CHAPTER 3

DIELECTRIC RESONATOR ANTENNA
- THEORY AND FABRICATION

This chapter describes the fundamental concepts and theory of the dielectric resonator antenna. The methodology for fabricating DRA and that for the experimental measurements of various parameters of the antenna such as return loss, gain, radiation patterns etc. are also explained. The method of characterization of the DR and the basic facilities used for the measurement of antenna parameters are also highlighted.

3.1 DIELECTRIC RESONATOR ANTENNAS

The DRA is an open resonating structure, fabricated from a low loss microwave dielectric material. Dielectric resonators (DR’s) have proved themselves to be ideal candidates for antenna applications by virtue of their high radiation efficiency, flexible feed arrangement, simple geometry, small size and the ability to produce different radiation pattern using different modes[1]. Feeding techniques like probe feed, aperture slot, microstrip line and coplanar line can be used with the DRAs, which enables them for integration with microwave printed technology.

Additionally, DRA’s avoid some limitations of the patch antenna including the high conductor losses at millimeter-wave frequencies, sensitivity to tolerances, and narrow bandwidth. DRA’s of cylindrical, hemispherical and rectangular shapes are most widely used and investigated. The rectangular shape is much easier to fabricate and one or more dimensional parameters are available as additional degrees of freedom for the design [2]. Impedance bandwidth varies over a wide range with resonator parameters. It
can be as small as a few percent with high $\varepsilon_r$ material or over 20% with small $\varepsilon_r$ in conjunction with certain geometries and resonant modes. Different far field radiation patterns are supported. For a given DRA geometry, the radiation pattern can be made to change by exciting different modes.

Systematic experimental investigations on dielectric resonator antennas (DRA’s) were first carried out by Long et al. [3],[4]–[6]. Since then, theoretical and experimental investigations have been reported by many investigators on DRA’s of various shapes such as spherical, cylindrical (or cylindrical ring), rectangular, etc. (e.g., [3],[7]–[12]). DRAs with various other shapes are also reported in different literature.

3.2 ADVANTAGES OF DRAs

DRAs offer a high degree of flexibility and versatility over a wide frequency range, allowing for designers to suit many requirements. DRAs offer the following advantages:

- DRAs come in simple geometries like circular cylinder, hemisphere, rectangular etc. are readily available and can be easily fabricated.

- The DRA size is proportional to $\frac{\lambda_0}{\sqrt{\varepsilon_r}}$, where $\lambda_0$ is the wavelength at resonant frequency and $\varepsilon_r$ is the dielectric constant of the DR. Thus for the same frequency there is a natural reduction in size, compared with their conventional counterparts like microstrip antennas. Also, different values of $\varepsilon_r$ (ranging from 4 to 100) can be used, thus allowing the designer the flexibility in controlling the size and bandwidth.
Depending on the resonator shape, various modes can be excited within the DRA element. These modes can produce different radiation patterns for various coverage requirements. Also, the Q-factor of some of these modes will depend on the aspect ratio of the DRA, thus allowing one more degree of flexibility in the design.

Many of the existing feeding schemes can be used (slots, probes, microstrip, coplanar waveguides, dielectric image guide, etc.). This makes them easy to integrate with existing technologies.

Compared with the microstrip antenna, DRA has a much wider impedance bandwidth. This is because the microstrip antenna radiates only through two narrow radiation slots, whereas the DRA radiates through the whole antenna surface except the grounded part. Moreover the operating bandwidth of a DRA can be varied by suitably choosing the dielectric constant of the resonator material and its dimensions.

DRAs have been designed to operate over a wide frequency range (1 GHz to 44 GHz) compared with other antennas existing in the literature.

DRAs have a high dielectric strength and hence higher power handling capacity. Moreover the temperature-stable ceramics enable the antenna to operate in a wide temperature range.

There is no inherent conductor loss for a DRA. High radiation efficiency is thus possible in case of DR antennas. It is especially attractive for high frequency millimeter wave applications, where the loss from metallic antennas can be high.
3.3 RECTANGULAR DRA

Rectangular DRAs offer practical advantages over cylindrical and spherical shape. For example, the mode degeneracy can be avoided in the case of rectangular DRA’s by properly choosing the three dimensions of the resonator. It may be noted that mode degeneracy always exists in the case of a spherical DRA [13] and in the case of hybrid modes of a cylindrical DRA [14]. The mode degeneracy can enhance the cross-polar levels of an antenna, thus limiting its performance. Further, for a given resonant frequency, two aspect ratios of a rectangular DRA (height/length and width/length) can be chosen independently. Since the bandwidth of a DRA also depends on its aspect ratio(s), a rectangular-shaped DRA provides more flexibility in terms of bandwidth control [3].

