Radiation Characteristics of Dielectric Resonator Antennas
Loaded with a Beam-Forming Ring

by Ahmed A. Kishk* and Atef Z. Elsherbeni*

A technique for improving the radiation characteristics of resonant dielectric cavity antennas by using a circular conducting ring is proposed. The ring is used as a director in front of the antenna in order to obtain equal \( E \) and \( H \) plane far field patterns in the forward direction. The analysis of the problem is expressed in terms of integro-differential equations for the equivalent surface electric and magnetic currents on the various dielectric and metallic interfaces of the antenna. The method of moments is then used to solve the resulting equations. The effects of the thickness, diameter and height of the ring on the \( E \) and \( H \) plane radiation patterns as well as the cross-polarized field are presented for cylindrical and hemi-spherical dielectric cavity antennas.

**Strahlungscharakteristik von einem strahlformenden Ring belasteten dielektrischen Resonator-Antennen**


1. Introduction

The basic geometry of a dielectric cavity antenna consists of a dielectric material which resides on a conducting ground plane and is excited by a coaxial probe through the back of the ground plane. In recent articles [1]-[3], attention has been focused on the study of the radiation characteristics of this type of antennas through the proper choice of shape, relative permittivity of the dielectric material and appropriate excitation. Previous attempts to improve the symmetry of the radiation pattern as well as to reduce the backlobe level by placing the antenna inside a (topless cylindrical cavity) are also reported [4]. The purpose of this paper is to provide a simple and at the same time, practical alternative method for improving the radiation characteristics of these antennas by loading the dielectric cavity by a beam forming conducting ring. The use of a conducting ring is to be investigated in order to improve the radiation characteristics of a dipole-disk antennas which can be used as a prime focus feed for reflector antennas [5].

Analytical analysis of the spherical dielectric cavity antenna was carried out using the method of separation of variables [3], while approximations were employed for the analysis of cylindrical [1] and rectangular dielectric cavity antennas [2]. Specifically, the radiation patterns are evaluated in terms of equivalent magnetic currents on the surfaces of the resonator, which are determined based on the fields that exist in the resonator while it is covered by a perfect magnetic conductor. Furthermore, it has been assumed that the groove plane is infinite, in order to use the image theory in the analysis. In this paper, the analysis of dielectric cavity antennas loaded by a conducting ring is based on deriving integro-differential equations for the equivalent surface electric and magnetic currents on the various dielectric and metallic interfaces of the antenna. The solution of these equations is then carried out numerically using the method of moments. The method is rather general and allows for the investigation of a wide variety of geometries as long as they are axisymmetrical.

2. Basic Formulation

Consider a general geometry of an antenna made of combinations of dielectric and conducting materials as illustrated in Fig. 1 a. For the shown geometry, there are three distinct regions: \( V_0 \) constituting the perfectly conducting materials; \( V_1 \), the exterior of the scatterer, characterized by the constitutive parameters \( \varepsilon_r \) and \( \mu_r \); and \( V_2 \), the dielectric region, characterized by \( \varepsilon_r \) and \( \mu_r \). The excitation is simulated by an infinitesimal \( z \) directed electric dipole of current moment \( I_j \). The dipole is located in the dielectric medium and produces the incident field components \( E \) and \( H \). The total electric and magnetic fields in region \( V \) are denoted by \( E \) and \( H \), respectively. The normal unit vector \( \mathbf{n} \) points into the region \( V \). The boundary

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surfaces of the region \( V_1 \) are denoted by \( s_1; s_2 \) denotes the boundary surface between \( V_0 \) and \( V_1 \), \( s_3 \) denotes the boundary surface between \( V_2 \) and \( V_1 \), and \( s_4 \) denotes the boundary surface between \( V_2 \) and \( V_0 \). The total electric and magnetic fields in regions \( V_1 \) and \( V_2 \) are unknown. By applying the equivalence principle \( [6] \), the original problem can be split into two auxiliary problems namely, the external and internal problems as shown in Fig. 1b and 1c, respectively. The unknown electric and magnetic fields can then be expressed in terms of unknown electric and magnetic surface currents \( J_1 \) and \( M_1 \), respectively as follows:

\[
J_2 = \eta_1 E_2 + H_2 \quad \text{on} \quad s_3, \\
M_2 = -E_2 \times H_2 \quad \text{on} \quad s_4.
\]

