INVESTIGATIONS OF A CYLINDRICAL DRA WITH CONTROLLABLE WIDE BANDWIDTHS AND MONOPOLE LIKE OMNI-DIRECTIONAL RADIATION PATTERNS AND BEAM FOCUSING PROPERTIES WHEN IMPLEMENTED IN A CIRCULAR ARRAY

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Mehak Garg
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The Undersigned Faculty Committee Approves the

Thesis of Mehak Garg:

Investigations of a Cylindrical DRA with Controllable Wide Bandwidths and
Monopole Like Omni-Directional Radiation Patterns and Beam Focusing
Properties When Implemented in a Circular Array

Satish K. Sharma, Chair
Department of Electrical and Computer Engineering

Ashkan Ashrafi
Department of Electrical and Computer Engineering

Luciano Demasi
Department of Mechanical Engineering

May 11, 2012
Approval Date
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DEDICATION

I dedicate this thesis work to my mother, who has been the inspiration of my life and always taught me to live to make my dreams come true and never give up. I would not have been able to achieve what I have, without her love and support.
ABSTRACT OF THE THESIS

Investigations of a Cylindrical DRA with Controllable Wide Bandwidths and Monopole Like Omni-Directional Radiation Patterns and Beam Focusing Properties When Implemented in a Circular Array

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Mehak Garg

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The dielectric resonator antenna is actually derived from the dielectric resonator (DR), which was previously used for miniaturization of the active and passive microwave circuit components such as filters and oscillators. The DRs have various advantages such as the compact size, light weight, high radiation efficiency, ease of excitation, high power handling capability, and wideband capability. The coaxial probe and aperture coupled excited cylindrical DRAs have shown wide-bandwidth and broadside radiation patterns. A DRA that radiates pure monopole like patterns is of great interest. The investigations results are presented of a cylindrical dielectric resonator antenna (DRA) with a central air gap and fed in using a coaxial probe connected to a wire monopole. The central air gap and monopole height were determined after parametric studies which offer controllable impedance bandwidth ($S_{11} \leq -10\text{dB}$) between 40% to 67%. The DRA provides monopole like omni-directional radiation patterns with low cross-polarization levels over a wide impedance bandwidth. Prototype DRA with a selected air gap and monopole height was fabricated and experimentally verified. Simulated and measured impedance bandwidths of 67%, and 64%, respectively, were obtained with acceptable peak realized gain over the bandwidth. The simulated and measured radiation patterns also show monopole like omni-directional radiation patterns within the bandwidth.

The investigation results on the beam focusing properties of a dielectric resonator antenna (DRA) array consisting of four-elements on a finite ground plane are also presented. Each of the DRA provides pure omni-directional radiation patterns while possessing 67% impedance bandwidth. One DRA is excited at a time while other DRAs in the array are either short circuited to ground, open circuited, or matched terminated. The best option among the short circuit, open circuit and matched terminated case is selected. The antenna beam provides both horizontal and vertical plane beam coverage with 60% fractional bandwidth. This type of antenna can find applications in Digital Radio Broadcast (DRB), mobile terrestrial communications, and smart adaptive antennas, where switched sector beams are desired. The beam focusing property study was further extended for a case with ground plane having skirt. The array is having six-elements, the center one being fed and one open circuited and rest short circuited to the ground plane surrounding the center element. This work can be carried out further and can be done as a future study of the work presented here.
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CHAPTER 1

INTRODUCTION

Dielectric resonator antenna (DRA) is actually derived from the dielectric resonator (DR), which was previously used for miniaturization of the active and passive microwave circuit components. The DRs have various advantages such as the compact size, light weight, high radiation efficiency, ease of excitation, and high power handling capability [1, 2]. With proper excitation mechanism, these resonators can be used as efficient radiators or antennas instead of energy storage devices. Depending on the resonator shape and feeding methods, various radiating modes can be excited within the DRA radiating element. The coaxial probe and aperture coupled excited cylindrical DRAs have shown wide-bandwidth and broadside radiation patterns [3]. Further a hexagonal DRA showed wideband performance with mix of omni-directional and directional radiation patterns [4]. A coax-fed thin-wire monopole antenna loaded by an annular cylindrical dielectric ring resonator containing steps in its outer radius can cover ultra-wide bandwidth frequency range showing monopole like radiation patterns [5]. In D. Guha et al., the cylindrical DRA is excited using monopole in addition to the fact that monopole itself is acting as another radiator [6]. It has been seen that four-element cylindrical dielectric resonator (CDR) array covers wideband (29%) frequency band and generates low profile monopole-like radiation pattern antenna [7]. Thus, in all above investigations, while effort is made to achieve wideband to ultra-wide bandwidth ranges with monopole like patterns, the antenna design involves exposed monopole antenna structure along with the DRA structures so that monopole kind patterns can be generated intentionally.

This Chapter will introduce DRA, various excitation methods that can be applied to the DRA, detailed analysis of the cylindrical DRA, wideband DRAs, the effect of air gap on the DRA, and arrays of these elements.

1.1 DIELECTRIC RESONATOR ANTENNA

The dielectric resonator (DR) has been primarily used in microwave circuits, such as oscillators and filters. The dielectric constant of DR is usually high. Initially the dielectric
resonator was usually treated as an energy storage device rather than as a radiator. Although open dielectric resonators were found to radiate many years ago, the idea of using the dielectric resonator as an antenna had not been widely accepted until the original paper on the cylindrical dielectric resonator antenna (DRA) was published in 1983 [1]. At that time, it was observed that the frequency range of interest for many systems had gradually progressed upward to the millimeter and near-millimeter range (100-300 GHz). At these frequencies, the conductor loss of metallic antennas becomes severe and the efficiency of the antennas is reduced significantly. Conversely, the only loss for a DRA is that due to the imperfect dielectric material, which can be very small in practice. After the cylindrical DRA had been studied, Long and his colleagues subsequently investigated the rectangular and hemispherical DRAs. The work created the foundation for future investigations of the DRA. Other shapes were also studied, including the triangular, spherical-cap, and cylindrical-ring DRAs. Figure 1.1 [2] shows a photo of various DRAs. It was found that DRAs operating at their fundamental modes radiate like a magnetic dipole, independent of their shapes [1].

A number of excitation methods have been developed. Examples are the coaxial probe, aperture-coupling with a microstrip feedline, aperture-coupling with a coaxial feedline, direct microstrip feedline, co-planar feed, soldered-through probe, slotline, stripline, conformal strip, and dielectric image guide. A photo of the coaxial probe excitation scheme is shown in Figure 1.2 [2], and that of the aperture-coupling excitation scheme is given in Figure 1.3 [2].