A rectangular DRA support two type modes, TM and TE, but TM modes have never been observed experimentally [3]. Therefore the existence of TM modes appears to be doubtful. Figure 3.3.1 shows a rectangular DRA with the corresponding coordinate system. The resonant modes can be TE to either dimension, denoted as TE\(x\), TE\(y\), or TE\(z\). A rectangular DR has three independent dimensions. The modes of a DR can therefore, be TE to any of three dimensions.

Referring to the DR and co-ordinates system shown in figure 3.3.1, the modes with lowest order indexes are \(TE^z_{111}\), \(TE^y_{111}\) and \(TE^x_{111}\) [15]. If the dimensions of the DR are such that \(a > b > d\), the modes in the order of increasing resonant frequency are \(TE_{111}\), \(TE^y_{101}\) and \(TE^x_{011}\). The analysis of all the modes is similar. For example, for TE mode,
the analysis for the field components inside the resonator can be done from the directed magnetic potential $\Phi^h$ [3],[16].

Figure 3.3.1: Isolated rectangular DRA.

\[ H_z = \frac{\left( k_x^2 + k_z^2 \right)}{j\omega\mu_0} A\cos(k_x)x\cos(k_y)y\cos(k_z)z \]

\[ H_x = \frac{k_x k_z}{j\omega\mu_0} A\sin(k_x)x\cos(k_y)y\sin(k_z)z \]

\[ H_y = \frac{k_x k_z}{j\omega\mu_0} A\cos(k_x)x\sin(k_y)y\sin(k_z)z \]

\[ E_x = Ak_y\cos(k_x)x\sin(k_y)y\cos(k_z)z \]

\[ E_y = -Ak_y\sin(k_x)x\cos(k_y)y\cos(k_z)z \]
where \( A \) is an arbitrary constant and \( k_x, k_y, \) and \( k_z \) denote the wavenumbers along the x, y, and z directions, respectively, inside the DR.

\[
k_x^2 + k_y^2 + k_z^2 = \varepsilon_r k_0^2
\]  \hspace{1cm} \text{----------------------(6)}

\[
k_z \tan \left( \frac{k_z d}{2} \right) \sqrt{(\varepsilon_r - 1) k_0^2 - k_z^2}
\]  \hspace{1cm} \text{----------------------(7)}

The dimensions of the radiating portion of the DR were determined using the equation (7) developed for the dielectric waveguide model (DWM) [15] for a rectangular resonator in free-space [3].

where

\[
Q = \frac{2\omega W_s}{P_{rad}}
\]  \hspace{1cm} \text{----------------------(8)}

\[
k_x = \frac{\Pi}{a}, k_y = \frac{\Pi}{b}, k_z = \frac{2\Pi \varepsilon_r}{C}, C = 3 \times 10^8 m / S
\]  \hspace{1cm} \text{----------------------(9)}

Figure shows the rectangular resonator with length a, breadth b and height d. Resonances can occur at the following frequencies

\[
f_{\text{res}} = \frac{1}{2\sqrt{\varepsilon \mu}} \sqrt{\left( \frac{m}{a} \right)^2 + \left( \frac{n}{b} \right)^2 + \left( \frac{p}{d} \right)^2}
\]  \hspace{1cm} \text{----------------------(10)}

where \( \varepsilon \) is the permittivity, \( \mu \) is the material permeability, and \( m, n \) and \( p \) are integers. In this configuration, TE\(^x\text{011}\) mode is the dominant mode, because it occurs at the lowest frequency at which a cavity resonance can exist. From equation (10) it can be seen that the frequency at which this dominant resonant mode can exists (the cutoff frequency) is inversely proportional to the square root of the product of material parameters, \( \varepsilon \) and \( \mu \).


3.4 DIFFERENT FEED TECHNIQUES FOR DR

Electromagnetic power can be coupled to the DR in several ways. These coupling mechanisms can have a significant impact on the resonant frequency and Q-factor. Numerous feeding techniques are available in the literature. Some of the commonly used techniques are,

3.4.1 SLOT/APERTURE COUPLING

Figure below depicts a DRA fed by an aperture. The aperture behaves like a magnetic current running parallel to the length of the slot, which excites the magnetic fields in the DRA. The aperture consists of a slot cut in a ground plane and fed by a microstrip line beneath the ground plane. This coupling mechanism has the advantage of having the feed network located below the ground plane, thus avoiding spurious radiation [17]. The microstrip stub can be designed to cancel out the reactive component of the slot, thus allowing for an impedance match to the DRA. Moreover, slot coupling is an attractive method for integrating DRAs with printed feed structures. The coupling level can be adjusted by moving the DRA with respect to the slot.

