1.1. INTRODUCTION

A microstrip radiator (antenna) consists of a conducting patch on a ground plane, which is separated by a dielectric substrate of relative permittivity in the range of 1.1 to 12. However, low dielectric constant substrates are generally chosen for maximum radiation. The concept of such antennas was undeveloped until the revolution in electronic circuit miniaturization and large-scale integration in 1970. While, the early research work of Munson on microstrip antennas for use as a low profile flush mounted antennas on rockets and missiles has proven that these antennas can be used for various purposes [1]. The radiating patch and the feed lines are usually photo etched on a dielectric substrate. The basic configuration of a microstrip antenna is shown in Figure 1.1.

![Figure 1.1 Basic configuration of a microstrip patch antenna](image)

Various mathematical models have been developed for this antenna and its applications being extended to many other fields. The number of research work carried on these antennas shows its gaining importance in microstrip technology. The conducting patch of the microstrip antenna can take any shape; rectangular, square, circle, triangular, etc but rectangular and circular configurations are the mostly preferred, Figure 1.2.
For a rectangular patch, the length $L$ of the patch is usually $0.3333 \lambda_o < L < 0.5 \lambda_o$, where $\lambda_o$ is the free-space wavelength. The patch is selected to be very thin such that $t << \lambda_o$ (where $t$ is the patch thickness). The height $h$ of the dielectric substrate is usually $0.003 \lambda_o \leq h \leq 0.05 \lambda_o$. This is because other structures are difficult to analyze. A microstrip antenna is characterized by its input impedance, gain, bandwidth, efficiency and radiation patterns. The length of the antenna is nearly half wavelength in the dielectric; it is a very important parameter, which governs the resonant frequency of the antenna hence other parameters.

Microstrip patch antennas radiate primarily because of the fringing fields between the patch edges and the ground plane. For good antenna performance, a thick dielectric substrate having a low dielectric constant is desirable since this provides better efficiency, larger bandwidth and better radiation [2]. However, such a configuration leads to a larger antenna profile. In order to design a compact microstrip patch antenna, substrates with higher dielectric constants must be used which are less efficient and result in narrower bandwidth. Hence a trade-off must be realized between the antenna geometry and antenna performances.

1.2. RADIATION MECHANISM OF A MICROSTRIP PATCH ANTENNA

Consider Figure 1.3 shown below when a microstrip patch is provided power, a charge distribution is seen on the upper and lower surfaces of the patch and at the bottom of the ground plane. This charge distribution is controlled by two mechanisms-an attractive
mechanism and a repulsive mechanism. The attractive mechanism is between the opposite charges on the bottom side of the patch and the ground plane, which helps in keeping the charge concentration intact at the bottom of the patch.

![Figure 1.3 Radiation mechanism of microstrip patch antenna](image)

The repulsive mechanism is between the like charges on the bottom surface of the patch, which causes pushing of some charges from the bottom, to the top of the patch. As a result of this charge movement, currents flow at the top and bottom surface of the patch.

In the Figure 1.3 the microstrip patch antenna is represented by two slots, separated by a transmission line of length $L$ and open circuited at both the ends. Along the width of the patch, the voltage is maximum and current is minimum due to the open ends. The fields at the edges can be resolved into normal and tangential components with respect to the ground plane. However the normal components of the electric field at the two edges along the width are in opposite directions and thus out of phase since the patch is $\lambda/2$ long and hence they cancel each other in the broadside direction. The tangential components, which are in phase, means that the resulting fields combine to give maximum radiated field normal to the surface of the structure. Hence the edges along the width can be represented as two radiating slots, which are $\lambda/2$ apart and excited in phase and radiating in the half space above the ground plane. The fringing fields along the width can be modeled as radiating slots and electrically the patch microstrip patch antenna looks greater than its physical dimensions. The cavity model assumes that the height to width ratio (i.e. height of substrate and width of the patch) is very small and as a result of this the attractive mechanism dominates and causes most of the charge concentration and the current to be
below the patch surface and much less current would flow on the top surface of the patch and as the height to width ratio further decreases, the current on the top surface of the patch would be almost equal to zero. It would not allow the creation of any tangential magnetic field components to the patch edges. Hence, the four sidewalls could be modeled as perfectly magnetic conducting surfaces, which imply that the magnetic fields and the electric field distribution beneath the patch would not be disturbed. However, in practice, a finite width to height ratio would be there and this would not make the tangential magnetic fields to be completely zero, but they being very small, the side walls could be approximated to be perfectly magnetic conducting. Since the walls of the cavity, as well as the material within it are lossless, the cavity would not radiate and its input impedance would be purely reactive. Therefore, in order to account for radiation and a loss mechanism, one must introduce a radiation resistance $R_R$ and a loss resistance $R_L$. A loss cavity would now represent an antenna and the loss is taken into account by the effective loss tangent $\delta_{eff}$ which is given as [2]:

