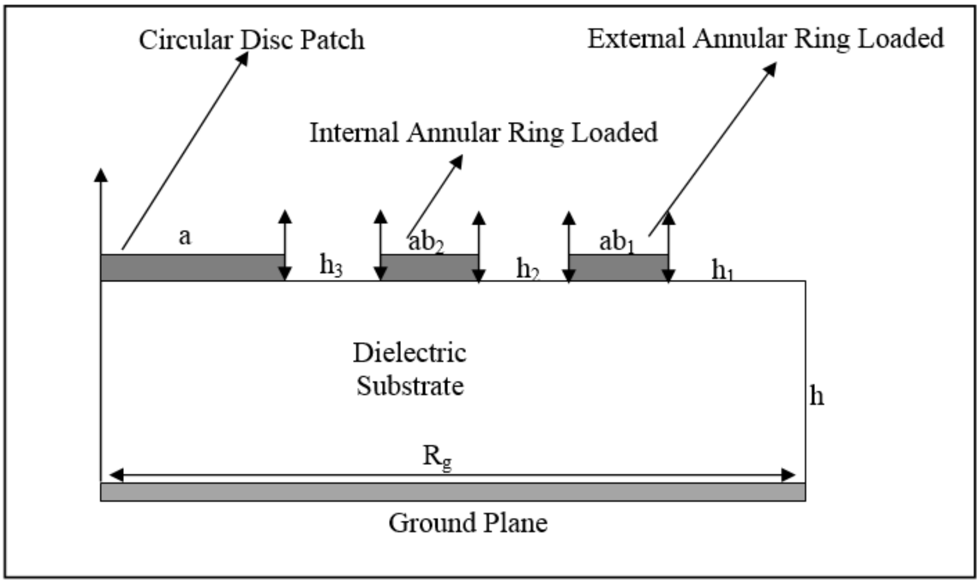

Figure (1) Design of Dimensions CMSL2AR

In Figure (1), we note different dimensions of the loading radiated area of the antenna as well as the dielectric zones which can be shortened by certain percentages to include all the components of the various important antenna parts in the formation of the radioactive structure process, where the value (R_{ab}) represents the ratio between the radius of area of the radiation loaded (b) to the radius the annular disk (a) which has a large and fundamental importance in improving the antenna coefficients because it represents the controlling of the area of the electromagnetic coupling between the external and internal rings and the annular disk, as well as the sum of the two rings space (ab) through which we control the area of the external and internal rings, and the total insulating layer on the outer surface of the antenna (h_{123}) through which the dimensions of the insulating zones can separately be reduced or increased.

2. Mathematical Analysis

The theoretical studies to solving the electromagnetic problem issues previously is based on Method of moment is used here to solve the electromagnetic problem the field components is distributed on the surfaces of a CMSAL2AR.

The high calculations ability of this method in solving the integral equation, since it included all the boundary conditions of the current densities on different surface and the easiest in suggest the testing functions as the conjugate of the basis function, according to Galerkin's model. Made it as the most important numerical techniques in addition the helped of the body of revolution principle and the application of equivalence principle. Now we will represent the electric and magnetic fields in these the volume V by the electric and magnetic equivalent surface currents, these currents are defined as[4] :

$$\bar{J}_S = \hat{n} \times \overline{H}_S \qquad \text{……..(1)}$$

$$\overline{M}_S = -\hat{n} \times \overline{E}_S \qquad \text{……..(2)}$$

Where $\vec{J}_S, \vec{M}_S$ represents electric and magnetic the surface currents densities respectively.

According to the equivalence, principle will be divided as two regions. There regions are the a finite region of volume V^e and a finite region of volume V^d as shown in figure (2) and (3) .

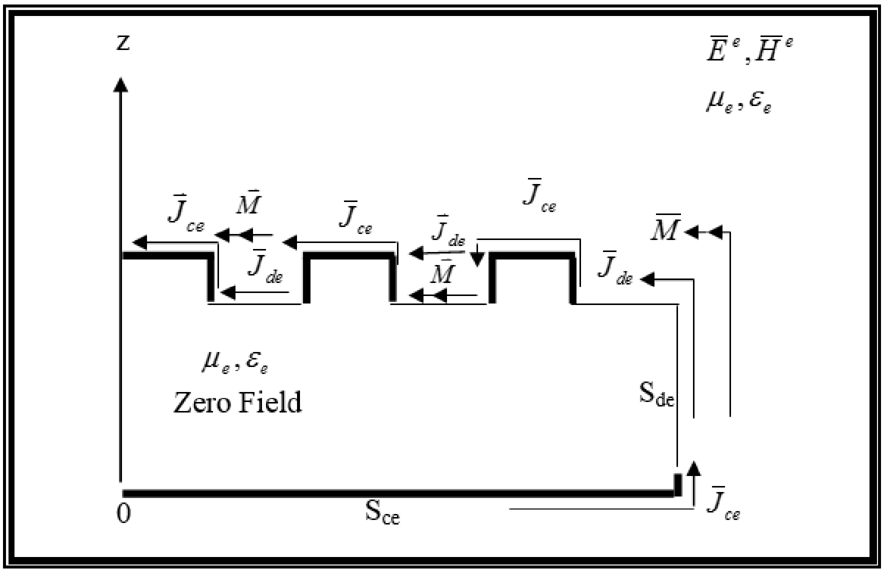

Figure (2): Equivalent for region V^e.