Figure 3.4.1.1: Slot fed DRA
3.4.2 COAXIAL PROBE COUPLING

The coaxial probe can either be located adjacent to the DRA or can be embedded within it. The amount of coupling can be optimized by adjusting the probe height and the DRA location. Also, depending on the location of the probe, various modes can be excited. For the probe located adjacent to the DRA, the magnetic fields of the $\text{TE}_{11\delta}$ mode of the rectangular DRA are excited (which radiate like a horizontal magnetic dipole). For a probe located in the centre of a cylindrical DRA, the $\text{TE}_{011}$ mode is excited (radiating like a vertical dipole) [17]. Another advantage of using probe coupling is that one can couple directly into a $50\Omega$ system, without the need for a matching network. Probes are useful at lower frequencies where aperture coupling may not be practical due to the large size of the slot required [18-21].

![Coaxial probe fed DRA](image)

**Figure 3.4.2.1:** Coaxial probe fed DRA

3.4.3 MICROSTRIP TRANSMISSION LINE / PROXIMITY COUPLING

Another common method for coupling to dielectric resonators in microwave circuits is by proximity coupling to microstrip lines. This approach is equally applicable
to DRAs as shown in figure. Microstrip coupling will excite the magnetic fields in the DRA to produce the short horizontal magnetic dipole mode. A metallic strip of definite width is etched on one side of a low loss dielectric substrate of known permittivity and thickness, the other side of which is metalized and grounded. An advantage of microstrip feed is that it is easier to fabricate, match and model. The feed is shown in Figure 3.4.3.1.

Coupling of EM energy and the input impedance are set by adjusting the lateral position of the DR with respect to the strip line [22, 23]. It is more convenient for the DRA arrays as well [24]. One disadvantage of this feed is that at higher frequency, surface wave modes are also excited in the substrate which affects the radiation pattern and efficiency of the DRA [25]. For lower permittivity values (necessary for DRAs requiring wide bandwidth), the amount of coupling is generally quite small.

**Figure 3.4.3.1:** Microstrip line fed DRA

### 3.4.4 COPLANAR SLOT FEEDS

Coupling to DRAs can also be achieved using co-planar feeds. Figure shows a ITDRA coupled to a co-planar loop. The coupling level can be adjusted by positioning
the DRA over the loop. The coupling behavior of the co-planar loop is similar to that of the coaxial probe, but the loop offers the advantage of being non intrusive. By moving the loop from the edge of the DRA to the centre, one can couple into either the HE_{11δ} mode or the TE_{011} mode of the cylindrical DRA [17, 26]. A coplanar slot can also be used to feed the DRA as shown in figure.

![Co-planar slot feed DRA](image)

**Figure 3.4.4.1:** Co-planar slot feed DRA

### 3.4.5 WAVE GUIDE FEED

The primary advantage of a waveguide is that it is extremely less lossy in the millimeter wave band. Since the wave is completely guided within the metallic structure, there is no threat of radiation loss when used as a feed line. As both the waveguide and DR are very low-loss, they form an excellent combination for low-loss millimeter-wave communication systems [27, 28]. Coupling to the DR can be achieved through a probe [29] or a slot [30]. A waveguide probe fed DRA is shown in Figure 3.4.5.1.
3.5 DIFFERENT DR GEOMETRIES

One of the attractive features of a DRA is that it can assume a number of shapes. Moreover the mode of operation and performance of a DRA can be varied by selecting a DR with desired structure [31]. Hence a number of DRA geometries have already been tried experimentally. The first systematic, theoretical, and experimental study was made on cylindrical disk DRA geometry. Later geometries such as split cylinder, sectored cylinder, cylindrical rings, metallized DRAs, triangular, rectangular, notched rectangular DRA, chamfered DRA, conical, elliptical, spherical, hemispherical, spherical cap, tetrahedral, perforated DRA, stepped DRAs, and hybrid DRAs, have been reported. It was found that DRAs operating at their fundamental modes radiate like an electric or magnetic dipole, which depends on the mode of excitation and geometry of the bulk dielectric material. Geometries like conical [32], stair [33], stacked triangular [34] etc emerged for dualband or wideband applications while those like cross [35], elliptical [36], hexagonal [37], cylindrical-comb [38] etc emerged for circular
polarization applications. Figure 3.5.1 shows the DR geometries, explained so far. Though several geometries have been introduced, the most studied and common structures are still the cylindrical and rectangular DRAs because of the simplicity in their design, fabrication, and analysis.

![Different DR geometries used](image)

**Figure 3.5.1:** Different DR geometries used

### 3.6 CHARACTERISTICS OF A DIELECTRIC RESONATOR

#### 3.6.1 DIELECTRIC CONSTANT

The dielectric constant of a material under given conditions reflects the extent to which it concentrates electrostatic lines of flux. In other words, it is the ratio of the amount of electrical energy stored in a material by an applied voltage, relative to that stored in a vacuum. An important property of a dielectric material is its ability to support
an electric field while dissipating minimal energy in the form of heat. The lower the
dielectric loss, the more effective a dielectric material is.