$$\delta_{eff} = \frac{1}{Q_T}$$  \hspace{1cm} (1.1)

Where $Q_T$ is the total quality factor of antenna and has been expressed in the form

$$\frac{1}{Q_T} = \frac{1}{Q_d} + \frac{1}{Q_e} + \frac{1}{Q_c}$$  \hspace{1cm} (1.2)

Where $Q_d$ represents the quality factor of the dielectric and is given by

$$Q_d = \frac{\omega_r W_T}{P_d} = \frac{1}{\tan\delta}$$  \hspace{1cm} (1.3)

Where

$\omega_r$ = angular resonant frequency

$W_T$ = total energy stored in the patch at resonance

$P_d$ = dielectric loss

$tan\delta$ = loss tangent of the dielectric substrate

$Q_c$ = quality factor of conductor and given as:
\[ Q_c = \frac{\omega_r W_T}{P_c} = \frac{h}{\Delta} \]  

(1.4)

Where

- \( P_c \) = conductor loss
- \( \Delta \) = skin depth of the conductor
- \( h \) = height of dielectric substrate and
- \( Q_r \) represents the quality factor for radiation and is given as:

\[ Q_c = \frac{\omega_r W_T}{P_r} \]  

(1.5)

Where \( P_r \) is the power radiated from the patch

\[ \delta_{eff} = \tan\delta + \frac{\Delta}{h} + \frac{P_r}{\omega_r W_T} \]  

(1.6)

Thus, the equation (1.6) describes the total effective loss tangent for the microstrip patch antenna.

### 1.3. ADVANTAGES AND DISADVANTAGES

Microstrip patch antennas are increasing in popularity for use in military and commercial applications mainly due to its low-profile. Therefore they are extremely compatible for embedded antennas in handheld wireless devices such as cellular phones, pagers etc... The telemetry and communication antennas on missiles need to be thin and conformal and are often in the form of patch antennas. Another area where they have been used successfully is satellite communication. Some of their principal advantages are given as;

- Light weight and low volume
- Low profile planar configuration which can be easily made conformal to host surface
- Low fabrication cost, hence can be manufactured in large quantities
- Supports both, linear as well as circular polarizations
- Can be easily integrated with microwave integrated circuits (MICs)
- Capable of dual and triple frequency operations
- Mechanically robust when mounted on rigid surfaces

Despite of these advantages, microstrip patch antennas suffer from some drawbacks as compared to conventional antennas, which are given as:

- Narrow bandwidth, low gain and low efficiency
- Alteration in resonance frequency in environmental conditions
- Extraneous radiation from feeds and junctions
- Poor end fire radiator except tapered slot antennas
- Low power handling capacity and surface wave excitation

In addition, microstrip patch antennas have a very high antenna quality factor ($Q$-factor), which represents the losses associated with the antenna. The large $Q$ also leads to narrow bandwidth and low efficiency. The $Q$-factor of the microstrip antenna can be reduced by increasing the thickness of the dielectric substrate. But as the thickness increases, an increasing fraction of the total power delivered by the source goes into the surface waves, as dissipation i.e. power loss.