As compared to the microstrip antenna, the DRA has a much wider impedance bandwidth (~ 10 % for dielectric constant $\varepsilon_r \sim 10$). This is because the microstrip antenna radiates only through two narrow radiation slots, whereas the DRA radiates through the whole DRA surface except the grounded part. Avoidance of surface waves is another attractive advantage of the DRA over the microstrip antenna. Nevertheless, many characteristics of the DRA and microstrip antenna are common because both of them behave like resonant cavities. For example, since the dielectric wavelength is smaller than the free-space wavelength by a factor of $\frac{1}{\sqrt{\varepsilon_r}}$, both of them can be made smaller in size by increasing $\varepsilon_r$. Moreover, virtually all excitation methods applicable to the microstrip antenna can be used for the DRA.
The DRs have various advantages such as the compact size, light weight, high radiation efficiency, ease of excitation, high power handling capability, and wideband capability [1, 2]. The coaxial probe and aperture coupled excited cylindrical DRAs have shown wide-bandwidth and broadside radiation patterns [3]. A DRA that radiates pure monopole like patterns is of great interest. In S. K. Sharma and P. R. S. Sodhi, the DRA is the radiating element and monopole is feeding it but as an independent radiator i.e. the monopole and DRA two antennas are working together such as in C. Ozzaim [4, 5]. Four-element cylindrical dielectric resonator (CDR) array also covers wide-bandwidth of 29% and generate low profile monopole-like antenna [6]. In the research, we incorporate such a DRA element that was proposed in D. Guha and Y. M. M. Antar which offers a simple DRA
Figure 1.2. Photo of a coaxial probe-fed DRA. (a) Above the ground plane are the coaxial probe and DRA. (b) Below the ground plane is the SMA connector for the coaxial probe. Normally the probe is inside the DRA. Source: K. M. Luk and K. W. Leung, *Dielectric Resonator Antennas*, Baldock, UK: Research Studies Press Ltd., 2003.
Figure 1.3. Photo of an aperture-coupled DRA. (a) Above the ground plane are the circular aperture and DRA. (b) Below the ground plane is the microstrip feedline. Normally the DRA covers the aperture. Source: K. M. Luk and K. W. Leung, *Dielectric Resonator Antennas*, Baldock, UK: Research Studies Press Ltd., 2003.
A simple analysis for the cylindrical DRA was carried out in C. Ozzaim using the magnetic wall model [5]. Figure 1.4 [5] shows the DRA configuration, along with standard cylindrical coordinates.

\[ J_n(X_{np}) = 0, \quad J_n'(X_{np'}) = 0, \quad n = 1, 2, 3, \ldots, \quad p = 1, 2, 3, \ldots, \quad m = 0, 1, 2, 3, \ldots \]

From the separation equation,

\[ k_p^2 + k_z^2 = k^2 = \omega^2 \mu \varepsilon \]  

(1.1)
Figure 1.5. The geometry of cylindrical DRA.
the resonant frequency of the $npm$ mode can be found as follows:

$$f_{npm} = \frac{1}{2\pi a^{2}\mu_{e}} \sqrt{\left\{ \frac{X_{n}^{2}}{X_{n}^{2}} \right\} + \left[ \frac{\pi a^{2}}{2d} (2m + 1) \right]^{2}}$$  \hspace{1cm} (1.2)$$

In practical applications, we are interested in the fundamental (dominant) mode, which has the lowest resonant frequency. It is found that the fundamental mode is the $TM_{110}$ mode, with the resonant frequency given by the following equation, where $X_{11}$ = 1.841.

$$f_{TM_{110}} = \frac{1}{2\pi a^{2}\mu_{e}} \sqrt{\left\{ X_{11}'^{2} \right\} + \left[ \frac{\pi a^{2}}{2d} \right]^{2}}$$  \hspace{1cm} (1.3)$$

1.3 Field Distributions of Cylindrical Dielectric Resonator Antennas

An electric current probe can be used to couple to the electric field lines when it is oriented along the electric field line along the magnetic field lines. Figures 1.6 [1] to 1.9 [1] present the sketches of the electric and magnetic field lines in two perpendicular planes respectively, for the TM01, HEM11, TE01 and HEM12 modes. Notice that these sketches are for the cases without ground plane.


It can be seen from the sketches of the field distributions of the modes that the TM01 and HEM11 modes can be supported if a perfectly conducting plane divides the DR


from the middle of its height, as shown in Figure 1.10 [1], the field distribution will not be disturbed and the resonant frequency will be of the same value as the full size disc. The radiation patterns for these modes are shown in Figure 1.11 [1] and Figure 1.12 [1] for the TM01 and HEM11 modes, respectively. The radiation pattern of the TM01 mode looks like a


quarter wavelength monopole above ground plane. The radiation pattern of the HEM11 looks like the radiation pattern of a half wavelength narrow slot on a ground plane or a half wavelength electric dipole parallel to the ground plane along the electric field lines and of a quarter wavelength above the ground plane.

1.4 Wideband Dielectric Resonator Antennas

Bandwidth enhancement techniques for the DRA have been a popular topic. It was first done in 1989 by Kishk et al., who stacked two different DRAs on top of one another [8]. Since the DRAs had different resonant frequencies, the configuration had a dual-resonance operation, broadening the antenna bandwidth. Sangiovanni et al. employed the stacking method with three DRAs to further increase the antenna bandwidth [9]. Leung et al. introduced an air gap between the stacking and active DRA elements [10]. They used a high-permittivity, low-profile DR as the stacking element, and good results were obtained. Junker et al. analyzed the stacking configuration that employs a conducting or high-\( \varepsilon_r \), loading disk [11]. Simon and Lee used another method in which two parasitic DRs were placed beside the DRA to increase the impedance bandwidth [12]. Alternatively, Leung et al. used the dual-disk method to enhance the bandwidth of the low-profile DRA of very high permittivity [13]. The above methods require extra DR elements. Some bandwidth-enhancement techniques are based on single-DRA configurations. For example, Wong et al. introduced an air gap inside a hemispherical DRA to widen the impedance bandwidth [14]. Ittipiboon et al. performed a similar work with the rectangular DRA [15]. Shum and Luk placed an air gap between the DRA and ground plane to broaden the impedance bandwidth [16]. Leung investigated the case where the air gap inside the DRA is replaced by a conductor [17]. Chen et al. added a dielectric coating to the DRA to increase the impedance bandwidth [18]. Similar work was also carried out by Shum and Luk [19]. Lately, a parasitic conducting patch has been used to increase the impedance bandwidth of the DRA [20, 21]. The new method does not require any extra DR elements or special DRAs and, hence, should facilitate designs of broadband DRAs.

Antennas are the critical component of all the wireless communication systems. They act as a transducer between the guided waves in transmission lines (and waveguides) to the unguided waves in free space. Being a transitional device, antennas have the characteristics of a circuit element and also a radiating element. The important characteristic to consider for an antenna being a circuit element is the impedance matching. The level of impedance matching determines the power delivered to the antenna which has direct consequence on its radiation characteristics especially gain which determines the range. For an antenna to have
good impedance matching, the reflection coefficient $\Gamma$ in Equation 1.4 should be less than a specific value depending on the application. It is usually expressed as a logarithmic quantity called as reflection coefficient magnitude as given in Equation 1.5.

$$\Gamma(\omega) = \frac{Z_L(\omega) - Z_0}{Z_L(\omega) + Z_0}$$

(1.4)

Reflection Coefficient Magnitude = $20 \log |\Gamma| \text{ dB}$

(1.5)

Reflection coefficient magnitude less than (or better than) -10dB is the widely accepted standard as it ensures that 90% of the input power is delivered to the antenna. It can be inferred from Equation 1.4 that the ideal situation would be when the load (antenna) impedance is same as that of line impedance $Z_0$. Otherwise the standing waves are created in the line which hampers the system efficiency depending on the degree of impedance mismatch between line and load. It can also be inferred from Equation 1.4 that the load (antenna) impedance is frequency dependent which makes the reflection coefficient frequency dependent. If the reflection coefficient magnitude is expressed as a positive quantity, it is called as Return Loss (RL) given by Equation 1.6.

$$RL = -20 \log |\Gamma| \text{ (dB)}$$

(1.6)

The same concept of impedance matching can also be expressed by specifying its Voltage Standing Wave Ratio (VSWR) given in Equation 1.7 which is the ratio of the maximum voltage level of standing wave to its minimum value.

$$\text{VSWR} = \frac{V_{\text{max}}}{V_{\text{min}}} = \frac{1+|\Gamma|}{1-|\Gamma|}$$

(1.7)

In ideal case, reflection coefficient should be zero which will translate to a VSWR of 1. Reflection coefficient should be better than 0.33 which will translate to a VSWR better than 2:1 to have the reflection coefficient magnitude better than -10dB.