The net flux density $D$ can be expressed as

$$D = \varepsilon_o E + P \quad \text{.......(11)}$$

where $E$ is the electric field intensity and $P$ is the net polarization given by

$$P = \varepsilon_o \chi E \quad \text{......... (12)}$$

Where $\chi$ is the electric susceptibility. Now Eq. (11) becomes

$$D = \varepsilon_o (1 + \chi) E \quad \text{........... (13)}$$

$$= \varepsilon_o \varepsilon_r E$$

Now we define the relative permittivity as,

$$\varepsilon_r = 1 + \chi \quad \text{........... (14)}$$

in the complex form,

$$\varepsilon_r = \varepsilon_r' - j\varepsilon_r'' \quad \text{............... (15)}$$

In Eq. (15), the real part is called the dielectric constant and the ratio $\frac{\varepsilon_r''}{\varepsilon_r'} = \tan \delta$ is called
the dissipation or loss tangent of the dielectric.

Hence it is clear that the dielectric properties of a DR are resulted from the
phenomenon called dielectric polarization that occurs when electromagnetic fields pass
through them. A DR at rest contains randomly oriented permanent electric dipoles. When
an external field is applied, the dipoles align themselves in the direction of the field and
the material is said to be polarized. For most materials $P$ vanishes as $E$ vanishes.
### 3.6.2 QUALITY FACTOR

The radiation Q-factor of the DRA is determined using [3]:

\[
Q = \frac{2\omega W_e}{P_{rad}}
\]

(16)

where \(W_e\) and \(P_{rad}\) are the stored energy and radiated power, respectively. These quantities are given by:

\[
W_e = \frac{\varepsilon_r \varepsilon_0 \omega d}{32} \left(1 + \frac{\sin(k_z d)}{k_z d}\right) \left(k_x^2 + k_y^2\right)
\]

(17)

\[
P_{rad} = 10k_0^4 |p_m|^2
\]

(18)

where \(p_m\) is the magnetic dipole moment of the DRA:

\[
p_m = -j\omega \varepsilon_0 \varepsilon_r (\varepsilon_r - 1) k_z \sin(k_z d / 2) \hat{z}
\]

(19)

The impedance bandwidth (BW) of the DRA can be estimated from the radiation Q-factor using:

\[
BW = \frac{S - 1}{Q \sqrt{S}}
\]

(20)

where \(S\) is the maximum acceptable voltage standing-wave ratio (VSWR). The above equations can be used to generate the graphs which plot the normalized Q-factor \(Q_e\) as a function of the DRA dimensions \(d/b\) for various values of dielectric constant and various values of \(a/b\). The normalized Q-factor is defined as:
Thus these curves can be used to estimate the Q-factor of a DRA without having to rely on the preceding equations.

3.7 FABRICATION OF THE DIELECTRIC RESONATOR

The DR is fabricated through the mixed oxide or solid state route which is a time consuming process, involving many steps \([39]\) such as Mixing and Grinding, Calcination process, Pellet shaping, Sintering process and Surface finishing. During these processes we can minimize the free energy of the material and redistribute the atoms. The minimization involves the reduction of internal surface area and an increase in the grain size.

3.7.1 MIXING AND GRINDING

Here, the preparation of DR sample from ZnTiO\(_3\) material is explained as an example as it is used for the work in this thesis. Titanates (TiO\(_2\)) have many uses in electronic and material industry due to its piezoelectric, ferroelectrics and other properties.

We start with the chemical equation of the compound ZnTiO\(_3\) which is

\[
\text{ZnO} + \text{TiO}_2 \rightarrow \text{ZnTiO}_3
\]

or in terms of atomic weight,

\[
(65.39 \,(\text{Zn})+15.999(\text{O})) \text{ gms of ZnO} + (47.67(\text{Ti})+2*15.999(\text{O})) \text{ gms of TiO}_2 \rightarrow \\
(65.39(\text{Zn})+47.67(\text{Ti})+3*15.999) \text{ gms of ZnTiO}_3
\]

or

\[ Q_e = \frac{Q}{\varepsilon_r^{3/2}} \quad (21) \]
81.389 gms of ZnO + 79.668 gms of TiO₂ → 161.057 gms of ZnTiO₃

This gives the fact that 1 gm of ZnTiO₃ requires 0.5053 gm of ZnO and 0.4947 gm of TiO₂. Thus the stochiometric quantities of ZnO and TiO₂ required for forming N gms of ZnTiO₃ as the final product can be calculated easily. The next step is mixing for eliminating aggregates and/or reducing the particle size. The weighed powders of ZnO and TiO₂ are mixed well with 100−200 % of distilled water for about 12 hrs in a ball-mill, which is a motor-driven barrel that rotates on its axis. The barrel is filled with the ceramic beads made of alumina or silicon carbide that act as the grinding medium for the powder. The creamy mixer is then dried in an oven at a 100°C.