1.4. APPLICATIONS OF MICROSTRIP PATCH ANTENNA

Microstrip patch antennas are found suitable for the applications in the various areas such as in the medical applications, satellites and of course even in the military systems just like in the rockets, aircrafts missiles etc. The usage of the microstrip antennas is spreading widely in all the fields and areas and now they are booming in the commercial aspects due to their low cost of the substrate material and the fabrication. It is also expected that due to the increasing usage of the patch antennas in the wide range could take over the usage of the conventional antennas for the maximum applications.

Some of the applications for the microstrip patch antenna are in the following system/devices:
1.5. FEEDING TECHNIQUES

In order to excite for radiation, a microstrip patch antenna is fed by a variety of methods. These methods can be classified into two categories: - contacting and non-contacting [3]. In the contacting method, the RF power is fed directly to the radiating patch using a connecting element such as a coaxial and microstrip line feeds. In the non-contacting scheme, electromagnetic field coupling is used to transfer power between the microstrip line and the radiating patch. However the four most popular feed techniques used for exciting the patch antennas are the microstrip line, coaxial probe (both contacting schemes), aperture coupling and proximity coupling (both non-contacting schemes).

1.5.1. Microstrip Line Feed

In this type of feed technique, a conducting strip is connected directly to the edge of the patch of microstrip antenna (Figure 1.4). The conducting strip in this case should be smaller than width compared to the patch. Such kind of feed techniques have the advantage that the feed can be etched on the same substrate providing a planar structure. The purpose of the inset cut in the patch is to provide impedance matching between the feed line and patch without adding any other matching element. This is achieved by properly controlling the inset location. That is, it is an easy feeding approach, as it provides ease of fabrication and simplicity in modeling as well as impedance matching.
However as the thickness of the dielectric substrate increases, surface waves and spurious feed radiation also increases, which hampers the bandwidth of the antenna. The feed radiation also leads to undesired cross polarized radiation.

1.5.2. Coaxial Feed

A coaxial feed also known as probe feed, is a very common feeding technique used to excite microstrip patch antennas. As can be seen, in Figure 1.5, the inner conductor of the coaxial connector extends through the dielectric and then soldered to the radiating patch, while the outer conductor is connected to the ground plane. The main advantage of this type of feeding approach is that the feed can be placed at any desired position inside the patch in order to achieve impedance matching.
This feed technique is easy to execute as well as has low spurious radiation. However, a disadvantage is that it provides narrow bandwidth and is difficult to model since a hole need to be drilled in the substrate and the connector protrudes outside the ground plane, hence not making it completely planar for thick substrates \((h > 0.02\lambda_0)\). Also, for thicker substrates, the increased probe length makes the input impedance more inductive, leading to impedance matching issue. Earlier it is seen that for a thick dielectric substrate, which provides broad bandwidth, the microstrip line feed and the coaxial feed, both suffer from numerous disadvantages. Therefore the non-contacting feed techniques are preferred to solve these problems.

1.5.3. Aperture Coupled Feed

In this type of feed technique, the radiating patch and the microstrip feed line are separated by the ground plane, and coupling between the patch and the feed line is made through a slot or an aperture in the ground plane, shown in Figure 1.6. The coupling aperture is usually centered under the patch, leading to lower cross polarization due to symmetry of the configuration.

![Figure 1.6 Aperture-coupled feed of microstrip patch antenna](image)

The amount of coupling from the feed line to the radiating patch is determined by the shape, size and location of the aperture. Since the ground plane separates the patch and the feed line, spurious radiation is minimized. Generally, a high dielectric constant material is
used for bottom substrate; however a thick, low dielectric constant material is used for the top substrate to optimize radiation from the patch. The major disadvantage of this feed approach that it is difficult in fabrication of the patch antenna due to multiple layers, which also increases the antenna thickness and it provides the narrow bandwidth.

1.5.4. Proximity Coupled Feed

This feed technique is also called as the electromagnetic coupling feeding. As shown in Figure 1.7, two dielectric substrates are used such that the feed line is between the two substrates and the radiating patch is on top of the upper substrate. The advantage of this feed method is that it eliminates spurious feed radiation and provides very high bandwidth (nearly 13%), due to overall increase in the thickness of the microstrip patch antenna. This scheme also provides choices between two different dielectric media, one for the patch and another for the feed line to optimize the individual performances. Matching can be achieved by controlling the length of the feed line and the width to line ratio of the radiating patch.