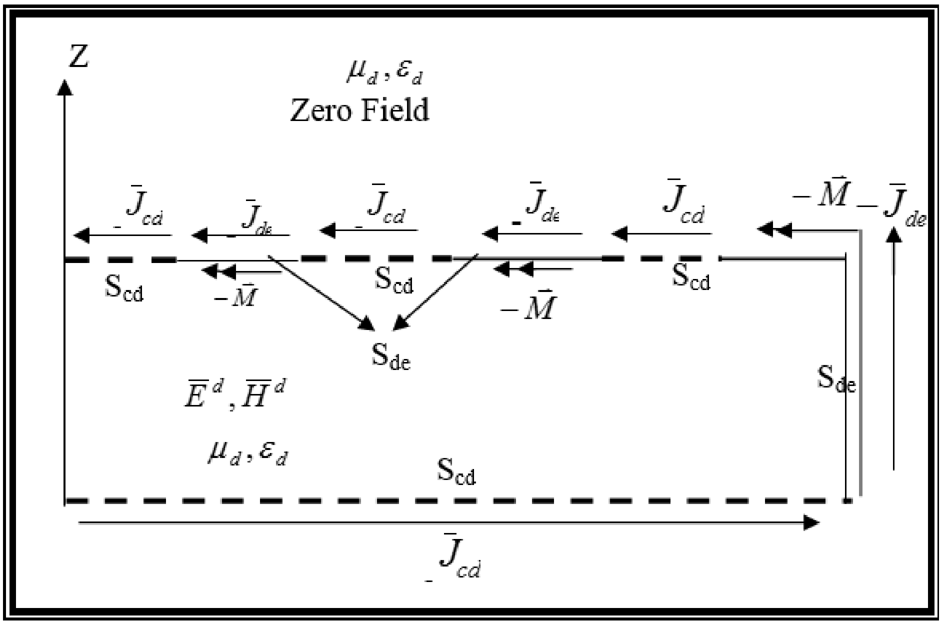

Figure (3): Equivalent for region V^d.

The microstrip antenna under test consists of a conductive material such as the patch, the ground plane, and two annular ring as well as the dielectric substrate between them. Because of these deferent surfaces, two type of boundary condition must be satisfied. These conditions requires vanishing of the tangential electric field component on the conductor surface, while the magnetic and electric fields continuity on the dielectric surface. The polarized currents within the dielectric surface and the current density on the conductor surface IE, therefore the boundary condition equation for antenna can describe as follows[5,6]:

$$\hat{n}\times\overline{E}^e = 0 \qquad \text{on } S_{ce}\ldots\ldots(3\text{-a})$$

$$\hat{n}\times\overline{E}^d = 0 \qquad \text{on } S_{cd}\ldots\ldots(3\text{-b})$$

$$\hat{n}\times\overline{E}^d = \hat{n}\times\overline{E}^e \qquad \text{on } S_{de}\ldots\ldots(3\text{-c})$$

$$\hat{n}\times\overline{H}^d = \hat{n}\times\overline{H}^e \qquad \text{on } S_{de}\ldots\ldots(3\text{-d})$$

Where $\hat{n}$ is the unit vector normal to the conductor and dielectric surface.

The surface equivalent electric and magnetic currents are:

$$\overline{J}_{ce} = \hat{n}\times\overline{H}^e \qquad \text{on } S_{ce}\ldots\ldots(4\text{-a})$$

$\overline{J}_{cd} = \hat{n}\times\overline{H}^{d}$ on S_{cd}…..(4-b)

$\overline{J}_{de} = \hat{n}\times\overline{H}^{e}$ on S_{de}…..(4-c)

$\overline{M} = -\hat{n}\times\overline{E}^{e}$ on S_{de}…..(4-d)

The equivalent electric currents ($\vec{J}_{ce}, \vec{J}_{cd}, \vec{J}_{de}$) are generated on the conductor and dielectric surface where the equivalent magnetic currents $\vec{M}$ is generated on the dielectric surface only. Applying the equivalent principle on internal and external equivalent region of the problem yields the integral equations as follows:

$$\hat{n}\times\overline{E}^{e}(\overline{J}_{ce}+\overline{J}_{de},\overline{M}) = 0 \quad \text{on } S_{ce} \text{ and } S_{de} \quad \text{.....(5-a)}$$

$$\hat{n}\times\overline{H}^{e}(\overline{J}_{ce}+\overline{J}_{de},\overline{M}) = 0 \quad \text{.....(5-b)}$$

$$\hat{n}\times\overline{E}^{d}(-\overline{J}_{cd}-\overline{J}_{de},-\overline{M}) + \hat{n}\times\overline{E}^{d}(\overline{J}^{id},0) = 0 \quad S_{cd} \text{ and } S_{de} \quad \text{.....(5-c)}$$

$$\hat{n}\times\overline{H}^{d}(-\overline{J}_{cd}-\overline{J}_{de},-\overline{M}) + \hat{n}\times\overline{H}^{d}(\overline{J}^{id},0) = 0 \quad \text{.....(5-d)}$$