3.7.2 CALCINATION PROCESS

In this process the endothermic decomposition reaction is taken place. Any salt such as carbonate or hydroxide, decomposes, leaving an oxide as a solid product by liberating a gas. This process causes the interaction of the constituents by the interdiffusion of their ions and so reduces the extent of the diffusion that must occur during sintering in order to obtain a homogeneous body. The calcinations conditions are important factors determining the shrinkage of the pellet during the sintering. The thermal conductivity of powdered materials is always low, so that a sufficiently uniform temperature can only be obtained through a depth of a few centimeters when the period at maximum temperature is 1 or 2 hours in most cases. If compound formation is to occur during calcinating or firing, the matter of neighboring particles must inter-diffuse and the time taken to complete the process is proportional to the square of the particle size. Here
the well mixed powder is taken in an alumina crucible and calcinated at a temperature of
1000°C in an electric muffle furnace, for 2 hrs.

3.7.3 PELLET SHAPING

After the calcination process, the powder is crushed well in an agate mortar to
form finer powder and then, mixed well with 4 % of Poly Vinyl Alcohol (PVA), an
organic binder. Mixing with the binder provides sufficient strength to resist the
disintegrating effect of small stress on the shaped pellets prior to sintering. Dry pressing
is carried out in dye with movable top and bottom punches, made of hardened steel as
shown in figure 3.7.3.1.

![Figure 3.7.3.1: Photograph of Dyes used for ITDR and Cylindrical DR](image)

The die used in this work is hexagonal in shape that can be used for fabricating
hexagonal DR as well. After fixing the dye cavity on the bottom punch, one of the top
punches is placed inside the cavity. To fabricate ITDR, the remaining void space in the cavity is filled with an adequate amount of free-flowing powder (already prepared) and the top punch is descended to compress the powder to a predetermined volume, to a set pressure (75–300 MPa). Highly polished dye and punch surfaces ensure reduced wall friction. Shapes with a uniform section in the pressing direction are the easiest to produce by dry pressing. The time taken on an automatic pressing machine varies from 0.2 second for pieces of diameter around 1 mm to 5 seconds for large complex shapes.

Figure 3.7.3.2: Schematic Sketches of the Dye and ITDR pellet
3.7.4 SINTERING PROCESS

Sintering converts the compacted powder into a denser structure of crystallites jointed to one another by grain boundaries, at elevated temperatures below the melting point of the material. The energetic basis for sintering lies in the reduction of surface energy by transferring matter from the interior of grains along the grain boundaries to adjacent pores, which are eventually filled. Usually the powder compact is heated at fixed sintering temperature, held at this temperature for the required time and finally cooled at the room temperature. This is referred to as isothermal sintering. The organic binder is burnt out at the lower sintering temperatures. In the present case, isothermal sintering of the pellets, placed on an alumina slab at 1150°C for 5 hours is carried out after which it is cooled to the room temperature.

3.7.5 SURFACE FINISHING

Tool wear during the pellet shaping and variations in shrinkage during sintering and drying contribute to 1−2 % variation in the dimensions of the sintered pellets. For experimental studies, especially in the case of material characterisation, the surfaces of the pellets need to be as smooth as possible. Usually it is done by grinding and lapping the dense sample with tools consisting of silicon carbide, diamond powder etc. Here we use a silicon carbide water roof paper for finishing the pellets. A photograph of the final DR samples is shown in Figure 3.7.5.1.
3.8 MICROWAVE SUBSTRATES

Selection of substrates in microwave circuits is very important. Low loss substrates are very important at microwave bands. As frequency of operation increases, the loss tangent of the material used for substrates slightly increases, which in turn adversely affect the efficiency of the antenna. The power handling capability of the antenna depends on the substrate materials also. At high power certain substrate materials cannot withstand. A variety of substrate materials are available in the market. Flexible substrate materials are also available, so that the antenna can be mounted on curved surfaces. The selection of dielectric constant of the substrate depends on the application of the antenna and the radiation characteristics specifications. It is worth noting that surface waves will be excited in high dielectric constant substrates. This will generate spurious radiations in unwanted directions from the antenna. In this thesis importance is
given to compactness of the antenna structure. Prototype of antennas was fabricated on FR4 substrate which has a dielectric constant = 4.4, tan δ = 0.02 and thickness = 1.6 mm.

### 3.9 CHARACTERIZATION METHODS OF DR

Dielectric constant and quality factor of the fabricated DRs are measured using the well-known cavity methods as described below.

