This chapter highlights the general characteristics, fabrication methodology and characterisation of dielectric resonators to be useful as dielectric resonator antennas. First part of the chapter is dedicated to various aspects of the DR and the DR antenna design. Basic measurement facilities used and the measurement details of the important antenna characteristics — return loss, impedance, radiation pattern, gain and radiation efficiency are explained in the second part.

3.1 INTRODUCTION

Dielectric resonators (DRs) emerged as a substitute to resonant metallic cavities and waveguides in microwave devices like filters, oscillators, and phase shifters. As for metallic cavities, the resonant frequency of a DR is determined by its dimensions and also exhibits high Q-factors. But the main difference between the two is that the wavelength in dielectric materials (non-magnetic) is reduced by a factor of one over square root of the dielectric constant, \( \varepsilon_r \), which is much higher than unity for most materials. Hence the resonator can be made smaller by choosing a high dielectric constant material. However, the reactive power stored in a DR during resonance is not strictly confined inside the resonator. The leakage fields from the resonator can be used for energy coupling, frequency tuning or radiation purpose. To be useful in practical applications, a DR basically requires a high dielectric constant (\( \varepsilon_r > 20 \)) for promising size reduction, high Q-factor (\( Q_d > \))
5000), hence a low dielectric loss factor \(\tan \delta \sim (Q_d)^{-1}\) for a stable resonance and a near-zero temperature coefficient of resonant frequency \(\tau_t \sim 0 \text{ ppm/°C}\) for temperature stability and hence better circuit performance. The implication is nothing but a drastic reduction in the total cost of the RF system.

### 3.2 CHARACTERISTICS OF A DIELECTRIC RESONATOR

#### 3.2.1 Dielectric Constant

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 proportion of energy lost as heat), the more effective a dielectric material is. Another consideration is the dielectric constant, the extent to which a substance concentrates the lines of electric flux within.

The net flux density \(D\) can be expressed as

\[
D = \varepsilon_0 E + P \quad (3.1)
\]

where \(E\) is the electric field intensity and \(P\) is the net polarization given by

\[
P = \varepsilon_0 \chi E \quad (3.2)
\]

where \(\chi\) is the electric susceptibility. Now Eq. (3.1) becomes

\[
D = \varepsilon_0 (1 + \chi) E \quad (3.3)
\]

\[
= \varepsilon_r \varepsilon_0 E
\]

Now we define the relative permittivity as,

\[
\varepsilon_r = 1 + \chi \quad (3.4)
\]
in the complex form,

\[ \varepsilon_r = \varepsilon_r' - j\varepsilon_r'' \]  (3.5)

In Eq. (3.5), 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 polarisation that occurs when electromagnetic fields pass through them. A DR at rest contains randomly oriented permanent electric dipoles. When an external electric field is applied, the dipoles align themselves in the direction of the field and the material is said to be polarised. For most materials \( P \) vanishes as \( E \) vanishes.

### 3.2.2 Resonant Mode and Resonant Frequency

The modes of a cylindrical DR, placed on a large ground plane as shown in Figure 1, can be classified into three distinct types: Transverse Electric (TE), Transverse Magnetic (TM) and hybrid (HE and EH) [1]. TE with respect to the \( Z \)-axis means that electric field component in \( Z \)-direction is zero. In other words, two components of electric field perpendicular to \( Z \)-axis exist in the structure. TM mode also can be explained in a similar way. These two modes are axisymmetric, meaning that they have no azimuthal (\( \Phi \)) variation. But the hybrid modes are \( \Phi \) dependant, which can further be subdivided into hybrid electric (HE) and hybrid magnetic (EH) modes. In the HE mode, the \( E_z \) dominates the \( H_z \) and all other field components are expressed in terms of \( E_z \) only. Reverse is applied for the EH
mode. In general, a mode can be expressed as the mode name subscripted by three indices $n$, $p$ and $m$ which respectively represent the field variation along azimuthal, radial and the axial directions.

![Diagram of a cylindrical DR over a ground plane at $z = 0$](image)

Figure 1: A cylindrical DR over a ground plane at $z = 0$

In electromagnetics, metallic waveguides and resonators are modeled using perfect electric conductor (PEC) wall or electric wall, for which the tangential component of electric field vanishes or $n \times E = 0$, $n$ being the normal to the conductor surface. Similarly, a cylindrical dielectric resonator is modeled using the cavity model where the magnetic wall boundary is imposed on the outer surfaces of the DR, to predict its in-field distribution and the resonant frequency. According to the magnetic wall condition, $n \times H = 0$ at all the outer surfaces of the resonator with normal $n$, on imposing which we get the axial field components as

$$
\begin{align*}
\text{TE: } H^\text{TE}_{zpm} &= J_n \left( \frac{X_{zp}}{a} \right) \left\{ \frac{\sin(n \phi)}{\cos(n \phi)} \right\} \sin \left[ \frac{(2m+1)\pi}{2d} z \right] \\
\text{TM: } E^\text{TM}_{zpm} &= J_n \left( \frac{X_{zp}^2}{a} \right) \left\{ \frac{\sin(n \phi)}{\cos(n \phi)} \right\} \cos \left[ \frac{(2m+1)\pi}{2d} z \right]
\end{align*}
$$

(3.6)

$n = 1, 2, 3 \ldots \quad p = 1, 2, 3 \ldots \quad m = 0, 1, 2 \ldots$
Other field components can be evaluated by substituting Eq. (3.6) in the Maxwell’s equations.

