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The Effects of Air Gap and Finite Ground Plane on Radiation Characteristics of Probe-Fed Rectangular Dielectric Resonator Antennas Using Method of Moment

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Abstract **— The rigorous moment method analysis of probe-fed rectangular dielectric resonator antenna on a finite ground plane is presented. We modeled the exact structure of coaxial cable and antenna. The antenna, coaxial cable and ground plane are modeled as surface electric currents, but the dielectric resonator and internal dielectric of coaxial cable are assumed as a volume polarization current. The dielectric resonator (DR) is treated via a set of combined field integral equations. The associated coupling is then formulated with sets of integral equations. The coupled integral equations are solved by the method of moments (MoM). The effects of air gap and finite ground plane on resonance frequency, input impedance, E-plane and H-plane patterns and moment current of the objects are presented. Finally, we demonstrate some important conceptual issues for designing and analyzing of RDRA in communication circuits such as mobile systems, IF- amplifiers and phase detection. Our methods and results obtained from the MoM have a very good agreement with measurements and are higher accurate and fast than simulated results with the other softwares.**

Index Terms **— Probe, coaxial cable, rectangular dielectric resonator, finite ground plane, air gap.**

I. Introduction

Dielectric resonators (DRs) are widely used in shielded microwave circuit such as filters and oscillators. In recent years the dielectric resonator antenna (DRA) has become the subject of many researches. This antenna offers advantages such as compact size, high radiation efficiency and wide bands simple feed structure over conventional types of antennas. Also they are very compatible with MMIC fabrication. Literature survey shows that dielectric resonators have been studied in hemispherical, cylindrical, cylindrical ring and rectangular geometry. By comparing the other geometries it can be found that RDRs have a few advantages in fabrication process, and electrically they have some independent aspect ratios, which could be chosen to provide the required radiation characteristics [1-8]. Input impedance and resonance frequency of an aperture coupled and probe-fed rectangular dielectric resonator antenna has been computed using method of moment [9-11]. In this paper, the dielectric and the internal dielectric of coaxial cable are modeled using volume polarization currents. The probe, coaxial cable and ground plane are modeled using surface currents. These are formulated as a set of coupled equations in the spatial domain using the integral equation formulation. The MoM is then applied to the resulting operator equation to obtain the unknown currents. The resulting current is then used to investigate secondary parameters such as input impedance and pattern of antenna versus air gap between dielectric and ground plane and the radius of ground plane.

II. Formulations and Integral Equations

Fig.1 shows the structure of a probe-fed RDRA on a finite ground plane with an air gap between dielectric resonator and ground plane. The diameter of ground plane is 100 mm. The excitation is very important for moment analysis of the structure. Two issues are very important for this new scheme of excitation:

1. The matching of coaxial cable with the antenna.
2. The exact experimental setup

With regards to these two issues we excited our system by a sinusoidal signal on the inner conductor and another sinusoidal signal with a reverse direction on the outer conductor of coaxial cable. The wavelength of this sinusoidal signal is the guided wavelength of coaxial cable:

$$\lambda_g = v_{coax}/f$$

In which v_{coax} and f are the group velocity of coaxial cable and frequency, respectively. Assuming this excitation, the source modeling is the same as experimental setup.

$$\vec{J}_{feed}=\begin{cases} -\sin(2\pi/\lambda_g.(z-L_{antenna}))\hat{z} & , \\ \text{on the antenna and core of coaxial cable} & \\ (Ra/Rb).\sin(2\pi/\lambda_g.(z-L_{antenna}))\hat{z} & , \\ \text{on the clad of coaxial cable} & \end{cases}$$