If an antenna has a reflection coefficient magnitude better than -10dB over a certain range of frequencies with upper cut-off frequency $f_U$ and lower cut-off frequency $f_L$, it is said to be impedance matched in that range and the band of frequencies $(f_U - f_L)$ is called the impedance bandwidth of the antenna under consideration. Narrowband antennas are usually described using a percentage bandwidth, which is calculated as the ratio of the bandwidth which meets the specific VSWR or reflection coefficient requirement to the center frequency of the band as given in Equation 1.8.
Bandwidth percentage = \( \frac{(f_u - f_l)}{f_c} \times 100\% \) \hspace{1cm} (1.8)

where, \( f_u \) is the upper frequency of the band, \( f_l \) is the lower frequency of the band, and \( f_c \) is the center frequency of the band, \( f_c = (f_u + f_l) / 2 \).

Antennas with wider bandwidths are often specified by the ratio between the upper and lower frequencies that define the bandwidth, such as 10:1 for an operating bandwidth between 1 and 10 GHz. Antennas with 40% impedance bandwidth are considered wideband but the frequency ratio is generally used for antennas with bandwidth equal or greater than 2:1 band (i.e. 66% fractional bandwidth).

The frequency response is important because without a reasonable input impedance match, a transmitting system may suffer from severe reflections that could damage other components and waste power, whereas receiving systems will suffer from reduced sensitivity and require additional signal amplification.

### 1.5 AIR GAP EFFECT ON THE DRA

In applying the probe feed method to the DRA, a hole has to be drilled inside the DRA for the probe placement. Since normally the probe does not perfectly fit the hole, an air gap usually exists between the probe and DRA. Junker et al. [22, 23] have studied the air gap effect for the broadside HEM\(_{11}\) (or TM\(_{110}\)) mode of the cylindrical DRA. It was noted that the air gap increases the operation frequency and, in addition, lowers the resonance impedance. The 3-dB bandwidth, however, is not significantly affected by the air gap. An air gap may also exist between the bottom of the DRA and ground plane. The problem was investigated, again, by Junker et al. [23, 24], but the resonant mode of the study was the endfire TM\(_{01}\) mode. It was found that, in general, the resonant frequency increases, the resonance resistance decreases, and the 3-dB impedance bandwidth is significantly broadened, as the size of the air gap becomes larger. Drossos et al. [25] studied the effect of an air gap between the DRA and microstrip substrate for a microstrip-fed DRA. Approximate correction factors were given that helps antenna engineers incorporate the air gap effect into their designs. It was pointed out that the air gap can be used to tune the operating frequency and to obtain a wider bandwidth. It is worth mentioning that the frequency tuning can also be achieved by using a top-loaded conducting plate [26].
1.6 DRA Arrays

As the wireless communication technology is changing rapidly, the antenna designers are looking for high gain compact antenna designs. Also as we have seen that dielectric resonator antenna gives various advantages over a microstrip patch antenna, they are more interesting to explore and look into for these applications. There are various shapes of DRA and various methods to excite it. It offers very wideband response since the antenna gain of a DRA is limited to 5dBi, different types of DRA arrays have been studied for increasing the antenna gain because of its demand in the area of applications.

1.7 Motivation of Research

As discussed in the above topics the importance of wideband antennas in the field of wireless communication applications. The research began with a simple novel cylindrical dielectric resonator antenna that was coaxial fed. As discussed this feeding mechanism at the center offers omnidirectional radiation patterns. There is a monopole that extends from the coaxial probe and goes inside the cylindrical dielectric resonator antenna. The various parametric studies of the various parameters in terms of DRA height, DRA radius, air gap height and radius, monopole height and ground plane variations were carried out. The wideband response with omnidirectional patterns was observed.

The research is further extended to investigate array configuration of the DRAs. The beam focusing properties of a circular array were studied with four such DRA elements. One DRA element was fed at a time and all other elements were short circuited, open circuited and matched terminated respectively. The azimuth and elevation cut patterns were observed and seen that the matched terminated case gave the best results in terms of beam focusing in the desired direction as directed by the DRAs. Such an array configuration can find applications in Digital Radio Broadcast (DRB), mobile terrestrial communications, and smart adaptive antennas, where switched sector beams are desired. Also the beam focusing properties of a circular array with ground plane having a skirt was carried out and studied. The center DRA element was fed and the surrounding elements are short circuited and one being open circuited.
1.8 Simulation Techniques and Experimental Verification Methods

There are different numerical techniques to solve for electric and magnetic fields for any arbitrarily shaped antenna geometry and the software package used for simulation in this thesis work is Ansoft Corporation’s High Frequency Structure Simulator (HFSS) version 11.0 and 13.0, a full-wave Finite Element Method (FEM) tool. There are adequate HFSS licenses and computers with high computation power available in Antenna and Microwave Laboratory (AML) of SDSU to be used for research. HFSS is a highly versatile tool with features to compute and plot scattering parameters, surface current density and vectors, electric and magnetic fields, efficiency, 2D and 3D radiation plots which are useful for analyzing and plotting antenna performance parameters. The structure is modeled in HFSS by specifying its geometry, assigning constituent materials and appropriate boundary conditions. HFSS automatically subdivides the structure into small tetrahedral mesh elements and solves for Maxwell’s equations. The initial size of the tetrahedral comprising the mesh is determined by the desired operating frequency and based on the convergence of results below the set threshold, additional meshing is done to improve accuracy. HFSS, being a frequency domain solver, simulated results will be highly accurate only for frequencies near the set adaptive frequency with adaptive frequency being set at middle or upper end of simulated end.

Once the concept is verified from simulations and design optimized, it is fabricated, measured and the results compared with the simulated data. There is a LPKF Protomat S42 milling machine available at the AML which was used for fabricating the dielectric resonator antenna. The photograph of the LPKF Milling Machine is shown in Figure 1.13.