The resonant frequency for the $\text{TE}_{npm} / \text{TM}_{npm}$ mode is determined by solving the separation equation,

$$k^2_t + k^2_z = \varepsilon_t \left( \frac{2\pi f}{c} \right)^2$$

and is given by,

$$f_{npm} = \frac{c}{2\pi a\sqrt{\varepsilon_t}} \sqrt{\left( \frac{X_{np}^2}{X_{np}^2} \right) + \left[ \frac{\pi a}{2d(2m+1)} \right]^2}$$

where $X_{np}$ is the root of the characteristic equation $J_n(X_{np}) = 0$, $X'_{np}$ is the root of $J'_n(X'_{np}) = 0$, $J_n$ is the $n^{th}$ order Bessel function of the first kind and $J'_n$ is the first derivative of $J_n$. Also $k_t$ and $k_z$ are the radial and axial wave numbers respectively. Top view of the field distributions for some modes of a cylindrical DR are shown in Figure 2 [2].

### 3.2.3 Quality Factor

Quality or Q-factor is a measure of the ability of the DR to store microwave energy with minimal signal loss. The inherent Q-factor of a DR solely depends on the loss factor of the dielectric material. But in practical applications, the resonator is always associated with metallic parts, in the form of shields or ground planes. In general, the loaded Q-factor of a resonant cavity can be defined as the ratio of the stored energy to the dissipated power [3].
\[
Q_L = \frac{\text{Stored Energy}}{\text{Dissipated Power}} = \frac{2\omega_a E_s}{P_{\text{dis}}}
\]

(3.9)

Figure 2: Field distributions inside a cylindrical DR

The denominator term, which is the total power dissipation, can occur in many ways such as conductor losses \((P_c)\), dielectric losses \((P_d)\), radiation losses \((P_r)\) and/or losses in the external circuits \((P_{ext})\).

i.e.; \[P_{\text{dis}} = P_c + P_d + P_r + P_{ext}\] (3.10)

Combining equations (3.9) and (3.10), we get

\[
\frac{1}{Q_L} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_r} + \frac{1}{Q_{ext}}
\]

(3.11)

where the unloaded Q is given by
Here, $1/Q_d = \tan \delta$ is the loss tangent of the dielectric. However, the material loss of the dielectric itself can be divided into intrinsic and extrinsic losses. Intrinsic losses depend on the material composition which causes the microwave energy to be converted into thermal energy. But the extrinsic losses result from the microstructural irregularities like porosity, grain boundaries, impurities etc. Theoretically, the Q-factor linearly decreases with frequency; therefore usually the product $Q_f$ in GHz is specified for DRs.

Another aspect of the Q-factor is that, if a DR of negligible losses is placed in the unshielded/open environment, then $Q_u = Q_r$, the radiation Q-factor for the lower order modes. For a given excited mode, the $Q_r$ depends on the aspect ratio and the dielectric constant for a cylindrical DR. Then the bandwidth of the DR over which its voltage standing wave ratio (VSWR) conforms to a specified value $S$ can be expressed as [1]

$$BW = \frac{S-1}{Q_u \sqrt{S}} \times 100 \%$$  \hspace{1cm} (3.13)
3.3 FABRICATION OF THE DIELECTRIC RESONATOR

The DR is fabricated through the mixed oxide or solid state route that involves the following steps [4]:

(i) Weighing, Mixing and Grinding
(ii) Calcination
(iii) Shaping
(iv) Sintering
(v) Finishing

3.3.1 Weighing, Mixing and Grinding

Here, the preparation of DR sample from the zinc titanate (ZnTiO₃) material is explained as an example.

We start with the chemical equation of the compound ZnTiO₃ which is

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

or in terms of atomic weight,

\[(65.39 + 15.999) \text{ gms of ZnO} + (47.67 + 2 \times 15.999) \text{ gms of TiO}_2 \rightarrow \]

\[(65.39 + 47.67 + 3 \times 15.999) \text{ gms of ZnTiO}_3 \]

or

81.389 gms of ZnO + 79.668 gms of TiO₂ → 161.057 gms of ZnTiO₃ or

Thus 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 ZnTiO$_3$ 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 TiO$_2$ 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 100°C.

### 3.3.2 Calcination

Calcination is the final step in the production of high purity ceramic powder. This process is the endothermic decomposition reaction in which, any salt such as carbonate or hydroxide decomposes, leaving an oxide as a solid product 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 the important factors determining the shrinkage of the pellet during the sintering. The surfaces of the container in immediate contact with the powder must not react with it in order to avoid contamination. 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. The process will clearly
be considerably slower if the particles consist of aggregates of crystals. In the present case, the well mixed powder is taken in an alumina crucible and calcined at a temperature of 1000°C in an electric muffle furnace, for 2 hrs.