Where $\vec{H}^{a}(\vec{J},\vec{M})$ and $\vec{E}^{a}(\vec{J},\vec{M})$ represent the magnetic and electric fields due to the currents $\vec{J}$ and $\vec{M}$, radiated in media characterized by (μ_a, ε_a) .The simple (a) represent to radiation media characterized by (μ_e, ε_e) and (μ_d, ε_d) , while $\vec{H}^{d}(\vec{J}^{id},0)$ and $\vec{E}^{d}(\vec{J}^{id},0)$ represents the magnetic and electric fields due to the currents of the feed $(\vec{J}^{id})$.

Reformulate the equation (4) for the equivalent surfaces (S_{cd}, S_{ce}, S_{de}) gives:

$$\left(\overline{J}_{ce}+\overline{J}_{de}\right) = \sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)} I_{ni}^{1e}\overline{J}_{ni}^{1e} + \sum_{i=1}^{2(N_d-4)} I_{ni}\overline{J}_{ni}^{2e}\right] \quad \text{......(6-a)}$$

$$\left(\overline{J}_{cd}+\overline{J}_{de}\right) = \sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)} I_{ni}^{1d}\overline{J}_{ni}^{1d} + \sum_{i=1}^{2(N_d-4)} I_{ni}\overline{J}_{ni}^{2d}\right] \quad \text{......(6-b)}$$

$$\overline{M} = \eta_e \sum_{n=-\infty}^{\infty}\sum_{i=1}^{2(N_d-4)} K_{ni}\overline{M}_{ni} \quad \text{......(6-c)}$$

Substitute equation (6) in to equation (5) to get:

$$\sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)} I_{ni}^{1e}\overline{E}_{\tan}^{e}\left(\overline{J}_{ni}^{1e},0\right) + \sum_{i=1}^{2(N_d-4)} I_{ni}\overline{E}_{\tan}^{e}\left(\overline{J}_{ni}^{2e},0\right) + \eta_e \sum_{i=1}^{2(N_d-4)} K_{ni}\overline{E}_{\tan}^{e}\left(0,\overline{M}_{ni}\right)\right] = 0$$

,on S_{ce} ......(7-a)

$$\sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)} I_{ni}^{1d}\overline{E}_{\tan}^{d}\left(\overline{J}_{ni}^{1d},0\right) + \sum_{i=1}^{2(N_d-4)} I_{ni}\overline{E}_{\tan}^{d}\left(\overline{J}_{ni}^{2d},0\right) + \eta_e \sum_{i=1}^{2(N_d-4)} K_{ni}\overline{E}_{\tan}^{d}\left(0,\overline{M}_{ni}\right)\right] = \overline{E}_{\tan}^{d}\left(\overline{J}^{id}\right)$$

,on S_{cd} ......(7-b)

$$\sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)}\left\{I_{ni}^{1e}\overline{E}_{\tan}^{e}\left(\overline{J}_{ni}^{1e},0\right) + I_{ni}^{1d}\overline{E}_{\tan}^{d}\left(\overline{J}_{ni}^{1d},0\right)\right\} + \sum_{i=1}^{2(N_d-4)}\left\{I_{ni}\overline{E}_{\tan}^{e}\left(\overline{J}_{ni}^{2e},0\right) + I_{ni}\overline{E}_{\tan}^{d}\left(\overline{J}_{ni}^{2d},0\right)\right\} + \eta_e \sum_{i=1}^{2(N_d-4)}\left\{K_{ni}\overline{E}_{\tan}^{e}\left(0,\overline{M}_{ni}\right) + K_{ni}\overline{E}_{\tan}^{d}\left(0,\overline{M}_{ni}\right)\right\}\right] = \overline{E}_{\tan}^{d}\left(\overline{J}^{id}\right)$$

, on S_{de} ......(7-c)

$$\sum_{n=-\infty}^{\infty}\left[\sum_{i=1}^{2(N-N_d-10)}\left\{I_{ni}^{1e}\overline{H}_{\tan}^{e}\left(\overline{J}_{ni}^{1e},0\right) + I_{ni}^{1d}\overline{H}_{\tan}^{d}\left(\overline{J}_{ni}^{1d},0\right)\right\} + \sum_{i=1}^{2(N_d-4)}\left\{I_{ni}\overline{H}_{\tan}^{e}\left(\overline{J}_{ni}^{2e},0\right) + I_{ni}\overline{H}_{\tan}^{d}\left(\overline{J}_{ni}^{2d},0\right)\right\} + \eta_e \sum_{i=1}^{2(N_d-4)}\left\{K_{ni}\overline{H}_{\tan}^{e}\left(0,\overline{M}_{ni}\right) + K_{ni}\overline{H}_{\tan}^{d}\left(0,\overline{M}_{ni}\right)\right\}\right] = \overline{H}_{\tan}^{d}\left(\overline{J}^{id}\right)$$

, on S_{de} ......(7-d)