Therefore in region \( V_1 \), at a field point \( r \), the total electric and magnetic fields can be expressed in terms of integro-differential operators \( L \) and \( \kappa \) as:

\[
\theta(r) E_1(r) = -L_1 J_1(r) + \kappa_1 M_1(r),
\]

\[
\theta(r) H_1(r) = -\kappa_1 J_1(r) - \eta_1 \kappa_2 L_1 M_1(r),
\]

whereas in region \( V_2 \) the corresponding field expressions are:

\[
\theta(r) E_2(r) = E_2(r) - L_2 J_2(r) + \kappa_2 M_2(r),
\]

\[
\theta(r) H_2(r) = H_2(r) - \kappa_2 J_2(r) - (\eta_1 \kappa_2)^2 L_2 M_2(r),
\]

where

\[
\theta(r) = \begin{cases} 1 & r \in V_1, \\ 0 & r \in V_2 \end{cases}, \quad \kappa_1 = 1.2, \quad \kappa_2 = 0.2.
\]

The Green’s function for an unbounded medium is denoted by \( \phi_0(r) = \exp(i x \cdot r)/(4\pi r) \) where \( x = \omega(\mu_0 J_0) \), \( \mu_0 = J_0 = \mu_0 J_0 \), and \( \mu = \mu_0 \nu_0 \), \( \nu_0 = (\mu_0 J_0)^{1/2} \). \( \mu_0 \) and \( \nu_0 \) are the permittivity and permeability of free space, respectively, and the subscript \( r \) denotes the relative constitutive quantity. When \( r \rightarrow \infty \) the integrals defining \( L_i \) and \( \kappa_i \) must be interpreted in the Cauchy principal value sense.

By enforcing the boundary conditions, i.e. the tangential components of the electric and magnetic fields must be continuous on the dielectric surface \( s_4 \), one finds that \( J_2 = J_1 \) and \( M_2 = -M_1 \) as long as \( \kappa_1 = -\kappa_2 \). Furthermore, on the conductor surfaces only the electric surface currents exist i.e. \( J_1 = J_2 = 0 \) on \( s_3 \) and \( J_2 = 0 \) on \( s_4 \). Using the field representations in eqs. (5) and (6), the boundary conditions reduce to:

\[
\frac{1}{\eta_1} (-L_1 J_1 + J_2) + \kappa_2 M_1 = \theta(r),
\]

\[
\frac{1}{\eta_1} (-L_1 J_1 - J_2) + \kappa_2 M_1 = \theta(r),
\]

\[
\frac{1}{\eta_1} (-L_1 J_2 + J_2) - \kappa_2 M_1 = \theta(r),
\]

\[
\frac{1}{\eta_1} (-L_1 J_2 - J_2) - \kappa_2 M_1 = \theta(r).
\]

3. Method of Moment Solution

The above surface integral equations (10) to (13) are then solved by the method of moments. Since the geometry of any of the considered dielectric resonator antennas is in general a body of revolution, the surfaces are divided into annular rings and the surface currents are expanded as

\[
J = \sum_{n=1}^{M} \left( i \xi_n P_n^1 + \eta \xi_n P_n^2 \right) \quad \text{on} \quad s_3, s_4, \text{ or } s_0.
\]

\[
M = \sum_{n=1}^{M} \left( \xi_n M_n^1 + \eta \xi_n M_n^2 \right) \quad \text{on} \quad s_3, s_4, \text{ or } s_0
\]

where

\[
J_n = \theta(r) \left[ \phi_n(i) \phi(i) \right] e^{i \phi(i)}.
\]

while \( \xi_n \) is a triangle function, \( \phi(i) \) is the radial distance from the axis of symmetry as defined in [7]. \( \phi \) is a unit vector in the \( z \) direction where a separate tangent component \( i \phi \) of the body of revolution. \( \xi \) and \( \mu \) are the unknown expansion coefficients of \( J_n \) and \( M_n \), respectively. By applying Galerkin's method on eqs. (10) to (13) with testing functions \( W_m = \left( e^{i \phi(i) \phi} \right) \), where \( \phi \) denotes the Fourier mode expansion, the following matrix equation is obtained.
where \( \eta \) and \( \psi \) are the permeability and conductivity, respectively. The first pair of suffixes identifies field surface and the second pair of suffixes identifies the source surface and the Fourier mode \( n \) is implied. Every sub-matrix consists of four sub-matrices as a result of the inner product of the vectorial testing functions and the left hand side of eq. (10) to (13).