#### 3.9.1 Measuring Dielectric Constant using Hakki - Coleman Method

Depending on the operating principle, methods for measuring the complex permittivity of materials at microwave frequencies can be classified as (1) methods that depend on the standing wave field within the dielectric (2) methods that depend on transmitted waves or waves reflected from the dielectric and (3) resonance methods. When the material is available only in small volume, cavity perturbation techniques are suitable, but the measurement accuracy is limited to dielectric constants less than 10. Hakki and Coleman method [42] is most suited when the ceramic samples have higher dielectric constant, and the method uses a dielectric post resonator for this purpose.

The measurement setup consists of a cylindrical DR puck sandwiched between two conducting plates (of infinite extent theoretically) to form a parallel-plate DR. This method restricts most of the stored energy to the dielectric and allows the experimental configuration to closely approximate the analytical model. If the distance between the two parallel plates is smaller than one-half wavelength, then the excited TE\(_{011}\) mode will not radiate [43] and the sides of the resonator can be left open for providing the coaxial
coupling probes. The maximum dimensions of the specimen are set by the diameter of the shorting plates while the minimum dimensions by the diameter of the coupling probes. The setup is shown in Figure 3.9.1.1.

![Hakki-Coleman setup for measuring dielectric constant](image)

**Figure 3.9.1.1:** Hakki-Coleman setup for measuring dielectric constant

Consider a cylindrical DR of length $L$ and radius $a$ placed in the above setup. Then the characteristic equation for the TE$_{0nl}$ mode of operation is given by

$$\frac{\alpha J_0(\alpha)}{J_1(\alpha)} = -\beta \frac{K_0(\beta)}{K_1(\beta)}$$  \hspace{1cm} (22)

where $J_0(\alpha)$ and $J_1(\alpha)$ are the Bessel functions of the first kind of orders zero and one respectively, while $K_0(\beta)$ and $K_1(\beta)$ are the modified Bessel functions of the second kind of orders zero and one respectively. Also

$$\alpha_n = \frac{2\pi a}{\lambda_0} \sqrt{\varepsilon_r - \left(\frac{l\lambda_0}{2L}\right)^2}$$  \hspace{1cm} (23)

$$\beta_l = \frac{2\pi a}{\lambda_0} \sqrt{\left(\frac{l\lambda_0}{2L}\right)^2 - 1},\hspace{1cm} (24)$$
where \( l \) is the axial wave number. Thus the dielectric constant can be obtained from (23) and (24) as

\[
\varepsilon_r = 1.0 + \left( \frac{c}{\pi D f_1} \right)^2 \left( \alpha_1^2 + \beta_1^2 \right) \tag{25}
\]

where \( c = 3 \times 10^8 \text{ m/s} \), \( \alpha_1 \) and \( \beta_1 \) are the first roots of the characteristics equation with \( n = l = 1 \) corresponding to the TE_{011} mode.

### 3.9.2 Measurement of Quality factor using Cavity Method

Q-measurement methods are mainly of two types – time domain and frequency domain. Time domain methods mainly depend on measuring the decay time constant \( \tau \) of the stored energy in the cavity at frequency \( f_0 \), and by using the following relation [44].

\[
Q_L = \frac{2\pi f_0 \tau}{2} \tag{26}
\]

Three useful frequency domain techniques are the reflection method, the reactance method and the transmission method. Transmission method is the simplest and requires a transmission type cavity as shown in Figure 3.9.2.1.

As shown in the figure, a microstrip transmission line is fabricated on a dielectric substrate. The DR is coupled magnetically to the transmission line by placing it nearby it on the substrate. The lateral distance \( d \) between the strip and the centre of the DR determines the coupling coefficient between them. By properly adjusting \( d \), the TE_{01\delta} mode can be excited in the DR. In order to suppress the radiation losses, the entire structure is covered with a metallic cavity of dimensions at least 3 times the size of the...
DR, with a top plate that can be moved up and down using a screw. The shielding conditions affect the resonant frequency and the Q of the DR.

Figure 3.9.2.1: Top view of the transmission type cavity setup for Q-factor measurement

The degree of coupling is adjusted such that the transmission loss is of the order of $-40$ dB. By bringing the top metal plate close to the DR, the TE$_{01\delta}$ resonant frequency can be observed increasing, indicating that the stored energy in the cavity is predominantly magnetic. If the stored energy is electric, then a decrease in the resonant peak is expected.

From the transmission coefficient ($|S_{21}|$) plot around the resonant frequency, the loaded and unloaded Q-factors $Q_L$ and $Q_u$ respectively can be calculated as illustrated in Figure 3.9.2.2.