3.3.3 Pellet Shaping

The calcined powder is crushed well in an agate mortar (bowl made of agate, a hard material) to form finer powder and mixed well with 4% of Polyvinyl 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 a die with movable top and bottom punches, made of hardened steel. The free-flowing powder is filled in the die-cavity, which is cylindrical in shape and the top punch is descended to compress the powder to a predetermined volume, to a set pressure (75–300 MPa). The green density of the pellet is not greatly increased by applying pressures exceeding 74–150 MPa. Highly polished die 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.

3.3.4 Sintering

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.

3.3.5 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.
3.4 MICROWAVE CHARACTERISATION OF THE DIELECTRIC RESONATOR

Dielectric constant ($\varepsilon_r$), quality factor (Q) and temperature coefficient of resonant frequency ($\tau_f$) of the fabricated DRs are measured using the microwave techniques as described below.

3.4.1 Hakki-Coleman Method for Measuring Dielectric Constant

There are numerous conventional methods for measuring the complex permittivity of materials at microwave frequencies. Depending on the operating principle, these methods 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 (3) resonance methods [5]. Cavity perturbation techniques [5, 6] are suitable for the measurement of materials available in a small volume but the measurement accuracy is limited to dielectric
constants less than 10. This is because, the sample volume required will be too small to be handled, when measuring high dielectric constant materials. For ceramic samples of higher dielectric constant, Hakki and Coleman method [7] employing a dielectric post resonator is used.

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 [8] 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 measurement setup is shown in Figure 4.

![Cylindrical DR sample](image)

**Figure 4:** Hakki-Coleman setup for dielectric constant measurement
Consider a cylindrical DR of length $L$ and radius $a$ placed in the above setup. Then the characteristic equation for the TE_{0pm} mode of operation is given by

$$\frac{J_0(\alpha)}{J_1(\alpha)} = -\beta \frac{K_0(\beta)}{K_1(\beta)} \quad (3.14)$$

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

$$\alpha_p = \frac{2\pi a}{\lambda_o} \sqrt{\varepsilon_r - \left(\frac{m\lambda_o}{2L}\right)^2} \quad (3.15)$$

$$\beta_m = \frac{2\pi a}{\lambda_o} \sqrt{\left(\frac{m\lambda_o}{2L}\right)^2 - 1} \quad (3.16)$$

where $m$ is the axial wave number. Thus the dielectric constant can be obtained from (3.15) and (3.16) as

$$\varepsilon_r = 1.0 + \left(\frac{c}{2\pi a f_0}\right)^2 \left(\alpha_1^2 + \beta_1^2\right) \quad (3.17)$$

where $c = 3\times10^8$ m/s, $\alpha_1$ and $\beta_1$ are the first roots of the characteristics equation with $p = m = 1$ corresponding to the TE_{011} mode.
3.4.2 Khanna-Garault Method for Measuring the Quality factor

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 [9].

$$Q_L = 2\pi f_0 \tau$$ (3.18)

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 5.

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$_{018}$ 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 tuning screw. The shielding conditions affect the resonant frequency and the $Q$ of the DR.
Figure 5: Top view of the Khanna-Garault cavity setup for Q-factor measurement

The degree of coupling is adjusted such that the transmission loss is of the order $\leq -30$ dB. By bringing the top metal plate close to the DR, the TE$_{018}$ 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 6 [10].

In the figure, the parameter $x$ is given by

$$x=3-10\log\left(1+10^{-0.1|S_{21}|dB}\right)$$

(3.19)

Now the Q-factor is given by the well-known equation, $Q = \frac{f_0}{\Delta f}$
Figure 6: Measurement of Q-factor from the S$_{21}$ curve

For the measurement of the temperature coefficient of resonant frequency ($\tau_f$) for a particular mode of the DR, any of the above two cavity arrangements (Fig. 4 and 5) can be used. In the present study, the Khanna-Garault transmission cavity, exciting the TE$_{015}$ mode of the DR is employed. The arrangement is mounted in a temperature stable furnace with outlets for signal coupling. The temperature is varied from 29°C to 70°C in steps, and the corresponding resonant frequencies are noted. Now $\tau_f$ can be calculated as given by Eq. (3.20).

$$\tau_f = \frac{1}{f_0} \cdot \frac{\Delta f_0}{\Delta T} \times 10^{-6} \text{ / °C or parts per million or ppm/°C} \quad (3.20)$$

Here $f_0'$ is the TE$_{018}$ frequency at room temperature (29°C) and $\Delta f_0'$ is the frequency shift for a temperature gradient of $\Delta T$. The value of $\tau_f$ can either be
positive or negative depending on whether the frequency is increasing or
decreasing respectively with the rise in temperature.