There is an Anritsu 37269D Vector Network Analyzer (VNA) that goes from 40 MHz – 40 GHz which was used for measuring the antenna’s scattering parameters shown in Fig. 1.13 and Figure 1.14 shows the DRA scattering parameters being measured when it is connected to the VNA. There is a fully anechoic chamber from Orbit/FR which goes from 800 MHz – 18 GHz which was used for measuring the radiation patterns of the fabricated antennas. There are two broadband horn antennas covering the frequency bands 0.8 – 12 GHz and 2 – 32 GHz. Since the measurement frequencies for this work is from 1 – 18 GHz, both the horn antennas had to be used one by one as reference or transmitting antennas in
measurement setup to characterize the Antenna Under Test (AUT) patterns. The radiation pattern recorded from chamber is only the raw data and the Friis transmission equation is used to calculate the antenna’s gain from the knowledge of transmitter gain and the distance between AUT and transmitting horn. The cylindrical DRA being measured for scattering parameter measurement using Vector Network Analyser is as shown in Figure. 1.15. The photograph of the cylindrical dielectric resonator antenna designed on finite ground plane being setup in anechoic chamber for pattern measurement is shown in Figure 1.16 and in Figure 1.17 to measure the XY-cut patterns respectively.

1.9 ORGANIZATION OF THESIS

The remaining thesis is organized as follows. Chapter 2 presents the geometry and results of the novel L-band single dielectric resonator antenna proposed in this thesis work.
Detailed parametric studies, their inferences and the results of experimental verification are reported in this chapter. Chapter 3 presents the results of Beam Focusing Properties of DRA based circular array, their parametric studies and simulation results. Chapter 4 presents the results of the Beam Focusing Properties of DRA based on a circular array having skirted ground plane, the parametric results and simulation results. Chapter 5 presents the conclusions of this thesis work and the scope for future studies.
Figure 1.15. Photograph of the Dielectric Resonator Antenna on a finite ground plane being set up for Return Loss Measurement using the Anritsu Vector Network Analyzer.
Figure 1.16. Photograph of the Dielectric Resonator Antenna on a finite ground plane being setup in Anechoic Chamber for Pattern Measurement in the Azimuth Cut.

Figure 1.17. Photograph of the Dielectric Resonator Antenna on a finite ground plane being setup in anechoic chamber for pattern measurement.
CHAPTER 2

L-BAND SINGLE DIELECTRIC RESONATOR
ANTENNA STUDY AND DESIGN

As discussed in the introductory chapter, the DRs have various advantages such as the compact size, light weight, high radiation efficiency, ease of excitation, and high power handling capability. With proper excitation mechanism, these resonators can be used as efficient radiators or antennas instead of energy storage devices. Depending on the resonator shape and feeding methods, various radiating modes can be excited within the DRA radiating element. This chapter discusses design and study of the single dielectric resonator antenna working in the L-Band (1 GHz to 2 GHz). The cylindrical DRA antenna presented here uses central air gap along with the wire monopole of equal height in the conventional cylindrical DRA which excites the dominant TM_{01δ} mode providing monopole like omni-directional radiation patterns throughout the impedance bandwidth. The proposed antenna is much simpler in architecture than the reported ones generating such a radiation pattern.

The investigations results are presented of a cylindrical dielectric resonator antenna (DRA) with a central air gap and fed in using a coaxial probe connected to a wire monopole. The central air gap and monopole height were determined after parametric studies which offer controllable impedance bandwidth ($S_{11} \leq -10\,\text{dB}$) 40% to 67%. The DRA provides monopole like omni-directional radiation patterns with low cross-polarization levels over a wide impedance bandwidth. Prototype DRA with a selected air gap and monopole height was fabricated and experimentally verified. Simulated and measured impedance bandwidths of 67%, and 64%, respectively, were obtained with acceptable peak realized gain over the bandwidth. The simulated and measured radiation patterns also show monopole like omni-directional radiation patterns within the bandwidth.

2.1 DIELECTRIC RESONATOR ANTENNA GEOMETRY

Figure 2.1(a) shows the 3D view of the proposed antenna placed on top of a finite size square shape ground plane. The cylindrical DRA is located at the center of the ground plane and centrally excited using a 50Ω SMA coaxial probe connected to a wire monopole.
Figure 2.1. (a) The 3D view of the DRA with finite size ground plane, (b) Photograph of the fabricated antenna, and (c) The cross-sectional or side-view of the DRA with SMA feed at the center with an central air gap and monopole.
structure. Diameter of wire monopole is same as the diameter of the center conductor of the SMA connector 1.2mm. The DRA uses a dielectric material of relative dielectric constant of $\varepsilon_r = 10.2$, $\tan\delta = 0.002$ from Rogers Corporation. Photograph of the fabricated antenna is shown in Figure 2.1(b). Figure 2.1 (c) shows the cross-sectional view of this DRA with SMA feed and monopole while showing the important design parameters. There is an air gap between the monopole and the dielectric resonator as shown in the figure. The monopole structure and its location in the DRA are responsible for monopole like omni-directional patterns. The detailed parametric study was performed based on the design parameters. The important parameters varied are: air gap radius ($C_r$), air gap height ($C_h$), monopole height ($M_h$), DRA radius ($D_r$), DRA height ($D_h$), and square ground plane dimensions ($G_x, G_y$).

During the parametric study, while a selected parameter was varied, other parameters were kept invariant. Based on this study, the selected parameters are: $C_r = 2\text{mm}$, $C_h = 40\text{mm}$, $M_h = 40\text{mm}$, $D_r = 12\text{mm}$, $D_h = 70\text{mm}$, and $G_x = G_y = 250\text{mm}$.

### 2.2 Parametric Study Results and Controllable Matching Bandwidth

The effect of varying $C_r$, and $C_h, M_h$ on the impedance matching bandwidth is shown in Figures 2.2 and 2.3, respectively. The effect of other design parameters on impedance matching are also shown following parametric results and table that shows the effect of the air gap height and monopole height on matching bandwidth and thus frequency range. A small variation of 1 mm in $C_r$ can vary the bandwidth drastically as shown in Figure 2.2. The widest bandwidth of 67% ($S_{11} \leq -10\text{dB}$) is achieved when $C_r = 2\text{mm}$.

Also variation in $C_h, M_h$ is a very crucial parameter as observed from Figure 2.3 both of which are changed simultaneously. The $C_h, M_h = 40\text{mm}$ gives the widest matching bandwidth of 67%. The height of air gap and monopole must be kept same for better matching. Effect of different heights for $C_h, M_h$ were also studied but are included here for the sake of brevity. Table 2.1 summarizes the controllable matched frequency range and achievable bandwidth for the variation of heights, between 30mm to 41mm, of both the air gap and monopole. The frequency range and thus the impedance bandwidth from 40% (1.45 to 2.175GHz) to 65% (1.025 to 2.025GHz) can be controlled as the two parameters are
The effect of air-gap radius, $C_r$, on the simulated reflection coefficient magnitude of the DRA.

changed simultaneously from 30mm to 41mm. Since the air gap and monopole heights are to remain same in each case, it is easy to implement for meeting a specific matching bandwidth.

As mentioned earlier, $C_h$, $M_h = 40$mm offers the widest bandwidth of 67%.

The effect of varying $D_h$, $D_r$, $G_x$ and $G_y$ on the impedance matching bandwidth is shown in Figures 2.4, 2.5 and 2.6, respectively. It can be seen the best matching value for $D_h = 70$mm as it gives the maximum matching bandwidth. Also for the $D_r = 12$mm, the -10dB criteria is well satisfied and gives good matching bandwidth as compared to other values. The ground plane dimensions also help in increasing fractional matching bandwidth. As observed if the ground plane dimensions are too small, it affects the matching bandwidth a lot. All the selected values of these parameters are shown in red curves for convenience. The widest bandwidth of 67% ($S_{11} \leq -10$dB) is achieved when these values are selected.
Figure 2.3. The effect of air-gap height, $C_h$ and monopole height, $M_h$ on the simulated reflection coefficient magnitude of the DRA.

