A Wide Band Rectangular Dielectric Resonator Antenna for S-Band Applications

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Abstract

This paper presents a wide band rectangular dielectric resonator antenna (DRA) with microstrip slot excitation valid for S frequency band applications. The resonant frequency and radiation Q-factor of the lowest order "magnetic dipole" mode are obtained using dielectric waveguide model with magnetic walls. The accuracy of the model in predicting the resonant frequency is verified by comparing the obtained results using FDTD numerical technique, with those of a ready made software based on finite element method. The paper also includes comparison between DRA and microstrip patch antenna, MPA, designed for the same range of frequency, and using the same feeding system which is aperture coupling microstrip line

1. Introduction

Dielectric resonators (DR's) offer attractive feature as antenna elements [1].Since DR first introduced by Long et al. [2] in 1983, the dielectric resonator antenna (DRA) has received increasing attention in the last two decades. The DRA has many advantages such as small size, low cost, low loss, light weight, and ease of excitation. It also has an advantage over the microstrip antenna in that it has a wider impedance bandwidth. Since then, theoretical and experimental investigations have been presented in many publications presenting various shapes of DRAs such as spherical, cylindrical, cylindrical ring, rectangular,…etc. The rectangular DRA has some advantages over the cylindrical and hemispherical DRAs . For example, by choosing proper dimensions of the rectangular DRA, the mode degeneracy problem can be avoided and, in addition, the bandwidth can be optimized [3,4].In recent years, efforts have been paid on investigating the linearly polarized (LP) wide-band DRAs [5]–[6] and several methods have been proposed. The first one is to use more than one dielectric resonator (DR) elements [5],which is often called as a "stacked" DRA, with different sizes and/or dielectric materials. This method, however, will increase the antenna size and cost. The second approach is to use special-shaped DRAs [7]–[8], not be easy

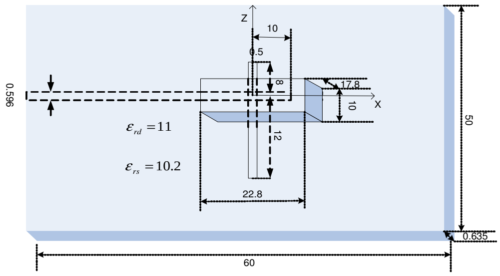

Fig.1 The configuration of DRA all dimensions in mm

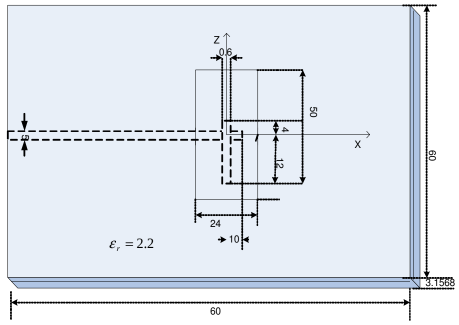

Fig.2 The configuration of MPA all dimensions in mm

to obtain commercially. Some feeding structures [7]–[9,10] can also be used to obtain a wide-band DRA.

In this paper, we analyzed, designed and fabricated a rectangular DRA. It operates in the S-band range of frequency with 20% bandwidth. The performance of this DRA is compared with the conventional MPA and found to have a much broader bandwidth.

2. Theory

In the model used in [4] the resonator is first considered as an infinite dielectric waveguide with perfect magnetic walls, to find the modal transverse fields. This guide is then truncated to form the rectangular resonator, and dominant transverse field components at these termination planes are matched with exponentially decaying fields outside the DR. Marcatili's approximation for the dielectric waveguide is used for the matching. Wave equations are solved in rectangular coordinates assuming sinusoidal steady-state time variation. Starting from magnetic and electric vector potential wave equations in the frequency domain given respectively by:

$$\nabla^2\overline{A}+k^2\overline{A}=0 \qquad (1)$$

and $$\nabla^2\overline{F}+k^2\overline{F}=0 \qquad (2)$$

where k is the complex propagation constant, assuming lossless dielectric region

$$k=\omega\sqrt{\varepsilon\mu} \qquad (3)$$

where ε and μ are the permittivity and permeability of the dielectric region, respectively. The rectangular components of $\overline{A}$ or $\overline{F}$ must satisfy the Helmholtz equation:

$$\nabla^2\psi+k^2\psi=0 \qquad (4)$$

where ψ is a scalar wave function. Helmholtz equation is solved in rectangular coordinate plane as a linear and inhomogeneous partial differential equation in three dimensions. For the TE^{z}_{111} mode (fig.1)