3.5 EVOLUTION OF DIELECTRIC RESONATOR AS A RADIATOR

Dielectric resonators were initially used in filters, oscillators and phase
shifters, by virtue of their high Q-factor, where the energy lost is negligible
compared to the energy stored. When used as an antenna, the DR is designed for
a low Q-factor, so that the energy lost through radiation is much higher than the
energy stored. This is achieved by the proper selection of dielectric constant,
geometry and feeding method for the DR placed in an open or unshielded
environment. The first antenna using a cylindrical DR was successfully devised by
S.A. Long et al. in 1983 [11], exciting the low-Q TM_{110} mode of the DR using a
coaxial probe. The antenna was then named as a dielectric cavity antenna which
was later transformed into the most popular name dielectric resonator antenna (DRA).
Since the DR is fully dielectric, no threat of ohmic loss and this results in higher
radiation efficiency when used as antenna. Since they are 3-D devices, when the
frequency of operation is halved, the antenna becomes eight times bulkier. This
puts a limit to the application of bare-DRs at frequencies below L-band, where
DR loaded patch antennas are being used as a compromise.

Merits of DRAs [26]

> A wide variety of geometries can be used allowing design flexibility

> High radiation efficiency
• Compatibility with numerous existing feeding mechanisms
• Good control over size and impedance bandwidth is achievable by using wide range of dielectric constants (10–100)
• A given geometry can support different radiation patterns based on the excitation
• Resistance to tolerance errors and proximity detuning

3.6 DIELECTRIC RESONATOR GEOMETRIES

To be useful in practical antenna designs, the geometry selection of the DR is very important. Cylindrical and hemispherical DRAs are well-known for their mechanical simplicity and easy analysis. Hemispherical geometry also offers the advantage of simple interface between the dielectric and the air, compared to other geometries [13]. Also there is only one design parameter (the diameter) for the aforesaid geometry. But they exhibit mode degeneracy which can increase the cross-polarisation levels of an antenna. This problem can be avoided in a rectangular DR by properly choosing its three dimensions. Also any of the two aspect ratios can be chosen independently for a given frequency of operation [12] of a rectangular DR. Several other geometries such as ring, triangular, split-cylinder, cross, conical, elliptical, hexagonal etc. have also been evolved. For a cylindrical-ring DR, the radiation Q-factor is lower than that of corresponding cylindrical DR for certain modes, facilitating more bandwidth [4, 25]. Triangular DR [14] has an advantage over the circular and rectangular ones in that it is smallest in size when the DRs have the same dielectric constant, thickness, and
operating frequency [15]. Reduced volume DR designs in the shape of split cylinder are also available [16, 17] which considerably allows low-profile, low-volume antenna applications.

Figure 7: DR geometries used as antennas
Geometries like conical [18], stair [19], stacked triangular [20] etc. emerged for dual-band or wideband applications while those like cross [25], elliptical [21], hexagonal [22], cylindrical-comb [23] etc. came out for circular polarisation applications. Figure 7 shows the DR geometries explained so far.

3.7 MICROWAVE POWER COUPLING TECHNIQUES

Microwave power is coupled to the DR through the feed. The feed geometry and its relative position determine the type and strength of the mode excited in the DR, on which the radiation pattern relies. Numerous feeding techniques are available in the literature [26–37]. Some of the commonly used techniques are discussed below

3.7.1 Coaxial Probe

This is the simplest means for coupling energy to a DR. As shown in Figure 8(a), the DR is placed on a conducting ground plane and the central conductor of a coaxial connector extends from the bottom to the top plane to make contact with the DR. The outer conductor of the connector makes contact with the ground plane. The probe can be placed either touching the periphery of the DR or inside a hole drilled on the bottom face of the DR [26]. Amount of coupling can be controlled by varying the probe position and/or length with respect to the DR.

When the probe is at or near the periphery of the cylinder, the broadside \( \text{HEM}_{118} \) mode is excited while when the probe is inserted at the centre, the
monopole -mode TM_{01}\textsubscript{2} is excited. However this kind of coupling requires drilling a hole through the DR especially when its dielectric constant is low, which is very difficult in practice. Any direct radiation from the probe can increase the cross-polarisation in the \textit{H}-plane of the DRA [12]. Also the probe introduces ohmic
loss and self-reactance at higher frequencies [13]. In addition, the air gap between the probe and the DR can prominently affect the DRA performance [24].

3.7.2 Microstrip Transmission Line

This kind of feed, which is compatible with microwave integrated circuits (MICs) couples energy magnetically to the DR. 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 metalised. Advantages of microstrip feed include easier fabrication, matching and analysis. Also these lines allow simple transition to/from coaxial circuits. The feed is shown in Figure 8(b).

Energy coupling and hence and the input impedance of the DRA is set by the relative position between the DR and the strip [27, 28]. Such a feed is shown to be more convenient with DRA arrays [29]. One disadvantage of the microstrip feed is that at higher frequencies, surface wave modes are also excited in the substrate which adversely affect the radiation pattern and efficiency of the DRA [1].