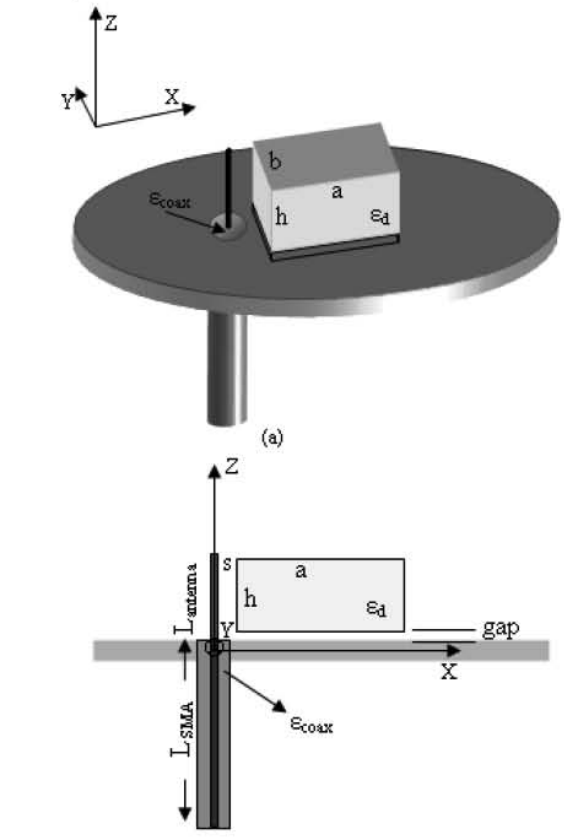


Fig. 1 The probe-fed rectangular dielectric resonator antenna on a finite ground plane. a) 3-D view. b) Front view.

We assume that both surface electric currents on the antenna are present since dielectric resonator has significant effects on the current of antenna along $\hat{\varphi}$, but the surface electric current on the core and shield of coaxial cable are only along $\hat{z}$. Also, the polarization currents in internal dielectric of coaxial cable have the radial direction. The polarization currents in the dielectric resonator have three directions. Finally; the surface currents on the finite ground plane are assumed in two directions. Now by these assumptions we can say that the total electric field at any point is:

$$\vec{E}_{total}=\vec{E}_{impressed}(\vec{J}_{feed})+\vec{E}_s(\vec{J}_{antenna})+ \vec{E}_s(\vec{J}_{core})+\vec{E}_s(\vec{J}_{clad})+\vec{E}_s(\vec{J}_{intdc})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g) \quad (1)$$

Where,

$\vec{E}_{impressed}$: Impressed field of the feeding currents on the core and clad of coaxial cable and the antenna.

$\vec{Es}$: Scattered field

The moment currents on each object have the following structure:

$\vec{J}_{antenna}=J_{a\varphi}\hat{\varphi}+J_{az}\hat{z}$

$\vec{J}_{core}=J_{corez}\hat{z}$, $\vec{J}_{clad}=J_{cladz}\hat{z}$

surface current on the core and sheild of coaxial cable;

$\vec{J}_{intdc}=J_{\rho c}\hat{\rho}$,

volume polarization current in internal dielectric of coaxial cable;

$\vec{J}_d=J_{dx}\hat{x}+J_{dy}\hat{y}+J_{dz}\hat{z}$,

volume polarization current of dielectric resonator;

$\vec{J}_g=J_{g\rho}\hat{\rho}+J_{g\varphi}\hat{\varphi}$,

surface current on ground plane

Also, the total current on the antenna, the core and the shield of coaxial cable are:

$\vec{J}_{TOTALantenna}=\vec{J}_{antenna}+\vec{J}_{feed}$

$\vec{J}_{TOTALcore}=\vec{J}_{core}+\vec{J}_{feed}$

$\vec{J}_{TOTALclad}=\vec{J}_{clad}+\vec{J}_{feed}$

By enforcing the boundary conditions on antenna and finite ground plane, the following equations must be satisfied on them:

$$\hat{z}\times[\vec{E}_{impressed}(\vec{J}_{feed})+\vec{E}_s(\vec{J}_{antenna})+\vec{E}_s(\vec{J}_{core})+ \vec{E}_s(\vec{J}_{clad})+\vec{E}_s(\vec{J}_{intdc})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g)]=0$$

on the surface of ground plane (2) , (3)

$$\hat{z}\bullet[\vec{E}_{impressed}(\vec{J}_{feed})+\vec{E}_s(\vec{J}_{antenna})+ \vec{E}_s(\vec{J}_{core})+\vec{E}_s(\vec{J}_{clad})+\vec{E}_s(\vec{J}_{intdc})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g)]=0$$

on the antenna , shield and core of coaxial cable (4-6)

$$\hat{\varphi}\bullet[\vec{E}_{impressed}(\vec{J}_{feed})+\vec{E}_s(\vec{J}_{antenna})+ \vec{E}_s(\vec{J}_{core})+\vec{E}_s(\vec{J}_{clad})+\vec{E}_s(\vec{J}_{intdc})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g)]=0$$

on the antenna (7)