![Proximity-coupled feed of microstrip antenna](image)

Figure 1.7 Proximity-coupled feed of microstrip antenna

The main disadvantage with feed scheme that, it is difficult to fabricate because of the two dielectric layers which need proper alignment. As well as there is an increase in the overall thickness of the antenna structure.

1.6. FUNDAMENTAL PARAMETERS OF ANTENNAS

There are several parameters which are useful to describe the performances of an antenna; they are given as;
**Radiation Pattern:** It is a graphical representation of the radiation properties of the antenna as a function of space coordinates. In most of cases the radiation pattern is determined in the far-field region.

**Directivity:** It is the ratio of radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions. If the direction is not specified, the direction of maximum intensity is implied.

**Gain:** The relative gain is the ratio of the power gain in a given direction to the power gain of a reference antenna in its referenced direction. In most of cases the reference antenna is a lossless isotropic source. When the direction is not specified, the power gain is usually taken in the direction of maximum radiation.

**Efficiency:** Antenna radiation efficiency is defined as the ratio of power radiated to the input power. It relates the gain and directivity. Radiation efficiency also takes into account conduction and dielectric losses.

**Bandwidth:** The range of frequencies within which the performance of the antenna conforms to a specified standard. For narrowband antennas, the bandwidth is expressed as a percentage of the frequency difference over the centre frequency. However, if the impedance is the limiting factor, the bandwidth is defined in terms of the Q.

**Polarization:** The polarization of an antenna in a given direction is the polarization of the wave radiated by the antenna. When the direction is not stated, the polarization is taken to be the polarization in the direction of maximum gain. Polarization describes the time varying direction and relative magnitude of the E-field.

**Input Impedance:** Impedance presented by an antenna at its terminals, or ratio of the voltage and current at a pair of terminals, or ratio of the appropriate components of the electric and magnetic field at a point.

### 1.7. METHODS OF ANALYSIS OF PATCH ANTENNAS

The preferred models for the analysis of microstrip patch antennas are the transmission line model, cavity model and full wave model (which include primarily integral equations/Moment Method). The transmission line model is the simplest of all and it gives good physical insight but it is less accurate. The cavity model is more accurate and gives good physical insight but is complex in nature. The full wave models are extremely accurate, versatile and can treat single elements, finite and infinite arrays, stacked
elements, arbitrary shaped elements and coupling. These give less insight as compared to the two models mentioned above and are far more complex in nature.

1.7.1. Transmission Line Model

The microstrip is essentially a non homogeneous line of two dielectrics, typically the substrate and air. This transmission line model represents the microstrip antenna by two slots of width $W$ and height $h$, separated by a transmission line of length $L$ [4, 5].

![Figure 1.8 (a) Electric field lines](image)

Hence, as seen from Figure 1.8 (a, b), most of the electric field lines reside in the substrate and parts of some lines in air. As a result, this transmission line cannot support pure transverse-electric-magnetic (TEM) mode of transmission, since the phase velocities would be different in the air and the substrate. Instead, the dominant mode of propagation would be the quasi-TEM mode. Hence, an effective dielectric constant ($\varepsilon_{\text{reff}}$) must be obtained in order to account for the fringing and the wave propagation in the line. The value of $\varepsilon_{\text{reff}}$ is slightly less than $\varepsilon_r$ because the fringing fields around the periphery of the patch are not confined in the dielectric substrate but are also spread in the air as shown in Figure 1.8. The expression for $\varepsilon_{\text{reff}}$ is given as [6]:

\[
\varepsilon_{\text{reff}} = \varepsilon_r - \frac{h}{W} \left( \varepsilon_r - \varepsilon_m \right)
\]
\[ \varepsilon_{\text{reff}} = \frac{\varepsilon_r}{2} + \frac{\varepsilon_r - 1}{2} \left[ 1 + 12 \frac{h}{w} \right]^{3/2} \]  \hspace{1cm} (1.7)