Here the parameter $x$ is given by [10]

$$x=3-10\log\left(1+10^{-0.1|S_{21}|_{\text{dB}}}\right) \quad (27)$$
Now the Q-factor is given by the well-known equation, \( Q = \frac{f_0}{\Delta f} \)

**Figure 3.9.2.2:** Measurement of Q-factor from the \( S_{21} \) curve

For the measurement of the temperature coefficient of resonant frequency (\( \tau_f \)), the Hakki-Coleman transmission cavity is placed in a temperature stable furnace with outlets for signal coupling. The temperature is varied over a desired range in discrete steps and the shift in the \( TE_{011} \) frequency is noted. Now \( \tau_f \) can be calculated as given by Eq. (28). Here \( f_0 \) is the \( TE_{011} \) resonant frequency at room temperature and \( \Delta f_0 \) is the frequency shift for a temperature gradient of \( \Delta T \). The \( \tau_f \) can be either positive or negative depending on whether the frequency is increasing or decreasing respectively with the rise in temperature.

\[
\tau_f = \frac{1}{f_0} \cdot \frac{\Delta f_0}{\Delta T} \text{ parts per million or ppm } /^\circ C \quad (28)
\]
3.10 EXPERIMENTAL CHARACTERIZATION SETUP

Antenna characteristics such as return loss, radiation pattern and gain are measured using the HP8510C and associated setup. The indegenously developed CREMA SOFT is used for the automatic measurement of the radiation properties using HP 8510C Network analyzer. The important systems used for the antenna characterization are Vector network Analyzer, Anechoic Chamber, Automated turn table etc.

3.10.1 HP 8510C VECTOR NETWORK ANALYZER

This is a sophisticated Vector Network Analyzer (VNA) from Hewlett Packard with time domain and frequency domain operation capability [45]. The NWA can measure the magnitude and phase of the S parameters. The microprocessor based system can measure two port network parameters such as $s_{11}$, $s_{12}$, $s_{21}$ and $s_{22}$ very accurately. The in built signal processing algorithms of the network analyzer process the transmit and receive data and finally displays the measured values in many plot formats. The schematic of the VNA is shown in Fig. 3.10.1.1.

The network analyzer consists of a microwave generator, S parameter test set, signal processor and the display unit as illustrated in Fig. 3.2. The synthesized sweep generator HP 83651B uses an open loop YIG tuned element to generate the RF stimulus. It can synthesize frequencies from 10 MHz to 50 GHz. The frequencies can be set in step mode or ramp mode depending on the required measurement accuracy.
Figure 3.10.1.1: Schematic diagram of HP 8510C vector network analyzer set up used for the characterization of the antennas

The antenna under test (AUT) is connected to the port of the S-parameter test set HP8514B and the forward and reflected power at the measurement point is separated and down converted to 20MHz using frequency down converter. It is again down converted to lower frequency and processed in the HP8510C processing unit. All the systems discussed above are interconnected using HPIB bus. A computer interfaced to the system is used for coordinating the whole operation remotely. Measurement data can be saved on a storage medium using it.
3.10.2 ANECHOIC CHAMBER

The anechoic chamber provides a quiet zone needed to simulate space environment required in pattern measurements. The absorbers used for building the chamber are made from high quality, low-density form impregnated with dielectrically/magnetically lossy medium. The wall of the chamber (24’ X 12’ X 10’ ) used for the measurements is properly shaped (tapered chamber) and covered with carbon black impregnated poly urethene (PU) foam based pyramidal, wedge, or flat absorbers of appropriate sizes. The PU foam structure gives the geometrical impedance matching while the dispersed carbon gives the required attenuation (up to –40 dB) for a wide frequency (500 MHz to 18 GHz) range. The chamber is made free of EMI by surrounding with thin aluminium sheet.

3.10.3 TURN TABLE ASSEMBLY FOR FAR FIELD RADIATION PATTERN MEASUREMENT

The turn table assembly consists of a stepper motor driven rotating platform for mounting the Antenna Under Test (AUT). The in-house developed microcontroller based antenna positioner STIC 310C is used for radiation pattern measurement. The main lobe tracking for gain measurement and radiation pattern measurement is done using this setup. A standard wideband horn (1-18GHz) is used as receiving antenna for radiation pattern measurements. The in-housed developed automation software ‘Crema Soft’ coordinates all the measurements.

3.10.4 MEASUREMENT PROCEDURE

The experimental procedures followed to determine the antenna characteristics are discussed below. The network analyzer in real practice is connected to
large cables and connectors. The connectors and cables will have its losses associated at higher microwave bands. Thus the instrument should be calibrated with known standards of open, short and matched loads to get accurate scattering parameters. There are many calibration procedures available in the network analyzer. Single port, full two port and TRL calibration methods are usually used. The two port passive or active device scattering parameters can be accurately measured using TRL calibration method. Return loss, VSWR and input impedance can be characterized using single port calibration method.