$A=0$ and $\overline{F}=\psi\,\hat{a}_z$

ψ is determined by solving Eq. (4) inside the resonator, by separation of variables. For symmetric modes around x, y and z, ψ may be written as

$$\psi=C\cos(k_x x)\cos(k_y y)\cos(k_z z) \qquad (5)$$

where C is an arbitrary constant. According to Harrington [6] the fields in the DR can be represented by:

$$\overline{E}=-\nabla\times\overline{F} \qquad (6)$$

and $$\overline{H}=-j\omega\varepsilon\overline{F}+\frac{1}{j\omega\mu}\nabla(\nabla\cdot\overline{F}) \qquad (7)$$

Substituting by $\overline{F}=\psi\,\hat{a}_z$ in Eqs.(6) and (7) using Eq. (5) for ψ we obtain

$$E_x=Ck_y\cos(k_x x)\sin(k_y y)\cos(k_z z) \qquad (8a)$$

$$E_y=-Ck_x\sin(k_x x)\cos(k_y y)\cos(k_z z) \qquad (8b)$$

$$E_z=0 \qquad (8c)$$

$$H_x=\frac{k_x k_z}{j\omega\mu}C\sin(k_x x)\cos(k_y y)\sin(k_z z) \qquad (9a)$$

$$H_y=\frac{k_y k_z}{j\omega\mu}C\cos(k_x x)\sin(k_y y)\sin(k_z z) \qquad (9b)$$

$$H_z=\frac{k_x^2+k_y^2}{j\omega\mu}C\cos(k_x x)\cos(k_y y)\cos(k_z z) \qquad (9c)$$

Considering a DR of dimensions a, b and d in the x, y and z directions, respectively and enforcing the perfect magnetic boundary conditions, PMC, $H_{\tan}=0$ and $E_{norm}=0$,

at $$x=\left|\frac{a}{2}\right| \Rightarrow E_x=0 \Rightarrow k_x=\frac{n\pi}{a}\ ,\quad n=1,2,\ldots,\infty \qquad (10a)$$

and at $$y=\left|\frac{b}{2}\right| \Rightarrow E_y=0 \Rightarrow k_y=\frac{m\pi}{b}\ ,\quad m=1,2,\ldots,\infty \qquad (10b)$$

Applying Marcatilis's approximation [4 and 11], to give an expression for the wave number k_z which has been derived from matching the DR's surface tangential fields at $z=\pm d/2$ with an external field which decays as it propagates normally from these surfaces. Expressing ψ as

$$\psi=X(x)Y(y)e^{-\beta z} \qquad (11)$$

where $\beta^2=k_{xo}^2+k_{yo}^2-k_o^2$. Thus, the field components outside the resonator are represented by:

$$E_x=Dk_{yo}\cos(k_{xo}x)\sin(k_{yo}y)e^{-\beta z} \qquad (12a)$$

$$E_y=-Dk_{xo}\sin(k_{xo}x)\cos(k_{yo}y)e^{-\beta z} \qquad (12b)$$

$$H_x=\frac{k_{xo}\beta}{j\omega\mu}D\sin(k_x x)\cos(k_y y)e^{-\beta z} \qquad (13a)$$

$$H_y=\frac{k_{yo}\beta}{j\omega\mu}D\cos(k_{xo}x)\sin(k_{yo}y)e^{-\beta z} \qquad (13b)$$

Applying mode matching at $z=+d/2$ for corresponding field components presented in Eqs. (8), (9) (12) and (13) and using mathematical manipulation the following result are obtained [3]

$$k_z\tan\left(\frac{k_z d}{2}\right)=\sqrt{(\varepsilon_r-1)k_o^2-k_z^2} \qquad (14)$$

Solving for k_o, which corresponds to the resonant frequency, $f_o=k_o c/2\pi$, of the TE^z_{111} mode, where c is the speed of light in free space. Adopting the same technique, one can design a rectangular DR resonate at a specific frequency, using the same set of equations and solving for the required dimension, d.

Different feeding systems may be used for the excitation of DRAs [12-13]. In this paper aperture coupling microstrip line feeding system is used, Fig.1. This choice is made mostly for the advantages of microstrip lines such as ease of integration, its low cost and compactness. The aperture coupled antenna may then be easily integrated on the same substrate with any other MIC subsystems. One of the main advantages of aperture-coupling is the isolation afforded between feed circuits and radiating elements which prevents the antenna radiation from altering the feed circuit's performance and also do not excite higher modes , also aperture could be used to enhance the bandwidth. Back radiation will be minimized since the aperture is normally chosen to be electrically small and nonresonant. The feed substrate parameters may be chosen to optimize the electrical function of transmitting the power to antenna elements by using electrically thin and high dielectric constant, low-loss material. This is in contrast to direct- or proximity-fed MPAs where a compromise must be made between the network's electrical function and the MPA cavity's radiation function. Another advantage of aperture feed is that it effectively and efficiently excites the lowest order resonant mode and yields good radiation patterns. As well, Pozar [14] noted that the bandwidth of the overall structure will be essentially that of the element and will not be affected by this coupling mechanism. Note that for DRAs, aperture-coupling involves only one substrate, for the feed network, whereas MPAs require two substrates to be aligned - one for the MPA and one for the feed- and the misalignment may affect the coupling, Fig.2.