In this proposed antenna using a Galerkin's method which is one of the most appropriate calculation methods for the selection of the weight functions ($W = J^*$) is used[7].

$$\overline{W}(\overline{r}) = \overline{W}^t(t,\phi) + \overline{W}^\phi(t,\phi) = \sum_{m=-\infty}^{\infty}\sum_{i=1}^{N-14}\left[\overline{W}_{mi}^t(t,\phi) + \overline{W}_{mi}^\phi(t,\phi)\right] \quad \text{……(8-a)}$$

$$\overline{W}_{mi}^t(t,\phi) = \hat{u}_t f_i(t)\ e^{-jm\phi} \quad \text{……(8-b)}$$

$$\overline{W}_{mi}^\phi(t,\phi) = \hat{u}_\phi f_i(t)\ e^{-jm\phi} \quad \text{……(8-d)}$$

And using inner products the weight of functions with equations (7) can be represent as the following matrix:

$$\begin{bmatrix} \left[Z_{ce,ce}^{1e}\right]_n & [0]_n & \left[Z_{ce,de}^{2e}\right]_n & \eta_e\left[Y_{ce,de}^{3e}\right]_n \\ [0]_n & \left[Z_{cd,cd}^{1d}\right]_n & \left[Z_{cd,de}^{2d}\right]_n & \eta_e\left[Y_{cd,de}^{3d}\right]_n \\ \left[Z_{de,ce}^{1e}\right]_n & \left[Z_{de,cd}^{1d}\right]_n & \left[Z_{de,de}^{2e} + Z_{de,de}^{2d}\right]_n & \eta_e\left[Y_{de,de}^{3e} + Y_{de,de}^{3d}\right]_n \\ \left[Y_{de,ce}^{1e}\right]_n & \left[Y_{de,cd}^{1d}\right]_n & \left[Y_{de,de}^{2e} + Y_{de,de}^{2d}\right]_n & \eta_e\left[Z_{de,de}^{3e} + Z_{de,de}^{3d}\right]_n \end{bmatrix} \begin{bmatrix} \left[I^{1e}\right]_n \\ \left[I^{1d}\right]_n \\ [I\ ]_n \\ [K\ ]_n \end{bmatrix} = \begin{bmatrix} [0]_n \\ \left[V_{cd}^d\right]_n \\ \left[V_{de}^d\right]_n \\ \left[I_{de}^d\right]_n \end{bmatrix}$$

……….(9)

and can be writing as:

$$[\vec{T}_{nm}][\vec{I}_n] = [\vec{V}_n] \quad \text{…… (10)}$$

Where $[\vec{T}_n]$ a generalized impedance and admittance matrices matrix is, $[\vec{I}_n]$ is a column matrix containing the unknown expansion coefficients, and $[\vec{V}]$ is the excitation column matrix [8].

The radiation field are calculated at the far field region (E_θ, E_ϕ) from induced electric and magnetic currents flowing on the surface of the conductor and the dielectric of the antennas so [9]:

$$E_\theta = -\frac{jw\mu_e}{4\pi\ r_o} e^{-jK_e r_o} F_1(\theta_o, \phi_o) \quad \text{…… (11-a)}$$

$$E_\phi = -\frac{jw\mu_e}{4\pi r_o} e^{-jK_e r_o} F_2(\theta_o, \phi_o) \quad \text{…… (11-a)}$$

$F_1(\theta_o, \phi_o)$ and $F_2(\theta_o, \phi_o)$ are the measurement coefficients given by :

$$F_1(\theta_o, \phi_o) = \int_s \left(\overline{J}(\overline{r}')\cdot\hat{\theta} + \frac{1}{\eta_e}\overline{M}(\overline{r}')\cdot\hat{\phi} \right) e^{-jK_e \hat{r}_o \bullet \overline{r}'} ds \quad \text{…… (12-a)}$$

$$F_2(\theta_o, \phi_o) = \int_s \left(\overline{J}(\overline{r}')\cdot\hat{\phi} - \frac{1}{\eta_e}\overline{M}(\overline{r}')\cdot\hat{\theta} \right) e^{-jK_e \hat{r}_o \bullet \overline{r}'} ds \quad \text{…… (12-b)}$$

where S is the exterior surface of the antenna, $\hat{r}_\circ$ a unit vector directed from the origin to the field point, and $\hat{\theta}_\circ, \hat{\phi}_\circ$ are the transverse unit vector in the direction of increasing θ and ϕ, respectively.

3. Results and Discussions

In this section we'll study the radiation patterns for circular microstrip antenna loaded two annular ring (CMSAL2AR) using the application of the principle equivalent to induce electric and magnetic currents densities on different surfaces of the propose antenna, and solving the integral equations by using method of moment, which is numerical method possesses the ability of high mathematical calculation. With the help of Galerkin's method and the symmetric characteristic of the BoR, and using fortran90 was the language using here for programming's the analysis of the proposed antenna and the plotting of the result. And the calculation antenna coefficients as the directive gain and bandwidth by determining better ratio (a/b) which is (R_{ab}= 5.5,6.5) as shown from the results previously, and knowing the effect of different dimensions on the radiation patterns for propose antenna, which increase the efficiency of the Antenna coefficients.