Since there are polarization currents present in the volume of DR and internal dielectric of coaxial cable, the integral equation in each volume is:

$$\vec{E}_{\text{impressed}}(\vec{J}_{\text{feed}})+\vec{E}_s(\vec{J}_{\text{antenna}})+\vec{E}_s(\vec{J}_{\text{core}})+$$
$$\vec{E}_s(\vec{J}_{\text{clad}})+\vec{E}_s(\vec{J}_{\text{intdc}})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g)=j\omega(\varepsilon_d-\varepsilon_0)\vec{J}_d$$

inside the dielectric resonator (8-10)

$$\vec{E}_{\text{impressed}}(\vec{J}_{\text{feed}})+\vec{E}_s(\vec{J}_{\text{antenna}})+\vec{E}_s(\vec{J}_{\text{core}})+\vec{E}_s(\vec{J}_{\text{clad}})$$
$$+\vec{E}_s(\vec{J}_{\text{intdc}})+\vec{E}_s(\vec{J}_d)+\vec{E}_s(\vec{J}_g)=j\omega(\varepsilon_{coax}-\varepsilon_0)\vec{J}_{intdc}$$

in the volume of internal dielectric of coaxial cable (11)

The coupled equations 2-11 can be written in an integral form by expressing the fields in terms of vector potential A:

$$G_0=\frac{e^{-jkR}}{4\pi R}$$
$$\vec{E}(\vec{J}_{cell},\vec{r})=-j\omega\vec{A}(\vec{J}_{cell},\vec{r})-\frac{j}{\omega\mu\varepsilon_0}\vec{\nabla}(\vec{\nabla}\vec{A}(\vec{J}_{cell},\vec{r}))$$
$$\vec{A}_V(\vec{J},\vec{r})=\mu_0\iiint_V \vec{J}.G_0 dv' \quad , \quad \vec{A}_S(\vec{J},\vec{r})=\mu_0\iint_S \vec{J}.G_0 dv' \qquad (12)$$

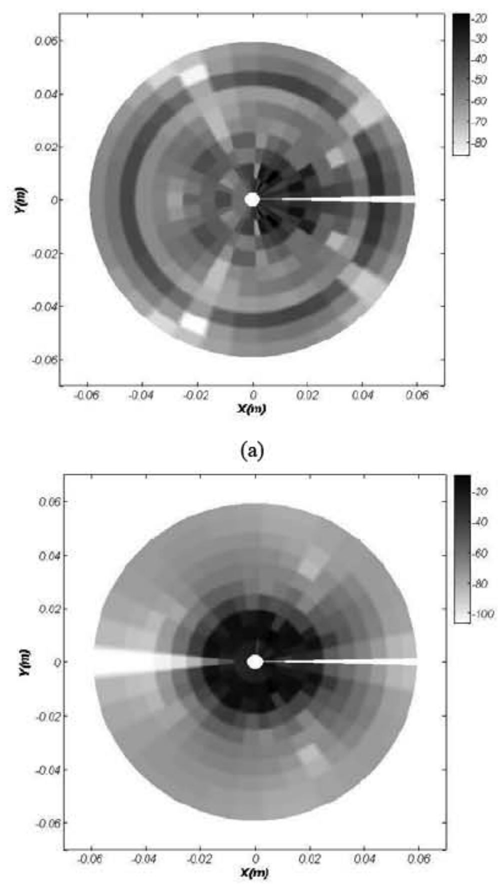
Where G_0 is the corresponding Green function in free space.

MoM procedure is used to solve the coupled equations 2-11. Because of complexity of the structure and choosing the best functions for convergence, pulse and Dirac functions was chosen respectively as a basis function for expanding the currents and as a testing function.

After calculating the unknown currents for all of the structure, the voltage between core and the shield of coaxial cable with the distance of (R_b–Ra) has been found according to the total of electric field in that region. After that, the input impedance will be obtained by dividing this voltage to the total current on the core of the coaxial cable.