Assuming that the aperture dimensions are negligible as compared to the dimensions of the DRA, and that the DRA is kept symmetrical with respect to the aperture, with d dimension of the DRA being parallel to the length of the aperture, the TE^{z}_{111} mode of the DRA is excited. The coupling between DRA and the aperture can be easily adjusted by controlling the length of the slot. If the dimensions of a given slot are such that maximum coupling to the DRA is more than the critical coupling, then a convenient method to obtain the critical coupling is to vary the position of the slot with respect to the DRA, i.e., by varying its position with respect to the z direction of the circuit. In this paper, the aperture was shifted about 4 mm to obtain a critical coupling.

3. FDTD Solution

A full-wave numerical analysis is the only way to obtain an exact estimation of all possible resonance frequencies of the DR. The DR is treated as a cavity surrounded by four PMC walls that is excited by Differential Gaussian Pulse at its center. Taflove [15] used orthogonal-grid Yee algorithm, Fig.3. Starting from Maxwell's curl equations in linear, isotropic, nondispersive, lossy materials and applying the concepts of FDTD numerical technique, the electric and magnetic fields at a fixed space point are expressed as [15]

$$\mathbf{E}^{n} = \frac{1-\dfrac{\sigma\Delta t}{2\varepsilon}}{1+\dfrac{\sigma\Delta t}{2\varepsilon}}\mathbf{E}^{n-1} + \frac{\Delta t/\varepsilon}{1+\dfrac{\sigma\Delta t}{2\varepsilon}}\nabla\times\mathbf{H}^{n-\frac{1}{2}} \qquad (15)$$

$$\mathbf{H}^{n+\frac{1}{2}} = \mathbf{H}^{n-\frac{1}{2}} - \frac{\Delta t}{\mu}\nabla\times\mathbf{E}^{n} \qquad (16)$$

Eqs.(15) and (16) are rewritten at the points $(i+1/2,j,k),(i,j+1/2,k),(i,j,k+1/2)$, according to Fig.3. Applying the boundary conditions at the surfaces of the DR and using the FFT, the resonance frequencies of the DR are obtained.

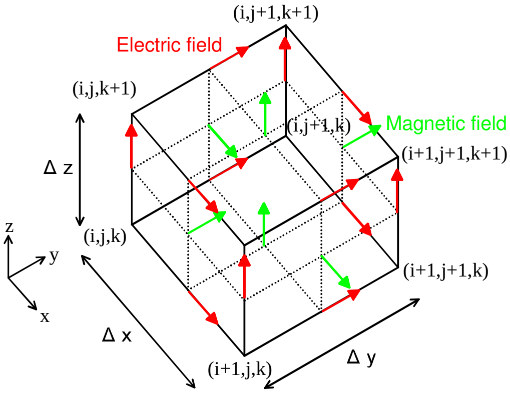

Fig.3 Orientation of both electric and magnetic fields in the Yee-space lattice [15]

4. Results

Matlab was used to build the code to solve the problem in which the FDTD numerical technique was adopted. A DRA using aperture coupling system was constructed in the microstrip laboratory in Electronics Research Institute, Egypt. The physical and geometrical properties of the DR used are as follows: $a = 22.8\,\text{mm}$, $b = 35.6\,\text{mm}$, $d = 10\,\text{mm}$ and $\varepsilon_r = 11$. The resonant frequencies obtained are shown in Fig.4.

The performance of the fabricated antenna was measured, using HP 8719E5 vector network analyzer. The characteristics of the circuit calculated using HFSS are: $Gain = 6.62dB$, $Directivity = 6.7dB$, $impedance\ \ beamwidth = 15.8\%$, $\eta = 98.18\%$ and $front-back-ratio = 13.5dB$.

Results of different methods are compared in Table 1. From Table 1, it is concluded that the results of FDTD adopted in this research achieves the least percentage error as compared to experimental results. Simulated results obtained from The HFSS software are shown in Figs.5 and 6,while fig.7 shows the simulated and measured return loss. It is clear from this figure that the measured impedance bandwidth is about 20%. A microstrip patch antenna was designed to operate at the same operating frequency of the DRA. The return loss and radiation patterns of the MPA are presented in Figs. 8,9 and 10.It is clear from fig.8 that the measured impedance bandwidth of the MPA is about 3.7%. This means that for the same type of feeding we obtained a wider bandwidth in the case of the DRA than the MPA (almost five times). One can not generalize this remark in all the cases.