Figures (4) show the effect of different ground plane radii (R_g) on the radiation pattern and directive gain of a proposed antenna is predicated in Table (1) which represents the antenna coefficients at those dimensions, as we note that the best value for the directive gain increasing to (G= 19.92 dB) as the ground plane radius $R_g = 0.31\,\lambda$ for the ratio (R_{ab}= 5.5) .To check this, bandwidth can be also calculated for the highest values of the directive gain in the Table (1),It is shown from this table that the greater value of the bandwidth is (BW%= 21.41) will be at $R_g = 0.32\,\lambda$, compared with the ratio (R_{ab}= 5.5) is the greater value (G= 14.36 dB) and (BW%= 19.9%).And study the effect of variant in the thickness of ground plane (R_{g1}) the directive gain increased to (G= 21.30 dB) when ($R_{g1} = 0.015\lambda$) for the ratio of (R_{ab}= 5.5), and bandwidth has been better (BW%= 19.9%) when ($R_{g1} = 0.01\lambda$) for the ratio (R_{ab}= 6.5), as shown in Figure (5) and the table (2).shown in Fig.(6),and table (3), the patch and two annular ring thickness not effected on the antenna coefficients for the ratio of (R_{ab}=6.5), but directive gain increasing to (G= 21.14 dB) when ($a_1 = 0.008\lambda$) at the ratio (R_{ab}= 5.5).As well as not effect for the antenna coefficients when change the dielectric subtract radius in the loading patch area for determine dimensions, shown in Figure (7) and the table (3).

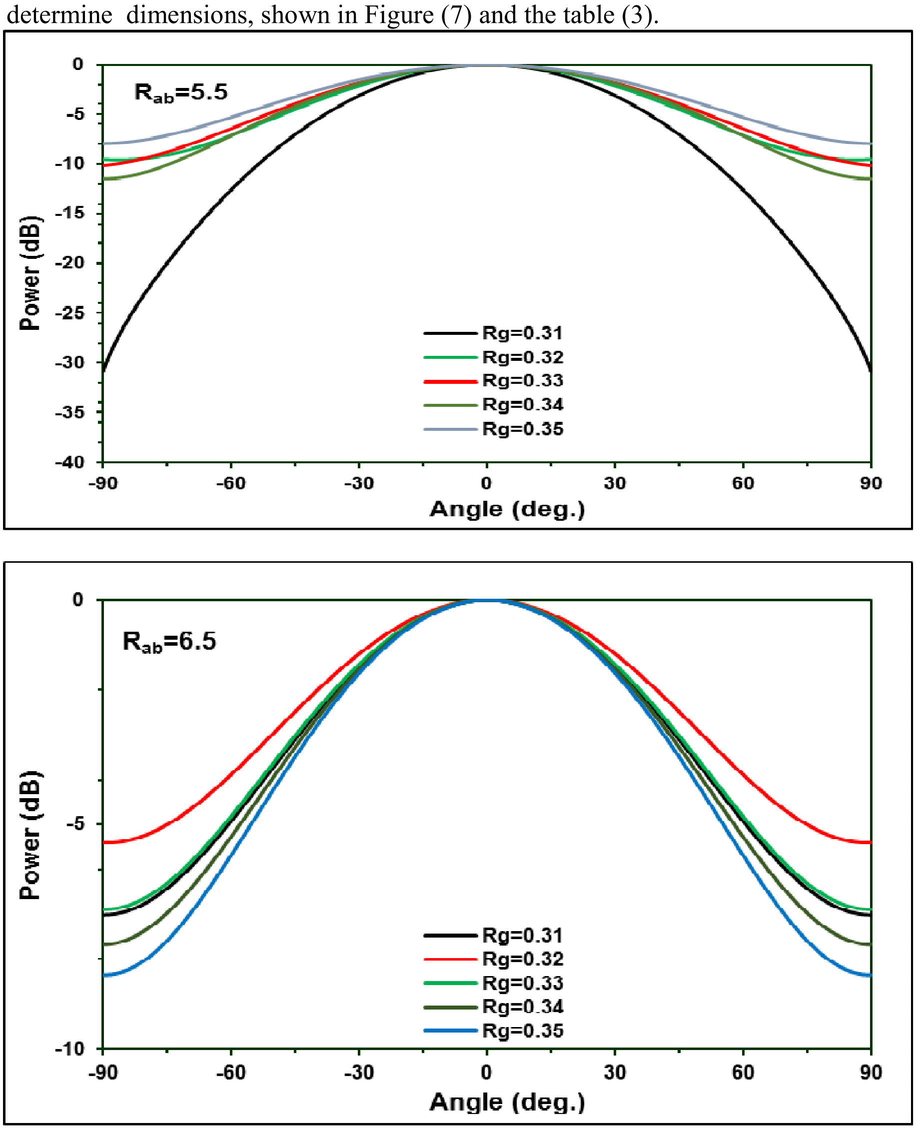

Fig. (5): The effect of ground plane radius radiation patterns of a CMSAL2A excited by TM_{11}-mode, for E-plane.