The current distributions are presented in Figure 2.7 for the different frequencies within the matching bandwidth. It can be seen that the current distribution is along the DRA height exciting the dominant mode (TM$_{01\delta}$) which supports the omni-directional radiation patterns.

The 3D omni-directional radiation pattern behavior is shown in Figure 2.8. It is evident that throughout the frequency band, the pattern is fairly omni-directional showing monopole kind radiation patterns. The pattern shows null along the DRA height and pattern around the DRA.
Table 2.1. Effect of Ch and Mh on the Controllable Matched Frequency Range and Bandwidth of the DRA

<table>
<thead>
<tr>
<th>Ch, Mh (mm)</th>
<th>Frequency Range (GHz)</th>
<th>Bandwidth (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>30</td>
<td>1.45 – 2.175</td>
<td>40</td>
</tr>
<tr>
<td>31</td>
<td>1.35 – 2.175</td>
<td>47</td>
</tr>
<tr>
<td>32</td>
<td>1.275 – 2.15</td>
<td>51</td>
</tr>
<tr>
<td>33</td>
<td>1.225 – 2.15</td>
<td>55</td>
</tr>
<tr>
<td>34</td>
<td>1.2 – 2.125</td>
<td>56</td>
</tr>
<tr>
<td>35</td>
<td>1.175 – 2.1</td>
<td>56</td>
</tr>
<tr>
<td>36</td>
<td>1.15 – 2.1</td>
<td>58</td>
</tr>
<tr>
<td>37</td>
<td>1.125 – 2.075</td>
<td>59</td>
</tr>
<tr>
<td>38</td>
<td>1.075 – 2.075</td>
<td>63</td>
</tr>
<tr>
<td>39</td>
<td>1.05 – 2.075</td>
<td>65</td>
</tr>
<tr>
<td>40</td>
<td>1.025 – 2.05</td>
<td>67</td>
</tr>
<tr>
<td>41</td>
<td>1.025 – 2.025</td>
<td>65</td>
</tr>
</tbody>
</table>

2.3 FABRICATED RESULTS AND PERFORMANCE RESULTS

The photograph of the fabricated DRA is shown in Figure 2.1(b). The DRA was fabricated using Rogers TMM10 material (εr = 10.2, tanδ = 0.002) of thickness 2.54mm. The DRA discs were glued to achieve the desired DRA height of 70mm. The impedance matching and radiation performance were measured in the Antenna and Microwave Lab (AML) at San Diego State University. The simulated and measured impedance bandwidth (S11 ≤ -10dB) of the final DRA is around 67% (1.05GHz to 2.05GHz) and 64% (1.14GHz to 2.215GHz), respectively, as shown in Figure 2.9. The difference in frequency range and matching bandwidths is attributed to fabrication error which may be also in terms of air gap and monopole height. Also the peak gain versus frequency variation can be seen in Figure 2.10. It can be observed that the peak gain varies between 0.4dBi and 4dBi and always remains positive. Also the simulated and measured peak gain values are following the same trend with a gain disagreement of around 1 dB except towards lower and upper end of the frequencies.
Figure 2.4. The effect of DRA height, \( D_h \), on the simulated reflection coefficient magnitude of the DRA.

Figure 2.11 shows the simulated and measured XZ-, YZ- and XY-cut patterns of the DRA at 1.1GHz, 1.52GHz, and 1.905GHz. It can be seen that, the elevation (XZ- and YZ-cuts) plane and azimuth (XY-cut) plane patterns show monopole like radiation patterns. Further, right from the start of the frequency band, the antenna radiation pattern is omni-directional in nature. The patterns are very symmetric throughout the frequency band. The simulated and measured co-polarization pattern components in all cases agree well. However, the measured cross-polarization levels in XZ and YZ-cuts are higher than the simulated level which is attributed to the alignment error of the antenna on the antenna under test mount. The measured cross-polarization level in XY-cut matches well with the simulated cross-polarization. The XY-cut or azimuthal plane pattern is also included here because it
suggests that pattern is invariant in azimuthal plane which is desired in many applications such as the direction finding system.

The chapter presented the investigation results of a wideband cylindrical DRA excited using a coaxial probe connected to a wire monopole on the finite ground plane to achieve monopole kind omni-directional radiation patterns. Controllable matching bandwidths from 40% to 67% can be obtained by controlling air gap and monopole height from 30mm to 41mm. Fabricated DRA’s impedance matching and radiation patterns are in reasonable agreement with the simulated antenna performance. This antenna operates around L-band frequency which covers several satellite communication bands and direction finding systems. In the next section, the beam focusing properties of a circular array based on the single element dielectric resonator antenna is presented. The single element DRA presented
Figure 2.6. The effect of ground plane dimensions, $G_x$ and $G_y$, on the simulated reflection coefficient magnitude of the DRA.

in this chapter is further implemented in a circular array and its beam focusing properties are studied. The work performed was also submitted to be published in *IEEE Transactions on Antennas and Propagation* and was accepted as a paper to be presented at a conference in April 2012 [27].
Figure 2.7. The current distribution plots showing TM_{015} mode excitation at various frequency points within the matching bandwidth (a) 1.1GHz, (b) 1.5GHz, (c) and 1.9GHz.
Figure 2.8. The omni-directional 3-D radiation patterns at (a) 1.1GHz, (b) 1.5GHz, (c) and 1.9GHz.
Figure 2.9. Comparison of the simulated and measured magnitude of the reflection coefficients, $S_{11}$, versus frequency for the proposed DRA.
Figure 2.10. The simulated and measured peak gain versus frequency for the proposed DRA.
Figure 2.11. The simulated and measured radiation patterns at 1.1GHz, 1.52GHz, and 1.905GHz for the proposed DRA: (a-c): YZ-cuts, (d-f): XZ-cuts and (g-i): XY-cuts respectively.
CHAPTER 3

BEAM FOCUSING PROPERTIES OF DRA BASED CIRCULAR ARRAY

This chapter discusses the beam focusing properties of a cylindrical dielectric resonator array antenna that has been extensively investigated by employing different coaxial probe feeding mechanism on a finite ground plane. The presented DRA element is monopole fed and has an air gap between the monopole and the DR. The advantage of this design over others reported is that it is a complete dielectric resonator that is radiating and monopole is to feed it. The air gap between the monopole and the DR helps improving the bandwidth as it can be seen from the results presented. The investigation results on the beam focusing properties of a dielectric resonator array (DRA) antenna consisting of four-elements on a finite ground plane, by having different exciting schemes as shown in Figure 3.1. The excitation of one element at a time and having all other elements either short circuited to ground, open circuited, and matched terminated provides enhanced horizontal and vertical plane peak directivities, and a reduced 3 dB beamwidth for the horizontal pattern.