3.7.3 Slot or Aperture Feed

In slot-fed DRA, shown in Figure 8(c), a narrow slot is formed on the ground plane of the previous structure (Figure 8(b)), through which energy is coupled to the DR. The slot acts as a magnetic current element perpendicular to the microstrip. The magnetic coupling through the slot avoids the drawbacks of the probe coupling [13]. This also has the advantage of isolating the radiator from
the feed as well as blocking the spurious radiations from the strip [30]. Also slot coupling provides low cross-polarisation level since both the slot and the DR radiate like horizontal magnetic dipoles [12].

Length of the microstrip stub that extends beyond the slot can be used to cancel the reactance of the slot, thus allowing good impedance matching. Also this feed is well-suited in monolithic microwave integrated circuits (MMICs). But at lower frequencies, the size of the slot becomes large so that such coupling is advised at higher frequencies [31].

3.7.4 Co-planar Feed

Here, both the feed and the ground plane are etched on the same side of a substrate [32, 33]. Figure 8(d) shows the details of a co-planar slot fed DRA. This kind of feed has been found the most suitable for MMICs, arrays, circularly polarised antennas, dual-frequency structures, wide-band structures, and active antennas [33]. Impedance matching is set by the geometry and the dimensions of the slot.

3.7.5 Waveguide Feed

The primary advantage of a waveguide is that it is extremely less lossy in the millimeter wave and higher frequencies. 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 [34, 35]. Coupling to the DR can be achieved through a probe [36] or a slot [37]. A waveguide probe fed DRA is shown in Figure 8(e).

3.8 APPLICATIONS OF DRs

Based on the above discussions, the key features of DRs which make them essential components in various microwave and millimeter wave systems [38] like satellite antenna, multi-channel microwave communications systems, radar systems, mobile phone systems, measuring equipments etc. are:

- High-purity, high-density ceramics minimise loss
- High dielectric constant makes possible the miniaturisation of components
- Temperature-compensated dielectric constant enables stable resonance
- A variety of shapes and coupling schemes are available for custom application requirements

These features enable DRs to be used as part of MIC/MMIC structures, forming high performance and highly stable oscillators, filters, duplexers, frequency discriminators etc. [39–42]. Recently, the application of DR as a loading element to the patch antennas and/or as a pure radiator [43] has received much attention. In addition, hollow DRs are used in active antenna applications simultaneously as the radiator and as the packaging cover, hence serving a dual-
function [44], and when also used as an oscillator load, it serves a tri-function [45]. Simultaneous use of a single cylindrical DR as an antenna as well as a filter has been recently reported [46]. Apart from antenna application, DR arrays are used as part of spatial millimeter wave power combining system [47]. DRs are also being used for material characterisation, by coating the DR with the material in the form of thin film [48], nanotubes or nanowires [49].

3.9 MEASUREMENT SETUP

Measurements of the characteristics of the DR and the radiation properties of the DRAs are carried out in the Center for Research in Electromagnetics and Antennas (CREMA), Department of Electronics, Cochin University of Science And Technology.

The basic measurement setup comprises of

- Network Analyser Unit
- Automated Antenna Positioner
- Standard Antenna
- Device-Under-Test (DUT)

When measuring DUTs like cavities, only the network analyser unit is used, but for antenna measurement, all the above four units are used.
3.9.1 Network Analyser Unit

Network analyser is a sophisticated instrument, generally used to measure the reflection and transmission of signals associated with an electrical network, especially at higher frequencies. The HP 8510C, a fully integrated vector network analyser (VNA) system, is used in the present study. It measures the magnitude and phase characteristics of electronic networks and components such as filters, amplifiers, attenuators and antennas. The instrument has four inputs, two independent measurement channels, and an internal microcomputer to automate measurements, conduct data processing, display results, and manage data input-output operations. The dedicated system bus provides fast digital communication between individual system instruments, allowing the network analyser to fully use the source and test set capabilities.

Figure 9: HP 8510C network analyser system
The minimum configuration [50] consists of a source, a test set, and the network analyser. Figure 9 shows the basic system with three major instruments.

The source uses an HP 83651B synthesised sweep oscillator to provide the RF signal. It combines the high performance and accuracy of a synthesised source with the speed and versatility of a sweep oscillator. The sweep oscillator uses an open-loop YIG (Yttrium Iron Garnet) - tuned source. It provides broadband frequency coverage (10 MHz to 50 GHz) with a precise frequency resolution of 1 Hz.

The test set HP 8517 B separates the signal produced by the source into an incident signal, sent to the DUT, and a reference signal against which the transmitted and reflected signals are later compared. The test set also routes the transmitted and reflected signals from the DUT to the receiver (IF/detector). It operates over the 45 MHz to 20 GHz range. Also two 90 dB step attenuators, which allow control of the port 1 and port 2 signal levels, are also built in the analyser.

The network analyser includes the HP 85101 display/processor and the 85102 IF/detector. The detector, together with the display/processor, processes the signals. Using its integral microprocessor, it performs accuracy enhancement and displays the results in a variety of formats.