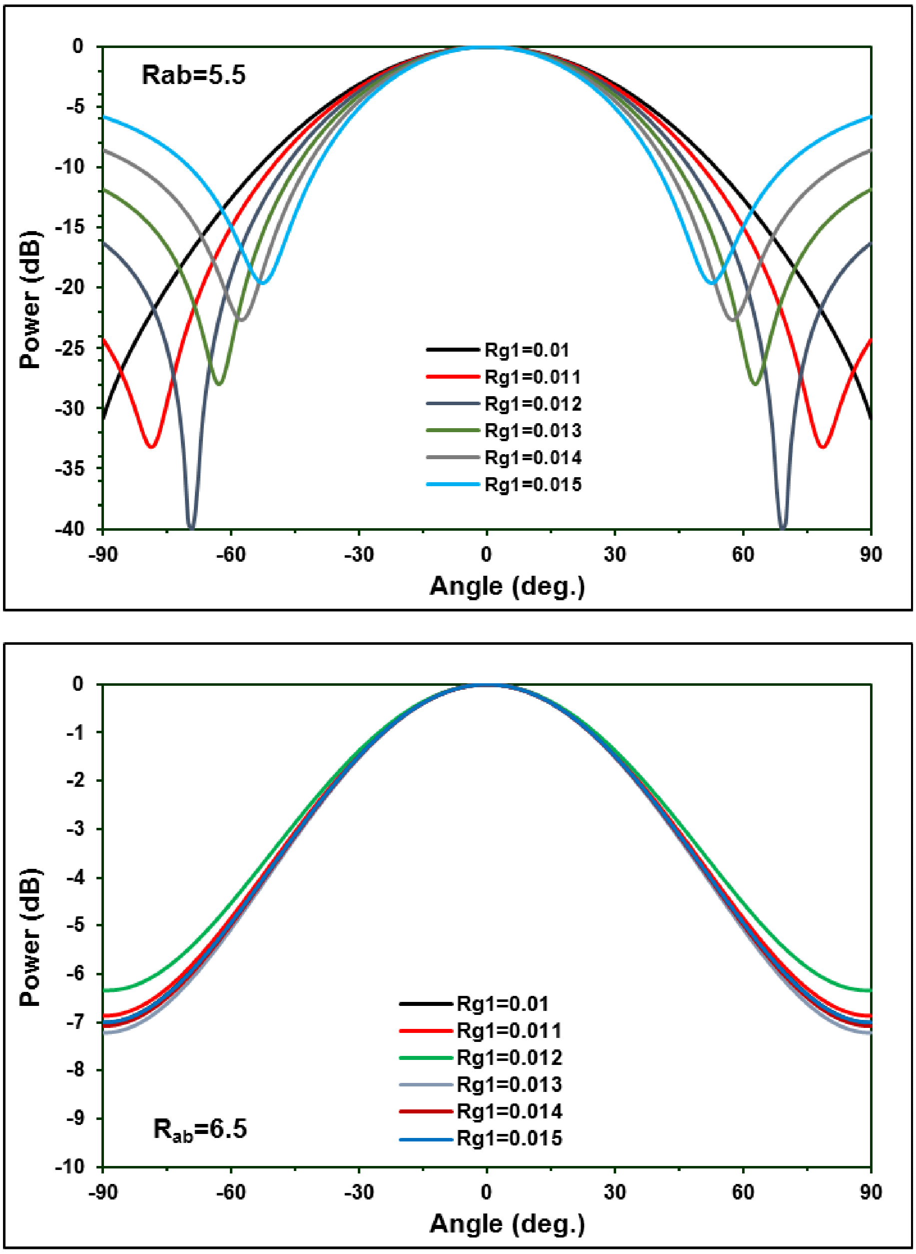

Fig. (6): The effect of ground plane thickness radiation patterns of a CMSAL2A excited by TM_{11}-mode, for E-plane.