Where
\[
\varepsilon_{\text{reff}} = \text{effective dielectric constant} \\
\varepsilon_r = \text{dielectric constant of substrate} \\
h = \text{height of dielectric substrate} \\
W = \text{width of the patch}
\]

Consider Figure 1.9, which shows a rectangular microstrip patch antenna of length \( L \), width \( W \) resting on a substrate of height \( h \). The co-ordinate axis is selected such that the length is along the \( x \) direction, width is along the \( y \) direction and the height is along the \( z \) direction.

In order to operate in the fundamental \( TM_{10} \) mode, the length of the patch must be slightly less than \( \lambda/2 \) where \( \lambda \) is the wavelength in the dielectric medium and is equal to \( \lambda_o / \sqrt{\varepsilon_{\text{reff}}} \) where \( \lambda_o \) is the free space wavelength. The \( TM_{10} \) mode implies that the field varies one \( \lambda/2 \) cycle along the length, and there is no variation along the width of the patch. The dimensions of the patch along its length have now been extended on each end by a distance \( \Delta L \) \cite{3}, which is empirically given by:

\[
\Delta L = 0.412 h \frac{(\varepsilon_{\text{reff}} + 0.3)(W/h + 0.264)}{(\varepsilon_{\text{reff}} - 0.258)(W/h + 0.8)} \]  \hspace{1cm} (1.8)

The effective length of the patch \( L_{\text{eff}} \) now becomes

\[
L_{\text{eff}} = L + 2\Delta L \]  \hspace{1cm} (1.9)
Hence, for a given resonance frequency $f_0$, the effective length is given as:

$$L_{eff} = \frac{c}{2f_0 \sqrt{\varepsilon_{reff}}} \tag{1.10}$$

For a rectangular microstrip patch antenna, the resonance frequency for any TM$_{mn}$ mode is given as [7]:

$$f_r = \frac{c}{2f_0 \sqrt{\varepsilon_{reff}}} \left[ \left( \frac{m}{L} \right)^2 + \left( \frac{n}{W} \right)^2 \right]^{1/2} \tag{1.11}$$

Where $m$ and $n$ are the modes along $L$ and $W$ respectively. However for efficient radiation the width $W$ is given as [3, 5]:

$$W = \frac{c}{2f_0 \sqrt{\frac{\varepsilon_{reff} + 1}{2}}} \tag{1.12}$$

### 1.7.2. Cavity Model

Although the transmission line model discussed in the previous section is easy to use, it has some inherent disadvantages. Specifically, it is useful for patches of rectangular design and it ignores field variations along the radiating edges. These disadvantages can be overcome by using the cavity model. A brief overview of this model is given as follows.

In this model, the interior region of the dielectric substrate is modeled as a cavity bounded by electric walls on the top and bottom; however the magnetic fields around the cavity. The basis for this assumption is the following observations for thin substrates ($h \ll \lambda$).

- Since the substrate is thin, the fields in the interior region do not vary much in the $z$ direction, i.e. normal to the patch.
- The electric field is $z$ directed only, and the magnetic field has only the transverse components $H_x$ and $H_y$ in the region bounded by the patch metallization and the ground plane. This observation provides for the electric walls at the top and the bottom.

Consider Figure 1.10 when the microstrip patch is provided power, a charge distribution is seen on the upper and lower surfaces of the patch and at the bottom of the ground plane.
The cavity model assumes that the height to width ratio (i.e. height of substrate and width of the patch) is very small and as a result of this the attractive mechanism dominates and causes most of the charge concentration and the current to be below the patch surface.