The analyser uses a 32-bit Motorola 68000 microprocessor equipped with 1 MB of RAM, and 512 KB of EEPROM. A firmware operating system is stored
permanently in the ROM and then loaded into RAM each time power is applied. Peripheral devices such as a controller PC, printer, plotter, and disc drive can be interfaced with the analyser via GPIB ports.

In a typical measurement [51], the signal source is swept from the lower measurement frequency to the higher measurement frequency using a linear ramp controlled by the 8510. Ramp sweep offers the fastest update of the measurement display. In step-sweep mode, the source is phase-locked at each discrete measurement frequency controlled by the 8510. At the first frequency conversion stage, signal separation components in the test set apply a portion of the incident signal and the responses from the DUT to the first stage. Digital communication between the receiver and the test set pre-tunes the 65 MHz to 300 MHz voltage-tuned local oscillator (VTO) so that one of its harmonics mixes with the stimulus to produce a 1st IF frequency close to 20 MHz. Fine tuning is accomplished by comparing the IF frequency with the internal 20 MHz crystal reference and sweeping the local oscillator to track the stimulus frequency. When the local oscillator reaches its upper frequency limit, the sweep is stopped, the local oscillator is tuned again, phase lock is reestablished, and the sweep is continued. Since the first local oscillator frequency is selected algorithmically from a known stimulus frequency, the measurement is free of harmonic skip.

The second frequency conversion produces an IF frequency of 100 KHz for application to the detection and data processing elements of the receiver. Because the frequency conversions are phase-coherent and the IF signal paths are carefully matched, magnitude and phase relationships between the input signals
are maintained throughout the frequency conversion and detection stages. Automatic, fully calibrated, auto-ranging IF gain steps maintain the IF signal at optimum levels for detection over a wide dynamic range.

3.9.2 Automated Antenna Positioner

This assembly is used for the far-field measurement of the antenna-under-test (AUT). It consists of a stepper motor with gear system for rotating a circular platform over 360° on which the AUT is attached. The height of the platform can be adjusted vertically for aligning the axis of the AUT with that of the standard antenna. The platform can be rotated to any desired angle by using the controller S310C, built by the Sophisticated Test and Instrumentation Center, Cochin University of Science And Technology. The angular position can be precisely controlled either manually or by a MATLAB® program stored in the PC, interfaced with both the analyser and the controller.

3.9.3 Standard Antenna

The DRH-0118 broadband double-ridged horn antenna is used as the standard (STD) antenna that is shown in Figure 10. It is linearly polarised and operates over a frequency range of 1 to 18 GHz. These antennas have high gain, bandwidth, and power handling characteristics. They have low dispersion when used with short-pulse signals. The coaxial input to the antenna is easily adaptable to many modern network analysers. The antenna is fabricated from aluminium alloys and RT/duroid and all supporting hardwares are non-corroding for reliable
operation and long term durability in both indoor and outdoor applications. A universal mounting bracket is supplied which allows the antenna to be positioned axially in 22.5 degree increments for polarisation-sensitive measurements. The mounting bracket also has provisions for tripod attachment with a standard 1/4–20 UNC threaded fitting.

![Photograph of the double ridged horn antenna](image)

Figure 10: Photograph of the double ridged horn antenna

3.9.4 Antenna under Test

The AUT is the antenna designed using the fabricated DR. The fabrication process has been explained in section (3.3). A few DR samples are fabricated and characterised. For the antenna study, three different DRs of dielectric constant $\varepsilon_r$, diameter $2a$ and height $d$ are used. Two of them are made of zinc titanate (ZnTiO$_3$) and the third is of titanium dioxide (TiO$_2$). The DR specifications are given in Table 1. Measured results show that titanium based compounds are having poor $\tau_f$. But, when the easy availability and low cost of the titanates are taken into account, these are utilised in the present work.
Table 1: DRs used for AUT

<table>
<thead>
<tr>
<th>DR</th>
<th>Diameter $2a$ mm</th>
<th>Height $d$ mm</th>
<th>Density (mass/volume), g/cm$^3$</th>
<th>Dielectric constant, $\varepsilon_r$</th>
<th>$Q$-factor</th>
<th>Temp. coeff. of res. freq $\tau_0$, ppm/°C</th>
</tr>
</thead>
<tbody>
<tr>
<td>DR-1</td>
<td>24</td>
<td>7.3</td>
<td>4.69</td>
<td>20.8</td>
<td>6558 at 3.94 GHz</td>
<td>-47.9</td>
</tr>
<tr>
<td>DR-2</td>
<td>27.3</td>
<td>8.4</td>
<td>4.56</td>
<td></td>
<td>''</td>
<td>''</td>
</tr>
<tr>
<td>DR-3</td>
<td>24</td>
<td>7.8</td>
<td>3.84</td>
<td>88.68</td>
<td>3100 at 2.45 GHz</td>
<td>+455.4</td>
</tr>
</tbody>
</table>

The DRA is fed by a 50 $\Omega$ microstrip transmission line of width ($w$) 3 mm and length 50 mm fabricated on one side of a 1.6 mm thick ($h$) microwave substrate of dielectric constant $\varepsilon_r = 4$, size 115 mm x 115 mm and having a copper cladding on the opposite side. The design equations for the feed are given as Eq. (3.21) and (3.22) where an effective dielectric constant ($\varepsilon_{\text{eff}}$) is used instead of the absolute value in the equation for the characteristic impedance ($Z_0 = 50 \, \Omega$). The merits of using microstrip feed are well-known, where fine adjustment of impedance matching between the feed and the DR can be easily achieved by adjusting the DR position relative to the feed.