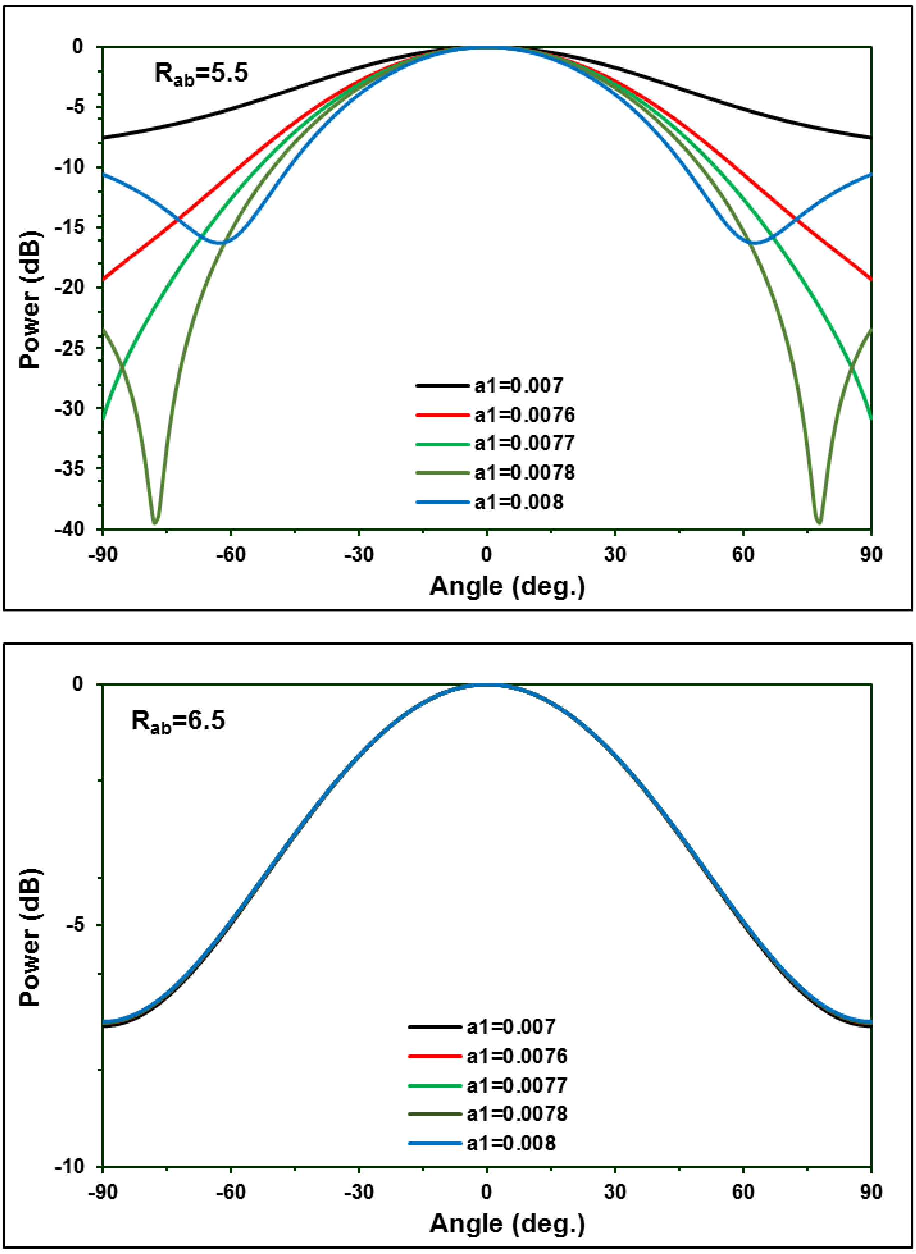

Fig. (7): The effect of conductor thickness (disc and two annular ring) tradiation patterns of a CMSAL2A excited by TM_{11}-mode, for E-plane.