![Figure 1.10 Charge distributions and current density creation on the microstrip patch antenna](image)

Much less current would flow on the top surface of the patch and as the height to width ratio further decreases, the current on the top surface of the patch would be almost equal to zero, which would not allow the creation of any tangential magnetic field components to the patch edges. Hence, the four sidewalls could be modeled as perfectly magnetic conducting surfaces. This implies that the magnetic fields and the electric field distribution beneath the patch would not be disturbed. However, in practice, a finite width to height ratio would be there and this would not make the tangential magnetic fields to be completely zero, but they being very small, the side walls could be approximated to be perfectly magnetic conducting. Since the walls of the cavity, as well as the material within it are lossless, the cavity would not radiate and its input impedance would be purely reactive. Hence, in order to account for radiation and a loss mechanism, one must introduce a radiation resistance $R_R$ and a loss resistance $R_L$. A lossy cavity would now represent an antenna and the loss is taken into account by the effective loss tangent $\delta_{eff}$ which is given as: (refer equations; 1.1 to 1.6).

The most important factor to take into account at the millimeter-wave band is attenuation, which is very high, by fog, water vapor, and other gases in the atmosphere. The section presents the outline of the theory for predicting the attenuation and scattering by rain as well as atmospheric gases.
1.8. ATTENUATION BY RAIN

Radio waves propagating through rain are attenuated due to absorption of power in the lossy dielectric medium represented by water [8]. The theory for rain attenuation is based on the calculation of the absorption of a single raindrop. This calculation is straightforward for the case of a spherical droplet of water having a radius no longer than $\lambda_0/10$. In this situation the low-frequency Rayleigh scattering theory can be applied. Since the radius of raindrops ranges from a fraction of millimeters up to several millimeters, the Rayleigh scattering theory is generally valid down to wavelengths of order 3 cm or somewhat less. This assumption of spherical droplets is not valid since raindrops take on an oblate spheroid or flattened shape under the influence of aerodynamic forces and pressure forces as they fall. However, at low frequency an equivalent spherical radius can be assumed. At microwave frequencies, it is important to first consider the drop shape and then determine its cross-sections. However, with modern computer facilities and techniques, the limitation is not in the computational aspect but in knowing the drop shape, which depends upon drop size and the velocity with which the drop falls.

Let us consider a plane wave incident on a spherical drop of water with a radius such that $a \ll \lambda_0$ (as shown in Figure 1.11), then the drop is characterized as a dielectric sphere with a complex dielectric constant $k = k' - jk''$, and then incident electric field may be chosen as

$$E_i = E_0 a_x e^{-jkoz}$$  \hspace{1cm} (1.13)

Over the extent of the drop, the incident field is essentially uniform and equal to

$$E_i' = E_0 a_r$$  \hspace{1cm} (1.14)

The polarization produced in the drop is thus the same as would be produced in a dielectric sphere under the action of a uniform static electric field. Solving this case, as boundary value problem, the dipole polarization $P$ per unit volume in the drop is given by
Figure 1.11 Plane wave incidents on spherical water drop

\[ P = 3 \frac{k - 1}{k + 2} \varepsilon_0 E_0 \alpha_z \]  

(1.15)

Hence, the total dipole moment of the water sphere is obtained by multiplying by the volume and is

\[ P_T = \frac{4}{3} \pi a^3 P = 4\pi a^3 \frac{k - 1}{k + 2} \varepsilon_0 E_0 \alpha_z \]  

(1.16)

Since \( a \ll \lambda_0 \) the far-zone scattered field from the sphere is the same as that radiated by a small electric dipole of total strength \( P_T \). Since a current element \( I dl \) is equivalent to the time derivative of the dipole moments, so \( j \omega P_T \) may be used to replace \( I dl \) for the far-zone radiated field; thus

\[ E_s = -\omega Z_0 k_0 P_T \sin \theta \frac{e^{-j k_0 r}}{4\pi r} \alpha_\theta \]  

(1.17)

The scattering pattern of such type of small dielectric sphere is the same as the radiation pattern of a small electric dipole; hence the total scattered power is given by

\[ P_s = \frac{1}{2} Y_0 \int_0^{2\pi} \int_0^\pi |E_s|^2 r^2 \sin \theta \, d\theta \, d\varphi = \frac{\omega^2 k_0^2 Z_0}{12\pi} |P_0|^2 \]  

(1.18)