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \frac{1}{\sqrt{1 + \frac{12h}{w}}} + 0.04 \left( \frac{1}{1 - \frac{w}{h}} \right)^2 \right], \quad \text{if } \frac{w}{h} < 1
\]

\[
= \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \frac{1}{\sqrt{1 + \frac{12h}{w}}} \right], \quad \text{otherwise}
\]
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\[ Z_o = \frac{60}{\sqrt{\epsilon_{\text{eff}}}} \ln \left( \frac{8h + w}{w + 4h} \right) , \text{ if } \frac{w}{h} < 1 \]

\[ = \frac{120\pi}{\sqrt{\epsilon_{\text{eff}}}} \frac{1}{\left( \frac{w}{h} + 1.393 + 0.677 \ln \left( \frac{w}{h} + 1.444 \right) \right)} , \text{ otherwise} \]  

\[ (3.22) \]

Photographs of the DRAs under test are shown in Figures 11 and 12. Figure 11 shows the fundamental DRA geometry that yields a narrow bandwidth and broadside radiation pattern. The above geometry is modified into that shown
in Figure 12 for a wide bandwidth and conical radiation pattern. Further details of the design are presented in Chapter 4.

As explained in section (3.2.3), the impedance bandwidth of a DRA is inversely proportional to its unloaded Q-factor \((Q_u)\) which in turn is directly related to the radiation Q-factor \((Q_r)\), provided the conductor and dielectric losses are negligible compared to the radiated power. The stored energy is directly proportional to the volume of the DR while the radiated power depends on its surface area in a similar way. In that case we can write [1],

\[
Q_r = \frac{2\omega_0 E_s}{P_r} \propto (\varepsilon_r)^p \left( \frac{\text{DR Volume}}{\text{DR Surface Area}} \right) \tag{3.23}
\]

where \(p = 1.5\) for magnetic dipole like modes

\(= 2.5\) for electric dipole like modes

where \(\omega_0\) is the resonant angular frequency, \(E_s\) is the peak stored energy and \(P_r\) is the radiated power. For a cylindrical DRA of a given volume, the arrangement shown in Figure 11 allows the top and side walls of the cylinder to radiate freely but the radiation from the bottom wall, lying on the substrate is restricted by the ground plane. If the bottom wall of the cylinder is also allowed to radiate as shown in Figure 12, the \(Q_r\) can be considerably reduced as per Eq. (3.23) to accomplish a wideband operation.
3.10 **MEASUREMENT PROCEDURE**

Measurement setup is similar to that in Figure 9, except for the DUT, which is substituted by a transmission cavity for DR characterisation or by the AUT connected at one port and the standard antenna at the other port for antenna characterisation. The arrangement for antenna measurement is shown in Figure 13. The analyser is interfaced to a computer that runs a MATLAB® based software to measure the antenna parameters.

![Diagram](image)

*Figure 13: Arrangement for antenna measurement*

### 3.10.1 Impedance and VSWR

These are the most important parameters as far as an antenna is concerned. Impedance of the antenna is a measure of the efficiency with which it acts as a
transducer between the source and the propagating medium. The impedance is a complex quantity with a real part, called the antenna resistance and the imaginary part, called the antenna reactance. When there is an impedance mismatch between the antenna and the source line, a part of the incident energy is reflected back to the source. The ratio of the reflected voltage (or current) to the incident voltage (or current) is termed as the input reflection coefficient ($\Gamma$). The magnitude $|\Gamma|$, when expressed in dB, it is called the return loss ($|S_{11}|$) of the antenna.

$$|S_{11}| = 20 \log_{10} (|\Gamma|) \quad (3.24)$$

The level of mismatch is also defined in terms of the voltage standing wave ratio ($VSWR$) defined as the ratio of the voltage maximum to minimum of the standing wave existing on the antenna input terminal.

$$VSWR = \frac{V_{\text{max}}}{V_{\text{min}}} = \frac{(1+|\Gamma|)}{(1-|\Gamma|)} \quad (3.25)$$

Once the AUT is connected to port 1, the frequency range is selected from the stimulus menu on the front panel of the analyser. The reflection mode is selected from the parameter menu. The display then gives the return loss of the AUT as a function of the frequency. By using markers, the resonant frequency, bandwidth etc. can be calculated. Usually the bandwidth is measured between the 2:1 $VSWR$ or $-10$ dB $|S_{11}|$ points of the plot. From the format menu, phase of $S_{11}$, $VSWR$, real and imaginary parts of the input impedance etc. can also be plotted.
3.10.2 Radiation Pattern

Radiation pattern of an antenna is the graphical representation of its radiation properties as a function of the space co-ordinates, which implies a three dimensional pattern. Because of the limits set by a practical measurement setup for measuring the 3-D pattern, usually two principal plane patterns are specified for antennas radiating in the broadside and three patterns for those with omnidirectional patterns. Generally, far-field patterns are specified for an antenna where the pattern is measured at a distance, \(d > 2D^2/\lambda\), where \(D\) is the largest dimension of the antenna and \(\lambda\) is the operating wavelength.