III. Numerical Results

After calculating the unknown currents, we can see that the currents on the ground plane are concentrated under the dielectric resonator (Fig. 2). Fig. 2 shows that the radial and angular currents are zero at the end of the ground plane; however, the radial current reaches its maximum value under the dielectric resonator and the angular current has the odd symmetry with respect to $\varphi=0$. Thus, these obtained currents on the ground plane in presence of dielectric resonator show that our results are correct. By comparing the input impedance obtained by method of moment with measurements and simulated results by HFSS [10], the higher accuracy of our method and modeling is noted. The parameters of the analyzed DRA are listed in Table I These parameters are



(a)

(b)

Fig.2 The obtained magnitude of current on the ground plane. a)Radial direction. b) Angular direction.

as the same as parameters in [10]. A good agreement between experimental and computed results has been found in Fig.3 and Fig.4.

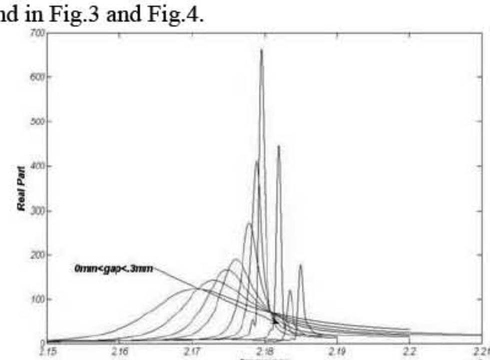


Fig. 3 The real part of input impedance versus frequency and air gap between resonator and ground plane.

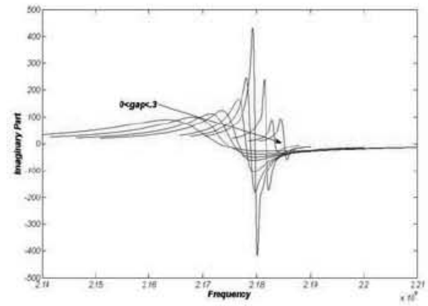


Fig. 4 The imaginary part of input impedance versus frequency and air gap between resonator and ground plane.

The investigation of the effects of air gap between dielectric and ground plane are shown in Fig.3 and Fig.4.we can see that by increasing the air gap, bandwidth and resonance frequency respectively has been decreased and increased. Also, by doing a comparison between these results and the experimental results, the actual value of the air gap can be predicted 30-50 micrometer. These figures are shown that the air gap have powerful effects on performance of the antenna and its matching procedure.

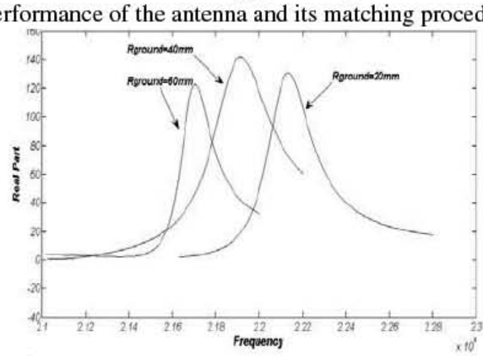


Fig. 5 The real part of input impedance versus frequency and the radius of ground plane.

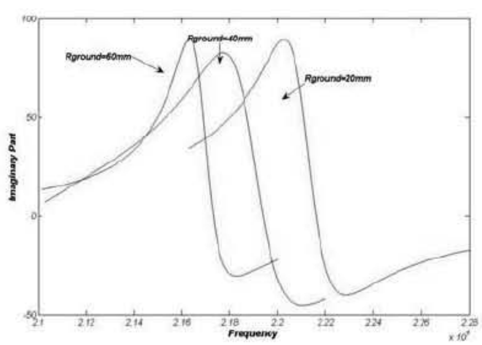


Fig. 6 The imaginary part of input impedance versus frequency and the radius of ground plane.

The effects of the radius of ground plane on input impedance are investigated in Fig. 5 and Fig. 6. These figures show that the variation of radius of ground plane

TABLE I

THE PARAMETERS OF THE ANTENNA STRUCTURE

$L_{antenna}$	L_{SMA}	R_b	Ra	h	a	b
9.5mm	15.5mm	2mm	.65mm	9.5mm	19mm	19mm

has a little effect on the shape of the input impedance; but the decrement of this radius has been caused the increasing of the resonance frequency and back-lobe of the pattern of the antenna; Also, we can see that the optimum bandwidth and matching can be achieved by choosing 30-40 millimeter for the radius of the ground plane.

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