3.1 PARAMETRIC STUDY RESULTS

Figure 3.1 shows a circular array of dielectric resonator antennas, having a radius of $D_r = 1.2\lambda = 12\text{mm}$, height $D_h = 3.5\lambda = 70\text{mm}$. The air gap dimensions has a radius of $C_r = 0.1\lambda = 2\text{mm}$, height $C_h = 2\lambda = 40\text{mm}$ and monopole height of $M_h = 2\lambda = 40\text{mm}$. The array radius is $R_a = 3\lambda = 60\text{mm}$, and is placed on a finite circular ground plane of radius $R_g = 17.5\lambda = 350\text{mm}$. The monopole #1 is fed using a 50 SMA connector from below the ground, and the remaining elements are shorted to the ground. The number of monopole elements selected for this study are $N = 4, 6, 8, 10$, and the reflection coefficient plot is shown in Figure 3.2 and the peak realized gain versus frequency plot in Figure 3.3 for the number of elements case. It can be seen that the reflection coefficient plot is meeting the criteria of $S_{11} \leq -10\text{dB}$ and also giving the best value of gain for four element case as compared to other cases. This is the reason four elements are considered for further study and other cases.
Figures 3.4 and 3.5 show the reflection coefficient magnitude plots for the parametric study done for the ground plane radius, $R_g$ and the array radius, $R_a$. The peak realized gain plots for these parameters are seen in Figures 3.6 and 3.7, respectively. The array radius is one of the crucial parameters that affect the bandwidth covered by the array and thus its overall beamwidth. With this excitation, the ground plane size, $R_g$, was varied from 250 to 350mm, keeping other parameters constant. The results are not affected significantly, but in Figure 3.4 the best bandwidth covered is for $R_g = 350$mm. The array radius was varied from $R_a = 35$ to 75mm, keeping other parameters constant. The bandwidth improves with $R_a$. The best results is for $R_a = 60$mm. The effect of $R_a$ by changing the number of elements were also studied, but were inferior to the four-element array.

### 3.2 Four Element DRA with Different Excitation Schemes

The selected array is a four-element DRA, with $R_a = 3\lambda = 60$mm, on a ground plane of radius $R_g = 3.5\lambda = 350$ mm, as in Figure 3.1. One element is selected for excitation at a
time, and other three are shorted to the ground. The excited element behaves as the feed for a cylindrical reflector simulated by the shorted elements. The proposed three symmetrical excitation schemes are when only one element is excited at a time (#1, 2, 3, and 4) and other three are short circuited to ground, open circuited, and matched terminated. For these symmetric excitations, the array beam remains along the X-direction. By sequentially changing the excited elements, the array beam can be rotated. For instance, connecting the excitation from element #1 to #2 will rotate the array beam by 90°.

### 3.2.1 Short Circuited Case

The reflection coefficient plot and peak realized gain plot are presented in Figure 3.8 and 3.9, respectively, when only one element is excited at a time (#1, 2, 3, and 4) and other three are short circuited to ground case. The matching bandwidth achieved for the best case with match terminating the other elements was 64% meeting the $S_{11} < -10$dB criteria from...
Figure 3.3. The plot of peak realized gain versus frequency for number of elements variation.

1.025GHz to 2GHz. Also the peak gain variation was seen from 2dBi to 9dBi for the array. Figure 3.10 shows the plot of xy-plane cut versus frequency when fed one at a time and all other elements are shorted to ground plane. It can be seen that the beam shifts as we change the fed element from 1 to 2 and so on. Also Figures 3.11, 3.12, 3.13 and 3.14 show the plots of yz-plane cut versus frequency when element #1, 2, 3 and 4 is fed, respectively, and all other elements are shorted to the ground plane. It can be seen that the beam is focused at around 54degrees with fairly good 3dB beamwidth values. of 140deg, 110deg, 90deg, 80deg, 80deg at 1.1GHz, 1.3GHz, 1.5GHz, 1.7GHz, 1.9GHz respectively.

### 3.2.2 Open Circuited Case

The reflection coefficient plot and peak realized gain plot are presented in Figure 3.15 and 3.16, respectively, when only one element is excited at a time (#1, 2, 3, and 4) and other three are open circuited to the ground plane. The matching bandwidth achieved for the best
Figure 3.4. The plot of reflection coefficient magnitude versus frequency by varying Rg.

case with match terminating the other elements was 44% meeting the $S_{11} < -10\text{dB}$ criteria from 1.05GHz to 1.65GHz. Also the peak gain variation was seen from 4dBi to 9dBi for the array. Figure 3.17 shows the plot of xy-plane cut versus frequency when fed one at a time and all other elements are open circuited to ground plane. It can be seen that the beam shift as we change the fed element from 1 to 2 and so on. Also Figures 3.18, 3.19, 3.20 and 3.21 show the plots of yz-plane cut versus frequency when element #1, 2, 3 and 4, respectively, is fed and all other elements are open circuited to ground plane. It can be seen that the beam is focused at around 54degrees with fairly good 3dB beamwidth values of 135deg, 100deg, 90deg, 80deg, 80deg at 1.1GHz, 1.3GHz, 1.5GHz, 1.7GHz, 1.9GHz respectively.

3.2.3 Matched Terminated Case

The reflection coefficient plot and peak realized gain plot are presented in Figure 3.22 and 3.23 respectively when only one element is excited at a time (#1, 2, 3, and 4) and other
Figure 3.5. The plot of reflection coefficient magnitude versus frequency by varying Ra.

three are matched terminated case. The matching bandwidth achieved for the best case with match terminating the other elements was 58% meeting the S11 < -10dB criteria from 1.05GHz to 1.9GHz. Also the peak gain variation was seen from 3dBi to 6.5dBi for the array. Figure 3.24 shows the plot of xy-plane cut versus frequency when fed one at a time and all other elements are matched terminated. It can be seen the beam shift as we change the fed element from 1 to 2 and so on. Also Figures 3.25, 3.26, 3.27 and 3.28 show the plots of yz-plane cut versus frequency when element #1, 2, 3 and 4 is fed and all other elements are matched terminated respectively. It can be seen that the beam is focused at around 54degrees with good 3dB beamwidth values of 150deg, 125deg, 110deg, 90deg, 90deg at 1.1GHz, 1.3GHz, 1.5GHz, 1.7GHz, 1.9GHz respectively.

It has been observed from the elevation plane cuts that the beam focusing is best when matched terminated as compared to when short circuited and open circuited. The beam
variation is less and the beam is more focused when matched terminated. It has been observed from the elevation plane cuts that the beam focusing is best when matched terminated as compared to when short circuited and open circuited. The beam variation is less and the beam is more focused when matched terminated. Also it is evident from the azimuthal and elevation plane cuts that the beam variation is less in this case as compared to the short and open circuited case.

3.3 MIMO Study Results

It was seen from the above results that the matched terminated case gave the best results in terms of reflection coefficient plot and peak realized gain plot. So this case was further taken and MIMO study was performed on it i.e. if this structure was to put in as a MIMO system, will it work or not. For this we would need the knowledge of Envelope Correlation and Mutual Coupling. In a rich multipath environment, the theoretical capacity of

![Figure 3.6. The plot of peak realized gain versus frequency for $R_g$ variation.](image-url)
a MIMO system increases linearly with the number of antenna elements $N$ in an $(N, N)$ MIMO system. However, in a more practical MIMO system, the capacity is reduced due to correlation between the signals in the receiver. Therefore, the correlation between the signals received from the different antenna elements is an important parameter in a MIMO system. As long as the envelope correlation $\rho_e$ is less than 0.5, the diversity gain could be obtained in a mobile phone. The envelope correlation between antennas $i$ and $j$ in the $(N, N)$ MIMO antenna system can be calculated. In the case of a $(4, 4)$ MIMO system, with $N=4$ antennas at both ends, the envelope correlation between antenna $i=1$ and $j=2$ could be calculated as presented in Equation 3.1.

$$\rho_{e(i,j,N)} = \frac{\left|\sum_{n=1}^{N} s_{i,n}^* s_{n,j}\right|^2}{\prod_{k=l}^{N} [1 - \sum_{n=1}^{N} s_{i,n}^* s_{n,j}]}$$ (3.1)

Figure 3.7. The plot of peak realized gain versus frequency for $R_a$ variation.