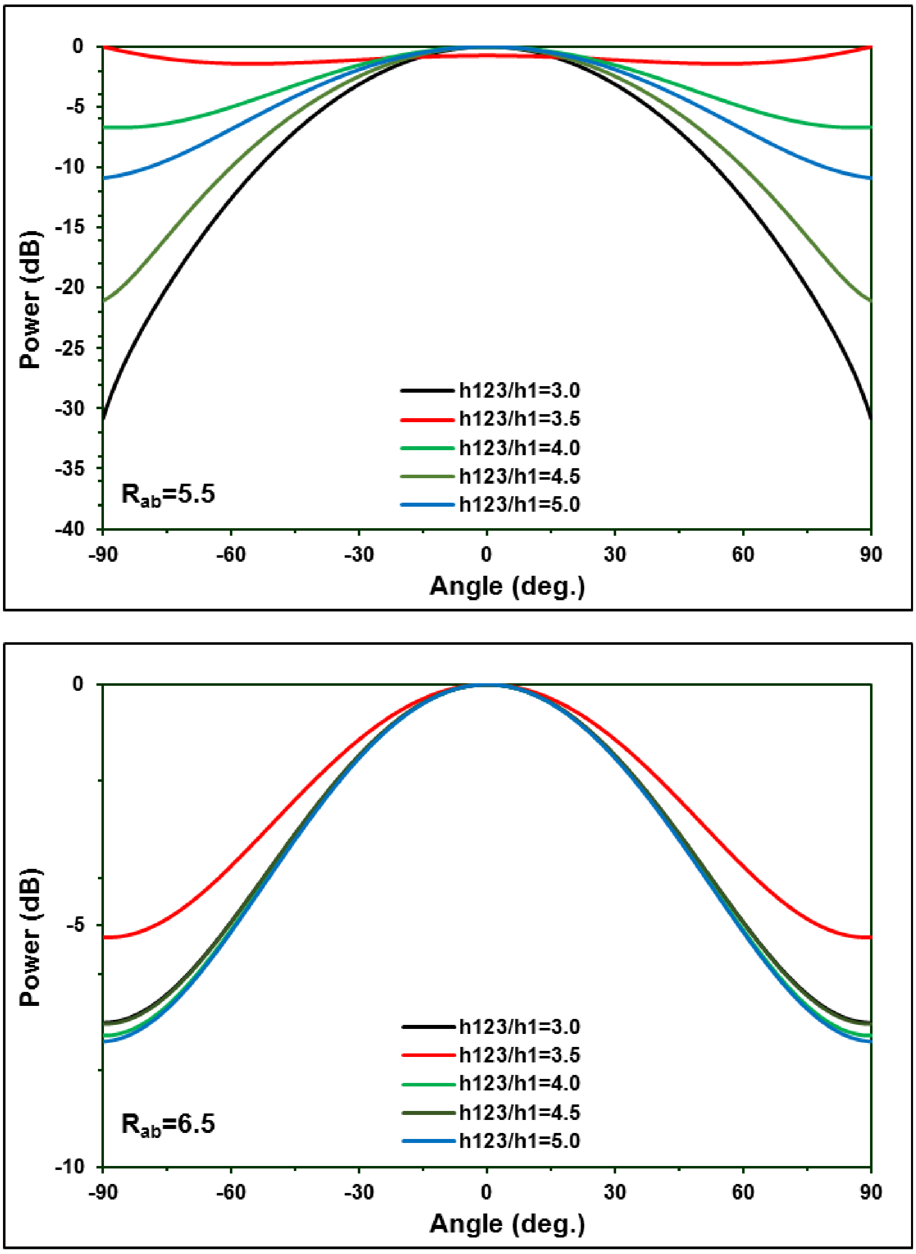

Fig. (8): The effect of dielectric subtract radius in the loading patch area for radiation patterns of a propose antenna excited by TM_{11}-mode, for E-plane.

Table (1): The directive gain and bandwidth under effect of ground plane radius for ratios (R_{ab}=5.5,6.5).

Antenna Type	*$R_g(\lambda)$*	*HP_H^o (deg)*	*HP_E^o (deg)*	*Directive gain (dB)*	*BW% S=2*
CMSAL2AR (R_{ab}=5.5)	**0.31**	**58**	**64**	**19.92**	**15**
	0.32	**72**	**68**	**17.52**	**21.41**
	0.33	**78**	**68**	**16.83**	**5.377**
	0.34	**76**	**66**	**17.30**	**13.01**
	0.35	**86**	**72**	**15.48**	**12.92**
CMSAL2AR (R_{ab}=6.5)	**0.31**	**88**	**80**	**14.36**	**19.9**
	0.32	**102**	**94**	**11.68**	**13.27**
	0.33	**90**	**86**	**13.54**	**15.09**
	0.34	**86**	**92**	**13.35**	**10.27**
	0.35	**82**	**100**	**13.04**	**17.81**

Table (2): The directive gain and bandwidth under effect of ground plane thickness for ratios (R_{ab}=5.5,6.5).

Antenna Type	*$R_{g1}(\lambda)$*	*HP_H^o (deg)*	*HP_E^o (deg)*	*Directive gain (dB)*	*BW% S=2*
CMSAL2AR (R_{ab}=5.5)	0.01	58	64	19.92	15
	0.011	56	66	19.96	15.56
	0.012	54	66	20.27	16.22
	0.013	52	66	20.60	12.45
	0.014	50	66	20.94	16.13
	0.015	48	66	21.30	14.15
CMSAL2AR (R_{ab}=6.5)	0.01	88	80	14.36	19.9
	0.011	90	78	14.34	19.81
	0.012	92	70	15.13	19.27
	0.013	88	82	14.15	18.36
	0.014	88	82	14.15	19.63
	0.015	88	80	14.15	18.72

Table (3): The directive gain and bandwidth under effect of conductor thickness (disc and two annular ring)for ratios (R_{ab}=5.5,6.5).

Antenna Type	*$a_1(\lambda)$*	*HP_H^o (deg)*	*HP_E^o (deg)*	*Directive gain (dB)*	*BW% S=2*
CMSAL2AR (R_{ab}=5.5)	0.07	82	68	16.39	15.66
	0.076	62	66	19.07	16.03
	0.077	58	64	19.92	15
	0.078	56	64	20.22	13.58
	0.08	52	62	21.14	10.94
CMSAL2AR (R_{ab}=6.5)	0.07	88	80	14.36	15.9
	0.076	88	80	14.36	19.09
	0.077	88	80	14.36	19.9
	0.078	88	80	14.36	19.45
	0.08	88	80	14.36	17.08

Table (4): The directive gain and bandwidth under effect of dielectric subtract radius in the loading patch area for ratios (R_{ab}=5.5,6.5).

Antenna Type	*h_{123}/h1)(*	*HP_H^o (deg)*	*HP_E^o (deg)*	*Directive gain (dB)*	*BW% S=2*
CMSAL2AR (R_{ab}=5.5)	3.0	58	64	19.92	15
	3.5	.....	72	…..	8.2
	4.0	88	70	15.52	13.2
	4.5	66	128	12.78	6.21
	5	76	74	16.31	14.9
CMSAL2AR (R_{ab}=6.5)	4	88	80	14.36	19.9
	3.5	104	68	14.32	18.18
	4.0	86	78	14.78	15.09
	4.5	88	76	14.81	13.81
	5	86	72	15.48	14

4.Conclusions

The mathematical analysis to solve the electromagnetic problem the field components of CMSAL2AR is MoM because included all the boundary conditions of the current densities on different surface, the result indicate is increase the bandwidth and directive gain for their applications in communications and microwave.

5.References

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