When we substitute \( P_T \) from equation (1.16), it gives

\[ P_s = \frac{4}{3} \pi a^2 (k_0 a)^4 Y_0 |E_0|^2 \left| \frac{k - 1}{k + 2} \right|^2 \]  

(1.19)

This is the low frequency formula for the scattered power.
In addition, the polarization current density in the sphere is \( j_p = j \omega P \) and is uniform. Thus total electric field \( E \) in the sphere will be related to \( P \) by the equation 
\[
P = (k - 1)\varepsilon_0 E \ .
\]
Hence the time average absorbed power is given by 
\[
P_a = \frac{1}{2} \text{Re} \int_0^\pi \int_0^{2\pi} \int_0^\infty E.J_p r^2 \sin \theta \ d\theta d\varphi dr = \frac{2}{3} \pi a^2 \text{Re}[E.J_p] 
\]
\[
= 6\pi a^2 k_0 Y_0 \frac{|k - 1|}{|k + 2|} \frac{k''|E_0|^2}{(k' - 1)^2 + (k'')^2} 
\]
\[
P_a = 12\pi a^2 (k_0 a) \frac{|k - 1|}{|k + 2|} \frac{k''}{(k' - 1)^2 + (k'')^2} 
\] (1.20)

In case an electromagnetic wave propagates through rain it encounters a great many water droplets with different radii. Let \( N(a) \ da \) be the number of drops per unit volume with the radii in the interval \( a \) to \( a+da \). Then the total power removed from a wave with power density \( P = \frac{1}{2} Y_0 |E|^2 \) by the drops in a volume element of unit cross-sectional area and thickness \( dz \) along \( z \) are;
\[
\frac{dP}{dz} = -\frac{1}{2} Y_0 |E|^2 \int_0^\infty \sigma_e(a)N(a)da 
\]
or
\[
= -P \int_0^\infty \sigma_e(a)N(a)da 
\] (1.21)
As a result of this power loss, the power flow decays at a rate \( 2\alpha \) where 
\[
2\alpha = \int_0^\infty \sigma_e(a)N(a)da = A \ 	ext{(say)} 
\] (1.22)
Equation (1.22) defines \( A \), which is termed as the specific attenuation per unit length along the propagation path. From equation (1.21 and 1.22) and we get
\[
\frac{dP}{dz} = -A(z).P 
\]
For which the solution is
\[
P(z) = P(0)e^{-\int_0^z A(z)dz} 
\] (1.23)
The drop size distribution may vary along the propagation path because of non uniform rain, and hence $A$ is a function of $z$, which accounts for the integral in the equation (1.21 - 1.23)

However, from view point of communication engineers what is needed is a relatively simple formula relating specific attenuation to rain rate, frequency, and temperature. Fortunately such a formula exists, and it is of the form:

$$A = \frac{aR^b dB}{km}$$

(1.24)

Where $R$ is the rain rate in millimeters per hour and $a$ and $b$ are constants that depend upon frequency and temperature of the rain. The temperature dependence is due to the variation of the dielectric constant of water with temperature. A detailed review of the theory and experimental data to a compilation of the values of two constants $a$ and $b$ has been given in [9]. They established the following empirical formulas for the constants $a$ and $b$ at a temperature of $0^0C$;

$$a = G_a f^{E_a} \quad f \text{ in GHz}$$

(1.25)

Where $G_a = 6.39 \times 10^{-5}$ \quad $E_a = 2.03$ \quad $f < 2.9GHz$

$G_a = 4.21 \times 10^{-5}$ \quad $E_a = 2.42$ \quad $2.9 GHz \leq f \leq 54GHz$

$G_a = 4.09 \times 10^{-5}$ \quad $E_a = 0.699$ \quad $54 GHz \leq f < 180GHz$

$G_a = 3.38$ \quad $E_a = -0.151$ \quad $180 GHz < f$

and

$$b = G_b f^{E_b} \quad f \text{ in GHz}$$

(1.26)

Where $G_b = 0.851$ \quad $E_b = 0.158$ \quad $f < 8.5GHz$

$G_b = 1.41$ \quad $E_b = -0.0779$ \quad $8.5 GHz \leq f < 25 GHz$

$G_b = 2.63$ \quad $E_b = -0.272$ \quad $25 GHz \leq f < 164 GHz$

$G_b = 0.616$ \quad $E_b = -0.0126$ \quad $164 GHz \leq f$
Some representative curves of attenuation in decibels per kilometers at frequencies of 10, 30, and 100 GHz as a function of rain rate were computed using equations (1.25) and (1.26) and are shown in Figure 1.12. At 10 GHz and below the attenuation due to rain is small. For moderate rain (5mm/h) it is only 0.074 dB/km at 10 GHz. The corresponding attenuation at 30 GHz is 0.85 dB/km, while at 100 GHz it is 3.42 dB/km. Since typical line-of-sight paths are 20 to 30 km in length, attenuation rates of 1 dB or more per kilometer can lead to decreases in signal strength. This attenuation must be offset by increased antenna gain or transmitter power, which is relatively expensive if a 1000-fold increase is required [10-15].

![Figure 1.12 Attenuation by rain at 10, 30 and 100 GHz as a function of rain rate](image)

### 1.9 ATTENUATION BY FOG

The attenuation of microwaves and millimeters waves by fog is governed by the same fundamental equations as attenuation by rain [8-10]. The main difference is that fog is suspended mist of very small water droplets with radii in the range 0.01 to 0.05 mm. For frequencies below 300 GHz the attenuation by fog is essentially linearly proportional to the water content per unit volume at any given frequency. The upper level for water content is around 1 g/m$^3$ with the content usually considerably less than this for most fogs. A concentration of 0.032 g/m$^3$ corresponds to a fog that is characterized by an optical
visibility of around 2000 ft. A concentration of 0.32 g/m$^3$ corresponds to an optical visibility range of around 400 ft. The attenuation by fog in decibels per kilometers as a function of frequency is shown in Figure 1.13, for the two concentration levels mentioned above. At a frequency of 300 GHz the attenuation in the more dense fog is still only about 1 dB/km. Hence, for communication link designs with sufficient signal margin built in to overcome the attenuation by rain, the attenuation by fog will not be the limiting factor.

Figure 1.13 Attenuation in fog as a function of frequency for two different concentrations

1.10. ATTENUATION BY SNOW AND ICE

In general, when water solidifies into snow and ice crystals, there is a significant change in the complex dielectric constant $k = k' - jk''$. For ice, $k'$ is nearly constant and equal to 3.17 for temperatures from 0 to 30$^\circ$C, throughout the centimeter and millimeter wave bands [16]. However the imaginary part is very small, nearly independent of frequency in the microwave and millimeters bands, and drops from a value of approximately $3.7 \times 10^{-3}$ at 0$^\circ$C to $5.2 \times 10^{-4}$ at 30$^\circ$C. The small of the imaginary part indicates relatively little attenuation by dry ice crystals. However, snow and hail consists of a mixture of ice crystals and water in many instances, so the attenuation is strongly dependent on the meteorological conditions. Furthermore the shape of snow and ice crystals is so varied that the calculation of absorption by a single typical particle is formidable task, however if needed a typical particle can be defined. Attenuation of microwaves in dry snow is at least
an order of magnitude less than in the rain for the same precipitation rate. However, attenuation by wet snow is comparable to that in rain and may even exceed that of rain at millimeter wavelengths. Even in dry snow, measurements have shown that the attenuation of 0.96 mm radiation may be greater than in the rain with same precipitation rate. Measurements have shown an attenuation of around 2dB/km at 35 GHz for wet snow and a precipitation rate of 5 mm/h. However dry snow the attenuation is comparatively less.

Therefore in the present THESIS the authors have studied the influences of abnormal conditions on the performances of microstrip antenna entitled “Study of Environmental Effects on the Performance Characteristics of Microstrip Antennas”. Accordingly, chapter two dedicated to the historical review of dielectric loading effects on various types of patch antenna with emphasis on theoretical analysis/ techniques and practical design for more than three decades.
REFERENCES