The right hand side column is the excitation matrix, due to the electric dipole, in which each element is given by

\[
E_1^r = \frac{1}{4\pi} \int \mathbf{W}^r \cdot \mathbf{E}_{\text{inc}}^r \, d\mathbf{x},
\]

\[
E_2^r = \frac{1}{4\pi} \int \mathbf{W}^r \cdot \mathbf{H}_{\text{inc}}^r \, d\mathbf{x},
\]

It is worth mentioning that the electric dipole is a simulation of the coaxial feed probe in the dielectric region. The radiated electric and magnetic fields produced by the electric dipole in an unbounded region are given by

\[
E_1^r = -j\omega \mu \mathbf{A} \cdot \mathbf{E},
\]

\[
H_2^r = \frac{1}{\mu_0} \mathbf{V} \times \mathbf{A},
\]

where

\[
\mathbf{A} = \frac{1}{4\pi} \int \frac{1}{r} \mathbf{I}_1 \cdot \mathbf{I}_2 \cdot k_1^2 (k_{1r} \mathbf{r} \cdot k_{1r} - \mathbf{r} \cdot \mathbf{r})
\]

\[
\Phi = \frac{\eta_0}{4\pi} \int \frac{1}{r} \mathbf{E}_1 \cdot \mathbf{E}_2 \cdot k_1^2 (k_{1r} \mathbf{r} \cdot k_{1r} - \mathbf{r} \cdot \mathbf{r})
\]

are, respectively, the magnetic vector and electric scalar potentials produced by the dipole. In eqs. (22) and (23) \( k_0^2 \) is the spherical Hankel function of the second kind and zero order, \( \mu_0 \) and \( \eta_0 \) are the permeability, wave number and intrinsic impedance of the dielectric medium, respectively and \( \mathbf{r}, \mathbf{r} \) are the positional vectors of the field and source points on the antenna surface, respectively.

By numerically solving the matrix system in eq. (17), the unknown expansion coefficients can be determined and \( J \) and \( M \) can then be evaluated. It should be noted that the matrix solution provides the currents on the outer and inner surfaces of the antenna.

The far field components \( E_\infty \) and \( H_\infty \) can be written in terms of the currents on the outer surfaces of the antenna as

\[
E_\infty = -\frac{j}{4\pi r_1} e^{-j\phi_0} F_1(\theta, \phi),
\]

\[
H_\infty = -\frac{j}{4\pi r_1} e^{-j\phi_0} F_2(\theta, \phi),
\]

where the angular functions \( F_1 \) and \( F_2 \) are given by

\[
F_1(\theta, \phi) = \int [J \cdot \mathbf{d}_k + (1/\eta_0) M \cdot \mathbf{d}_k] \exp(-j k_1 r_1 \phi) \, d\phi
\]

\[
F_2(\theta, \phi) = \int [J \cdot \mathbf{d}_k + (1/\eta_0) M \cdot \mathbf{d}_k] \exp(-j k_1 r_1 \phi) \, d\phi
\]

(26)

(27)

where \( j \) is the outer total surface of the antenna \( \mathbf{d}_k \) is a unit vector towards the field point, \( \phi \) is a vector representing the source point on a while \( k_1 \) and \( \eta_0 \) are the wave number and intrinsic impedance in free space, respectively. \( \mathbf{d}_k \) and \( \mathbf{d}_k \) are unit vectors in the direction of increasing \( \phi \) and \( \phi \) respectively.