5. CONCLUSION

In this paper a wide band rectangular dielectric resonator antenna with microstrip slot excitation is presented using FDTD technique. It was found that this antenna has a broader bandwidth as compared with the conventional microstrip antenna using the same type of feeding. This remark can not be generalized.

TABLE 1. RESONANT FREQUENCY OF DRA OBTAINED USING DIFFERENT METHODS

	Resonance frequency in GHz	Error in resonance frequency	Quality Factor
Closed formula	3.2311	-9.8	7.3682
FDTD	3.4493	-3.8	5.3488
HFSS	3.4343	-4.2	5.4638
Experimental	3.586	0.00	4.4279

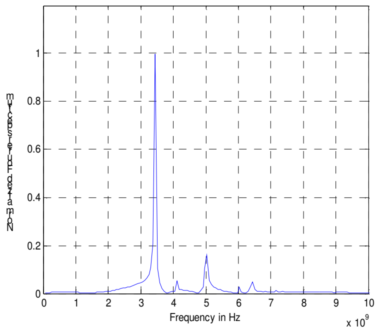

Fig.4 FDTD results using FFT on the DRA problem shown in Fig.1

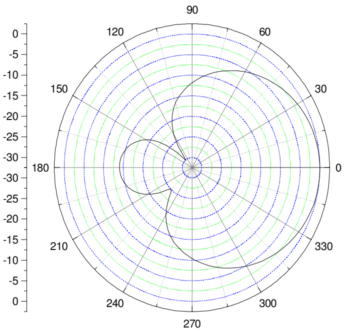

Fig.5 E-plane radiation pattern of TE_{111}^{z} mode of microstrip slot coupled rectangular DRA (shown in Fig.1).

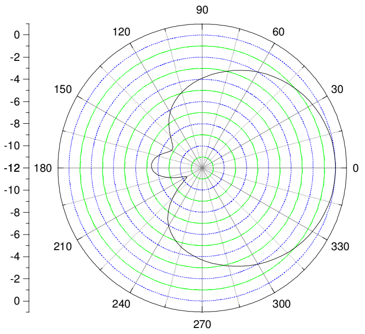

Fig.6. H-plane radiation pattern of TE_{111}^{z} mode of microstrip slot coupled rectangular DRA (shown in Fig.1).

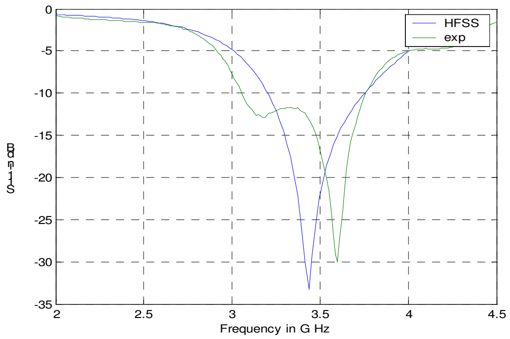

Fig.7 Return loss of the DRA shown in Fig.1 experimentally and using the HFSS

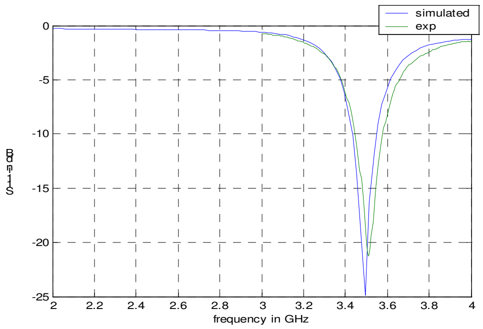

Fig.8 Return loss of the MPA shown in Fig.2 experimentally and using the HFSS

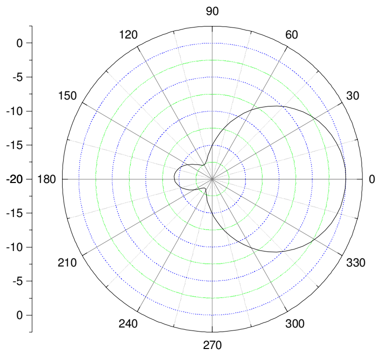

Fig.9. E-plane radiation pattern of MPA (shown in Fig.2).

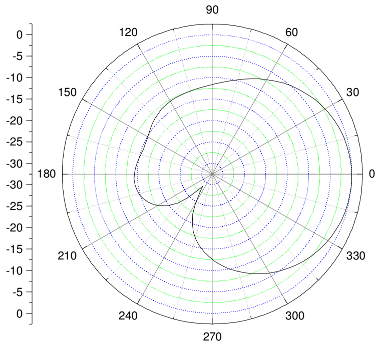

Fig.10 H-Plane radiation pattern of MPA (shown in Fig.2).

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