In this case, correlation between antenna elements i.e. DRA #1 and 2 is presented where we have 4 antennas array system. The formula was thus derived using the above
equation for envelope correlation and was found to be as follows after simplification of the complex numbers.

$$\rho_{e\ (1,2,4)} = \frac{|S_{11}^*S_{12} + S_{12}^*S_{22} + S_{13}^*S_{32} + S_{14}^*S_{42}|^2}{\left(1 - (|S_{11}|^2 + |S_{21}|^2 + |S_{31}|^2 + |S_{41}|^2)\right)\left(1 - (|S_{12}|^2 + |S_{22}|^2 + |S_{32}|^2 + |S_{42}|^2)\right)}$$

As it needs the real and imaginary values of the S-Parameters in the above formula for computation of the envelope correlation between two antennas, these values were taken from HFSS. The real and imaginary plots were separately taken as it varies with frequency as well. The mutual coupling plot is shown below in Figure 3.29 and it can be seen that is below -11dB value throughout the frequency band. The envelope correlation is also presented in Figure 3.30 and it depicts a good MIMO system as the value of envelope correlation is far below 0.5 as can be seen.

Figure 3.8. The plot of reflection coefficient magnitude versus frequency when fed one at a time and all other elements are shorted to ground plane.
The chapter presented the investigation results of beam focusing properties of a DRA based circular array. The beam focusing was achieved by feeding one of the elements and match terminating all the others. The results are presented for the number of elements variation and then selecting from short circuiting, open circuiting and matched termination. The matching bandwidth achieved for the best case with match terminating the other elements in terms of the beam focusing property was 58% meeting the S11 < -10dB criteria from 1.05GHz to 1.9GHz. Also the peak gain variation was seen from 3dBi to 6.5dBi for the array. MIMO study is also presented on the final array case and the mutual coupling plot and envelope correlation plot shows that it exhibits good MIMO behavior which can be used in satellite applications. The work performed to study the beam focusing properties of the circular array was also submitted and accepted for publication as [28].

Figure 3.9. The plot of peak realized gain versus frequency when fed one at a time and all other elements are shorted to ground plane.
Figure 3.10. The plot of xy-plane cut versus frequency when fed one at a time and all other elements are shorted to ground plane.
Figure 3.11. The plot of yz-plane cut versus frequency when element #1 is fed and all other elements are shorted to ground plane.
Figure 3.12. The plot of yz-plane cut versus frequency when element #2 is fed and all other elements are shorted to ground plane.
Figure 3.13. The plot of yz-plane cut versus frequency when element #3 is fed and all other elements are shorted to ground plane.
Figure 3.14. The plot of yz-plane cut versus frequency when element #4 is fed and all other elements are shorted to ground plane.
Figure 3.15. The plot of reflection coefficient magnitude versus frequency when fed one at a time and all other elements are open circuited.
Figure 3.16. The plot of peak realized gain versus frequency when fed one at a time and all other elements are open circuited.
Figure 3.17. The plot of xy-plane cut versus frequency when fed one at a time and all other elements are open circuited.
Figure 3.18. The plot of yz- plane cut versus frequency when element #1 is fed and all other elements are open circuited.
Figure 3.19. The plot of yz-plane cut versus frequency when element #2 is fed and all other elements are open circuited.
Figure 3.20. The plot of $yz$-plane cut versus frequency when element #3 is fed and all other elements are open circuited.
Figure 3.21. The plot of yz-plane cut versus frequency when element #4 is fed and all other elements are open circuited.
Figure 3.22. The plot of reflection coefficient magnitude versus frequency when fed one at a time and all other elements are matched terminated.
Figure 3.23. The plot of peak realized gain versus frequency when fed one at a time and all other elements are matched terminated.
Figure 3.24. The plot of xy-plane cut versus frequency when fed one at a time and all other elements are matched terminated.
Figure 3.25. The plot of yz-plane cut versus frequency when element #1 is fed and all other elements are matched terminated.
Figure 3.26. The plot of yz-plane cut versus frequency when element #2 is fed and all other elements are matched terminated.
Figure 3.27. The plot of $yz$-plane cut versus frequency when element #3 is fed and all other elements are matched terminated.
Figure 3.28. The plot of yz-plane cut versus frequency when element #4 is fed and all other elements are matched terminated.
Figure 3.29. The plot of mutual coupling between the array elements.
Figure 3.30. The plot of envelope correlation between array elements 1 and 2 in a 4 element array MIMO system.
CHAPTER 4

BEAM FOCUSING PROPERTIES OF DRA BASED
CIRCULAR ARRAY HAVING SKIRTED GROUND
PLANE

This chapter discusses the beam focusing properties of a cylindrical dielectric resonator array antenna having skirted ground plane that has been extensively investigated by employing different coaxial probe feeding mechanism on a finite ground plane. The presented DRA element is the same as described in Chapter 2 and implemented in Chapter 3. Typically, a limited ground surface will introduce a change in the principal direction of the radiation. Unlike the ideal case, the radiation maximum of a DRA or DRA array on a finite ground is elevated above the horizontal. As a consequence, the terrestrial coverage of the radiator will be severely degraded when the antenna is mounted at height above the true ground as in this case. The simple solution to reduce main lobe elevation is to extend the horizontal ground area. This method has obvious practical limits, so an alternative option is to attach a sleeve or skirt on the perimeter of the ground plane. This effectively enlarges the ground surface, without increasing the antenna’s horizontal area. The basic structure of a sleeved DRA that surrounds the source coaxial cable is shown in Figure 4.1. This chapter describes a skirted parasitic array that exhibits radiation controlled by the switching of parasitic antenna elements.

The investigation results on the beam focusing properties of a dielectric resonator antenna (DRA) array consisting of six-elements on a finite ground plane having skirt are shown in this chapter. Each of the DRA provides pure omni-directional radiation patterns while possessing 67% impedance bandwidth. The center DRA is excited, one DRA element is open circuited and other four elements in the array are short circuited to the ground plane. The five elements are surrounding the center element in a circular fashion. The short circuited elements form a passive reflector and beam is pushed in the direction of the open circuited element. As will be shown in Figure 4.2, DRA #1 is fed by a coaxial
Figure 4.1. Geometry of an N-element DRA on a finite skirted ground plane fed using a SMA probe.

SMA line and DRA #2 is open circuited and other elements i.e. DRA #3, 4, 5, 6 are all short circuited to the ground plane.