As shown in Figure 13, the AUT is connected to port 1 and the STD antenna is connected to port 2 of the analyser. The height and polarisation of both antennas are then aligned for maximum transmission (\(|S_{21}|\)) between them. The frequency range over which \(|S_{11}| < -10\ dB\) is selected using the parameter menu. Now thru calibration of the analyser is selected from the calibrate menu. This calibrates the \(|S_{21}|\) data to 0 dB for every frequency point in the band. In order to suppress the spurious reflections from the nearby objects, the time domain gating facility of the analyser is used. The gate span is selected according to the largest dimension of the radiator.

The positioner is now set to home, which automatically sets the current angular position of the antenna as 0°. The computer software now invokes the radiation pattern routine and reads the normalised \(|S_{21}|\) data for the specified frequency band, as a function of the angular position of the AUT.
3.10.3 Antenna Gain

Gain is the logarithm of the ratio of the intensity of an antenna’s radiation pattern in the direction of strongest radiation to that of a reference antenna, when both antennas are fed with the same input power. If the reference antenna is isotropic, the gain is often expressed in units of dBi. The gain of an antenna is a passive phenomenon — power is not added by the antenna, but simply redistributed to provide more radiated power in a certain direction than would be transmitted by an isotropic antenna. In this thesis, the gain transfer method is used to calculate the absolute gain of the AUT. The experimental setup for gain measurement is the same as that for radiation pattern measurement. Here, an antenna of known gain $G_{\text{ref}}$ (dBi) is used as the reference antenna. Initially, the AUT for the pattern measurement setup is replaced by the reference antenna. It is then positioned for maximum radiation in the direction of the STD antenna and the transmission coefficient $|S_{21}|_{\text{ref}}$ (dB) is displayed on the analyser. A thru calibration is performed and the data is stored in the cal set. This is the reference gain for the AUT. Now the reference antenna is replaced with the AUT and the transmission coefficient $|S_{21}|_{\text{AUT}}$ (dB) is recorded, which gives the relative gain. The absolute gain can then be calculated as

$$G \text{ (dBi)} = G_{\text{ref}} \text{ (dBi)} + |S_{21}|_{\text{AUT}} \quad (3.26)$$
3.10.4 Radiation Efficiency

Radiation efficiency of an antenna quantifies the resistive loss of the antenna in terms of the proportion of power radiated versus the power fed to the antenna.

\[
\text{Radiation Efficiency, } \eta = \frac{P_{\text{rad}}}{P_{\text{in}}} = \frac{P_{\text{rad}}}{P_{\text{rad}} + P_{\text{loss}}} = \frac{R_{\text{rad}}}{R_{\text{rad}} + R_{\text{loss}}} \tag{3.27}
\]

Where

- \( P_{\text{rad}} \) = power radiated (W)
- \( P_{\text{in}} \) = power fed to antenna (W)
- \( P_{\text{loss}} \) = power lost by the antenna (W)
- \( R_{\text{rad}} \) = radiation resistance of the antenna (\( \Omega \))
- \( R_{\text{loss}} \) = loss resistance of the antenna (\( \Omega \))

For physically small antennas, the Wheeler cap method [52] 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 its near fields as illustrated in Figure 14, 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 [53]. 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 as in Eq. (3.27).

Another approach to the Wheeler cap measurement is by measuring the Q-factor of the antenna using the formula given below.

\[
\eta = \frac{P_{rad}}{P_{rad} + P_{loss}} = 1 - \frac{Q_o}{Q_{loss}} \tag{3.28}
\]

where \(Q_o\) = total Q-factor

\(Q_{loss}\) = dissipation Q-factor

Figure 14: Wheeler cap method for measuring radiation efficiency
The Q-factor can be related to the $\pm 10$ dB bandwidth as,

$$Q = \frac{S - 1}{BW \sqrt{S}} \times 100\%$$  \hspace{1cm} (3.29)

The bandwidth, $BW$ is expressed as the ratio of the frequency bandwidth over which $VSWR$ is below a specific value $S$ (usually 2), to the centre frequency. Thus, without the cap, the measured bandwidth corresponds to the $Q_o$. When the cap is placed, the radiation is suppressed and the measured bandwidth represents $Q_{int}$. The size and shape of the cap is not critical in the above method. But the centering of the cap with respect to the radiator and good electrical contact between the cap and the antenna ground plane are very important [51].

**CONCLUSION**

The fabrication process of dielectric resonators, their general characteristics and microwave characterisation were discussed in this chapter. A detailed account of the measurement facilities used, the proposed antenna design and the measurement procedure of important antenna properties were also described.
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