4. Results and Discussion

The excitation of the considered dielectric resonator antenna is modeled by an infinitesimal electric dipole located at \( x = 0.16 \) and \( z = 0.05 \) inside the dielectric material where \( \phi = 0 \) is the free space wavelength. In Fig. 2, the cylindrical dielectric material has a diameter \( d = 3.34 \) and \( h = 0.11 \lambda \) while its relative permittivity \( \varepsilon_r = 8.9 \). These parameters are selected in order to excite the TM11 mode for which the radiation pattern has a peak in the broad side direction \( \phi \). It is also worth mentioning that since the TM11 mode is the dominant mode in our calculations, the asymmetric nature of the radiation patterns is due to contributions of the higher order modes. For the antenna in Fig. 2, an infinite perfectly conducting ground plane is assumed. The image theory is then used to construct an equivalent problem in order to solve numerically for the radiated fields in the upper region. From the resulting radiation patterns, it is obvious that the H plane pattern \( E_\phi \) is symmetric around the \( z = 0 \) and vanishes at \( \phi = 90^\circ \). On the other hand the variations in the E plane pattern \( E_\theta \) is limited to 3 dB, while the resulting cross-polarization level, in the planes \( \phi = 45^\circ \) and \( \phi = 225^\circ \), is high due to the unequal E and H plane patterns. For the same antenna parameters as in Fig. 2, a conducting ring of diameter \( d = 0.28 \) and thickness \( h = 0.02 \) is placed just on top of the cylindrical dielectric material where the distance between the top surface of the ground plane and the center of the ring (idented by \( k_1 \)) is set to be equal to 0.112 z as shown in Fig. 3. The resulting beamwidths of the E and H plane pattern became almost identical, consequently the cross-polarization level is reduced by approximately 7 dB. Furthermore, another antenna whose dielectric material is constructed from a semi-sphere is considered. This is
shown in Fig. 4 where the sphere diameter $d$ is chosen to be equal to 0.32 $\lambda$ and $\varepsilon_r = 8.9$ in order to support the TM$_{11}$ mode [3]. The radiation patterns of the hemispherical dielectric resonator antenna above an infinite ground plane (as shown in Fig. 4) are found to be similar to those of the cylindrical dielectric resonator antenna in Fig. 3. When a conducting ring with $d_r = 0.2040 \lambda$, $a_s = 0.1202 \lambda$, and $a_t = 0.02 \lambda$ is added, as shown in Fig. 5, one can easily notice the resulting improvements in the symmetry of the $E$ and $H$ plane patterns as well as the reduction of the cross-polarized field level. It is worth mentioning that the values of the ring diameter, thickness and height that we used in Figs. 3 and 5 are the optimal parameters for these specific antennas. While optimizing the ring parameters, it is found that the best height is that when the ring is almost touching the top surface of the dielectric material of the cylindrical antenna and when $d_r$ is approximately equal to 2 $\lambda$, for the hemispherical antenna. Moreover, the effect of the ring parameters ($a_s$, $d_r$, and $a_t$) on $E$ and $H$ plane beamwidths and the cross-polarization level at $\phi = 45^\circ$, $\theta = 90^\circ$ of the proposed resonator antennas are presented in Tables 1 to 4 for the cylindrical antennas shown in Fig. 2 and in Tables 5 to 8 for the hemispherical antenna shown in Fig. 4. In all considered cases, the cross-polar-
<table>
<thead>
<tr>
<th>( \alpha_0 ) (%)</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.02</td>
<td>72.61</td>
<td>75.11</td>
<td>-12.16</td>
</tr>
<tr>
<td>0.04</td>
<td>73.70</td>
<td>77.03</td>
<td>-11.79</td>
</tr>
<tr>
<td>0.06</td>
<td>79.03</td>
<td>79.56</td>
<td>-11.86</td>
</tr>
<tr>
<td>0.08</td>
<td>76.78</td>
<td>81.48</td>
<td>-10.64</td>
</tr>
<tr>
<td>0.1</td>
<td>89.14</td>
<td>86.35</td>
<td>-9.80</td>
</tr>
</tbody>
</table>

Table 2. The effect of the ring diameter on the radiation characteristics of a cylindrical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.3412 \), \( \beta = 0.12 \), \( a_0 = 0.12 \), and infinite ground plane.

<table>
<thead>
<tr>
<th>( \delta ) (%)</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2404</td>
<td>98.82</td>
<td>81.57</td>
<td>-10.47</td>
</tr>
<tr>
<td>0.26</td>
<td>72.27</td>
<td>78.80</td>
<td>-11.66</td>
</tr>
<tr>
<td>0.28</td>
<td>72.61</td>
<td>75.11</td>
<td>-12.16</td>
</tr>
<tr>
<td>0.3</td>
<td>63.66</td>
<td>70.61</td>
<td>-11.13</td>
</tr>
<tr>
<td>0.32</td>
<td>58.93</td>
<td>68.47</td>
<td>-10.78</td>
</tr>
</tbody>
</table>

Table 3. The effect of the ring height on the radiation characteristics of a cylindrical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.3412 \), \( \beta = 0.12 \), \( a_0 = 0.12 \), and infinite ground plane.

<table>
<thead>
<tr>
<th>( \alpha_0 ) (%)</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.12</td>
<td>72.61</td>
<td>75.11</td>
<td>-12.16</td>
</tr>
<tr>
<td>0.14</td>
<td>82.75</td>
<td>79.69</td>
<td>-12.68</td>
</tr>
<tr>
<td>0.16</td>
<td>83.07</td>
<td>82.41</td>
<td>-12.02</td>
</tr>
<tr>
<td>0.18</td>
<td>89.64</td>
<td>84.90</td>
<td>-11.23</td>
</tr>
<tr>
<td>0.2</td>
<td>96.40</td>
<td>87.48</td>
<td>-10.35</td>
</tr>
<tr>
<td>0.25</td>
<td>-</td>
<td>94.80</td>
<td>-7.92</td>
</tr>
</tbody>
</table>

Table 4. The frequency characteristics of a cylindrical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.3412 \), \( \beta = 0.12 \), \( a_0 = 0.28 \), \( \alpha_0 = 0.2 \), and infinite ground plane.

<table>
<thead>
<tr>
<th>Frequency ( f/\lambda_0 )</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.95</td>
<td>77.94</td>
<td>77.82</td>
<td>-14.92</td>
</tr>
<tr>
<td>0.96</td>
<td>76.72</td>
<td>77.44</td>
<td>-14.00</td>
</tr>
<tr>
<td>0.97</td>
<td>75.22</td>
<td>76.97</td>
<td>-13.93</td>
</tr>
<tr>
<td>0.98</td>
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<td>-12.43</td>
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<td>0.99</td>
<td>73.19</td>
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<td>-12.83</td>
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<td>1.0</td>
<td>72.61</td>
<td>75.11</td>
<td>-12.16</td>
</tr>
<tr>
<td>1.01</td>
<td>70.95</td>
<td>75.96</td>
<td>-11.43</td>
</tr>
<tr>
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<td>1.03</td>
<td>71.94</td>
<td>73.70</td>
<td>-9.92</td>
</tr>
<tr>
<td>1.04</td>
<td>73.08</td>
<td>72.92</td>
<td>-9.14</td>
</tr>
<tr>
<td>1.05</td>
<td>75.75</td>
<td>72.12</td>
<td>-8.36</td>
</tr>
</tbody>
</table>

Table 5. The effect of the ring height on the radiation characteristics of a semi-spherical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.32 \), \( \beta = 0.12 \), \( a_0 = 0.12 \), \( \alpha_0 = 0.02 \), and infinite ground plane.

<table>
<thead>
<tr>
<th>( \delta ) (%)</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.2404</td>
<td>73.31</td>
<td>76.70</td>
<td>-18.64</td>
</tr>
<tr>
<td>0.26</td>
<td>81.39</td>
<td>80.00</td>
<td>-14.36</td>
</tr>
<tr>
<td>0.28</td>
<td>73.52</td>
<td>76.94</td>
<td>-15.63</td>
</tr>
<tr>
<td>0.3</td>
<td>69.97</td>
<td>73.46</td>
<td>-15.95</td>
</tr>
<tr>
<td>0.32</td>
<td>55.81</td>
<td>66.22</td>
<td>-11.70</td>
</tr>
</tbody>
</table>

Table 6. The effect of the ring height on the radiation characteristics of a hemi-spherical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.32 \), \( \beta = 0.12 \), \( a_0 = 0.12 \), \( \alpha_0 = 0.02 \), and infinite ground plane.

<table>
<thead>
<tr>
<th>( \delta ) (%)</th>
<th>E-Plane 3 dB-BW</th>
<th>H-Plane 3 dB-BW</th>
<th>Cross-Pol. Level (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1202</td>
<td>73.31</td>
<td>76.70</td>
<td>-18.64</td>
</tr>
<tr>
<td>0.14</td>
<td>103.86</td>
<td>85.86</td>
<td>-10.81</td>
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<td>0.16</td>
<td>111.66</td>
<td>88.20</td>
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<td>0.18</td>
<td>117.50</td>
<td>89.96</td>
<td>-9.84</td>
</tr>
<tr>
<td>0.2</td>
<td>102.90</td>
<td>91.10</td>
<td>-9.40</td>
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<tr>
<td>0.25</td>
<td>-</td>
<td>96.00</td>
<td>-8.20</td>
</tr>
</tbody>
</table>

Table 7. The effect of the ring height on the radiation characteristics of a semi-spherical resonator antenna with \( \varepsilon_r = 9.5 \), \( \delta = 0.32 \), \( \beta = 0.2404 \), \( a_0 = 0.02 \), and infinite ground plane.
conducting ring. Therefore Fig. 6 shows the E and H plane radiation patterns as well as the cross-polarized field of a cylindrical dielectric antenna. The relative permittivity of the dielectric material is 8.9 while \( d = 3.34 \lambda = 3.3412 \lambda = 1.0 \text{ cm} \) and the circular ground plane diameter \( d_g = 2.594 \lambda = 7.6 \text{ cm} \). It should be noted that the numerical results shown in Fig. 6 are normalized so as to match the experimental data at \( \theta = 0^\circ \) [1]. It is obvious that the computed patterns are in good agreement with their corresponding experimental ones. It should also be noted that the radiation characteristics of the unloaded antenna, as shown in Fig. 6, indicate different E and H plane patterns. Upon loading the antenna by a conducting ring where \( d = 0.28 \lambda, h_l = 0.112 \lambda \) and \( a = 0.02 \lambda \) as shown in Fig. 7, almost identical E and H plane patterns, in the mainlobe region, are observed. As a result a lower cross-polarization level is obtained while the backlobe level is reduced by approximately 8 dB. It has been reported that the ground plane size has a very strong effect on the radiation patterns of dielectric resonator antennas [4]. Therefore, the ground plane diameter is set to be equal to \( 1 \lambda \) and the resulting radiation patterns are shown in Fig. 8. One can easily notice that by using a smaller ground plane the dip of the E plane pattern in the forward direction disappeared. When a conducting ring \( d = 0.28 \lambda, h_l = 0.1202 \lambda \) and \( a = 0.02 \lambda \) is added to the same antenna shown in Fig. 8, improved radiation characteristics are obtained as shown in Fig. 9. The radiation patterns of a hemispherical dielectric resonator antenna above a finite circular ground plane of diameter \( d_g = 1 \lambda \) are also shown in Figs. 10 and 11 without and with a conducting ring \( d = 0.2404 \lambda, h_l = 0.1120 \lambda \) and \( a = 0.02 \lambda \).
respectively. Whereas, Figs. 12 and 13 show the radiation patterns of a hemispherical dielectric resonator antenna of the same antenna parameters as in Figs. 10 and 11, respectively, but with a ground plane of larger diameter i.e. \( d_g = 4 \). Examination of Fig. 12 indicates that the diffractions fields from the edge of the ground plane greatly affect the smoothness of the field patterns in the forward directions and introduce a high backlobe levels. The ripples in the forward direction in the \( E \) plane pattern in Fig. 12 can be reduced in number and level by using a conducting ring as clearly shown in Fig. 13. Furthermore, it is found that the backlobe levels are reduced by approximately 8 dB in this case.
5. Conclusion

Although the symmetry of the $E$ and $H$ plate pattern of dielectric resonator antennas is improved by reducing the ground plane size, it is obvious that by using a beam steering, conducting ring the symmetry of the resulting patterns is greatly improved with a significant reduction in the backlobe level. Furthermore, the proper choice of the ring diameter, height and thickness yield a lower cross-polarized field and enhances the field distribution in the forward direction which in most cases is disturbed by the diffracted rays from the edges of the finite ground plane.

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