4.1 PARAMETRIC STUDY RESULTS

Figure 4.1. shows a circular array of dielectric resonator antennas, having a radius of $D_r = 0.06\lambda = 12\text{mm}$, height $D_h = 0.35\lambda = 70\text{mm}$. The air gap dimensions has a radius of $C_r =$
0.01\lambda = 2\text{mm}, height \( C_h = 0.2\lambda = 40\text{mm} \) and monopole height of \( M_h = 0.2\lambda = 40\text{mm} \). The array radius is \( d_r = 0.45\lambda = 90\text{mm} \), and is placed on a finite circular ground plane of radius \( G_r = 0.575\lambda = 115\text{mm} \), and having skirt of height \( G_h = 0.65\lambda = 130\text{mm} \). The DRA #1 is fed using a 50 SMA connector from below the ground, and DRA #2 is open circuited and the remaining elements are shorted to the ground as seen in Figure 4.2.

Figure 4.3 shows the reflection coefficient plots for the parametric study done for the ground plane radius and the array radius. The array radius is one of the crucial parameters that affect the bandwidth covered by the array and thus its overall beamwidth. With this excitation, the ground plane size, \( G_r \), was varied from 115 to 515mm, keeping other parameters constant. The results are not affected significantly, but in Figure 4.3 the best bandwidth covered is for \( G_r = 115\text{mm} \). The ground plane height was also varied from 30mm
to 230mm and did not show much change in the reflection coefficient plot as seen in Figure 4.4, but the best value was selected to be 230mm. The array radius was varied from $d_y = 25$ to 100mm, keeping other parameters constant as shown in Figure 4.5. The bandwidth improves with $d_y$. The best results is for $d_y = 90$mm.

### 4.2 Six Element DRA with Skirted Ground Plane

The array is a six-element DRA, with $d_y = 0.45\lambda = 90$mm, on a ground plane of radius $G_r = 0.575\lambda = 115$ mm, ground plane height $G_h = 0.65\lambda$ i.e. skirt height as in Figure 4.1. The center element is selected for excitation, and one is open circuited and other four are shorted circuited to the ground. Each of the DRA provides pure omni-directional radiation patterns while possessing 67% impedance bandwidth. The five elements are surrounding the center element in a circular fashion. The short circuited elements form a passive reflector and beam is pushed in a direction because of such arrangement. As shown in Figure 4.2. DRA #1 is fed
Figure 4.4. The plot of reflection coefficient magnitude versus frequency by varying Gh.

by a coaxial SMA line and DRA #2 is open circuited and other elements i.e. DRA #3, 4, 5, 6 are all short circuited to the ground plane.

Figure 4.6 and 4.7 show the reflection coefficient plot and peak realized gain plot respectively for the array. The reflection coefficient plot shows a matching bandwidth of 43% meeting the S11 < -10dB criteria from 1.165GHz 1.795GHz. Also the peak realized gain plot shows a variation from 3dBi to 6.8dBi throughout the bandwidth as can be seen from the figure.

Figure 4.8 and 4.9 show plots of the radiation patterns at 1.3GHz, 1.5GHz and 1.7GHz at Theta = 90deg and Phi = 90deg respectively. Figure 4.8 shows that the beam is being pushed in the direction of the open circuited element by the other short circuited ones. Also the beam focusing can be seen in Figure 4.9 at an angle of around -45deg. This geometry can be further studied and explored for proper beam focusing in the direction we want and can be further optimized.
Figure 4.5. The plot of reflection coefficient magnitude versus frequency by varying dy.

The chapter presented the investigation results of beam focusing properties of a DRA based circular array having skirted ground plane. The array is having six elements where the center element is fed by a coaxial probe and one element is open circuited and rest four elements surrounding the center element are short circuited to the ground plane having skirt. The parametric results are also shown for the array radius, skirt height variation and ground plane radius variation on the matching bandwidth. The reflection coefficient plot shows a matching bandwidth of 43% meeting the S11 < -10dB criteria from 1.165GHz to 1.795GHz.

Also the peak realized gain plot shows a variation from 3dBi to 6.8dBi throughout the bandwidth. This array cannot be realized for the beam focusing property and thus cannot be used for this application.
Figure 4.6. The plot of reflection coefficient magnitude versus frequency.
Figure 4.7. The plot of peak realized gain versus frequency.
Figure 4.8. The plot of radiation pattern at 1.3GHz, 1.5GHz and 1.7GHz when Theta = 90deg.
Figure 4.9. The plot of radiation pattern at 1.3GHz, 1.5GHz and 1.7GHz when Phi = 90deg.
CHAPTER 5

CONCLUSIONS AND FUTURE STUDY

Dielectric Resonator Antennas have various advantages as discussed such as compact size, light weight, high radiation efficiency, ease of excitation, high power handling capability, and wideband capability. The coaxial probe and aperture coupled excited cylindrical DRAs have shown wide-bandwidth and broadside radiation patterns. A DRA that radiates pure monopole like patterns is of great interest and one such design is studied in this thesis.

Chapter 1 summarized the research goals, analysis and theory of cylindrical dielectric resonator antenna, wideband DRAs, air gap effect on the DRA, followed by DRA arrays.

Chapter 2 introduced the novel L band single DRA study and design. The DRA design geometry with coaxial fed SMA and an air gap that shows improved performance in impedance matching and peak gain with omnidirectional radiation patterns. The parametric results are presented in terms of DRA height, DRA radius, air gap height, air gap radius and monopole height with ground plane size variations as well. The parametric studies showed that the matching bandwidth could be controlled with the help of air gap height and monopole height. Also these two parameters when have the same value give the best result in terms of the matching bandwidth and the radiation patterns. The investigations results are presented of a cylindrical dielectric resonator antenna (DRA) with a central air gap and fed in using a coaxial probe connected to a wire monopole. The central air gap and monopole height were determined after parametric studies which offer controllable impedance bandwidth ($S_{11} \leq -10\text{dB}$) between 40% to 67%. The DRA provides monopole like omni-directional radiation patterns with low cross-polarization levels over a wide impedance bandwidth. Prototype DRA with a selected air gap and monopole height was fabricated and experimentally verified. Simulated and measured impedance bandwidths of 67%, and 64%, respectively, were obtained with acceptable peak realized gain over the bandwidth. The simulated and measured radiation patterns also show monopole like omni-directional radiation patterns within the bandwidth.
Chapter 3 introduced the investigation results on the beam focusing properties of a dielectric resonator antenna (DRA) array consisting of four-elements on a finite ground plane. Each of the DRA provides pure omni-directional radiation patterns while possessing 67% impedance bandwidth. One DRA is excited at a time while other DRAs in the array are either short circuited to ground, open circuited, or matched terminated. The best option among the short circuit, open circuit and matched terminated case is selected. The antenna beam provides both horizontal and vertical plane beam coverage with 60% fractional bandwidth. This type of antenna can find applications in Digital Radio Broadcast (DRB), mobile terrestrial communications, and smart adaptive antennas, where switched sector beams are desired.

Chapter 4 introduced the investigation results on the beam focusing properties of a dielectric resonator antenna (DRA) array consisting of six-elements on a finite ground plane having skirt. Each of the DRA provides pure omni-directional radiation patterns while possessing 67% impedance bandwidth. The center DRA is excited, one DRA element is open circuited and other four elements in the array are short circuited to the ground plane. The five elements are surrounding the center element in a circular fashion. The short circuited elements form a passive reflector and beam is pushed in the direction of the open circuited element.
REFERENCES