### 3.10.4.1 RETURN LOSS, RESONANT FREQUENCY AND BANDWIDTH

The return loss characteristic of the antenna is obtained by connecting the antenna to any one of the network analyzer port and operating the VNA in $s_{11}/s_{22}$ mode. The calibration of the port is done for the frequency range of interest using the standard open, short and matched load. The calibrated instrument including the port cable is now connected to the device under test. The frequency vs reflection parameter ($s_{11}/s_{22}$) values is then stored on a computer using the ‘Crema Soft’ automation software.

The frequency for which the return loss value is minimum is taken as resonant frequency of the antenna. The range of frequencies for which the return loss value is within the -10dB points is usually treated as the bandwidth of the antenna. The antenna bandwidth is usually expressed as percentage of bandwidth, which is defined as

$$
%\text{Bandwidth} = \frac{\text{bandwidth}}{\text{centrefrequency}} \times 100
$$
At -10dB points the VSWR is ~2. This implies that at resonance the VSWR value approaches unity. The above bandwidth is sometimes referred to as 2:1 VSWR bandwidth.

### 3.10.4.2 FAR FIELD RADIATION PATTERN

The measurement of far field radiation pattern is conducted in an anechoic chamber. The AUT is placed in the quite zone of the chamber on a turn table and connected to one port of the network analyzer. A wideband horn is used as a transmitter and connected to the other port of the network analyzer. The turn table is controlled by a STIC positioner controller. The automated radiation pattern measurement process is coordinated by the ‘Crema Soft’ software in the remote computer.

In order to measure the radiation pattern, the network analyzer is kept in $S_{21}/S_{12}$ mode with the frequency range within the -10dB return loss bandwidth. The number of frequency points are set according to the convenience. The start angle, stop angle and step angle of the motor is also configured in the ‘Crema Soft’. The antenna positioner is boresighted manually. Now the THRU calibration is performed for the frequency band specified and saved in the CAL set. Suitable gate parameters are provided in the time domain to avoid spurious radiations if any. The Crema Soft will automatically perform the radiation pattern measurement and store it as a text file.

### 3.10.4.3 ANTENNA GAIN

The gain of the antenna under test is measured in the bore sight direction. The gain transfer method using a standard gain antenna is employed to determine the absolute
gain of the AUT [46-47]. The experimental setup is similar to the radiation pattern measurement setup. An antenna with known gain is first placed in the antenna positioner and the THRU calibration is done for the frequency range of interest. Standard antenna is then replaced by the AUT and the change in S21 is noted. Note that the AUT should be aligned so that the gain in the main beam direction is measured. This is the relative gain of the antenna with respect to the reference antenna. The absolute gain of the antenna is obtained by adding this relative gain to the original gain of the standard antenna.

3.10.4.4 RADIATION EFFICIENCY OF DRA

The radiation efficiency $\eta_{rad}$ describes the losses within the antenna structure. It is defined by the ratio of the radiated power $P_{rad}$ over the power $P_{in}$ going into the antenna terminal.

$$\eta = \frac{P_{rad}}{P_{in}} = \frac{P_{rad}}{P_{rad} + P_{loss}} = \frac{R_{rad}}{R_{rad} + R_{loss}} \quad (3.27)$$

Where $P_{rad}$ = power radiated

$P_{in}$ = power fed to antenna (W)

$P_{loss}$ = power lost by the antenna (W)

$R_{rad}$ = radiation resistance of the antenna (Ω)

$R_{loss}$ = loss resistance of the antenna (Ω)

For physically small antennas, the Wheeler cap method [48] is highly preferred for measuring the radiation efficiency. According to this method, if a radiation shield is placed around the antenna so as to enclose the near fields of the antenna, the radiation resistance of the antenna is reduced to zero while the loss resistance and the stored energy
remain the same as for the unshielded antenna [49]. When covering the antenna with a metal cap, the radiation is suppressed and the input power (proportional to the input resistance) is equal to the power loss (proportional to the loss resistance). Without the cap, the input power is equal to the radiated power plus the power loss (input resistance + loss resistance). The radiation efficiency of the antenna can be obtained from these two parameters.

### 3.11 ANTENNA UNDER TEST

The antenna under test (AUT) is the DRA designed using the fabricated ITDRs. Mainly three feeding methods are used. One with microstrip line feed alone and the other two are microstrip line feed with slotted ground plane. The slots provided in the ground plane are slightly different in shape for the use in Design 5-1 and Design 5-2 as would be seen in section 4.5.1 and 4.5.2 in chapter 4. The feed with substrate is shown in the figure.

![Antenna Under Test](a)(b)
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Figure 3.11.1: Photographs of the (a) feed used in Design 5-1 (b) feed used in Design 5-2 and (b) the antenna configuration with microstrip feed alone.

REFERENCES:


